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-: HAND WRITTEN NOTES:-

OF

ELECTRICAL ENGINEERING

1

-: SUBJECT:-

POWER ELECTRONICS

3

TOPICS

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1. Power semiconductor devices - 25%
2. Phase controlled rectifiers - 35%

   Application → DC drives
   → Charging batteries
   → Solar batteries

3. Inverters - 12%
4. Choppers - 12 - 15%
5. AC Voltage controllers & cycloconverters - 3 to 4%
6. Other applications - 7 - 10%

   → AC Drives
   → HVDC
   → SMPS

1-12-11 # POWER SEMICONDUCTOR DEVICES (4)

Power Electronics - deals with control & conversion of high power applications.

Power Semiconductor devices - Should be capable to handle large magnitudes of power.
eg Power diode, SCR(P↑), LASCR, GTO, ASCR, & RCT, TRIAC, DIAC, Power transistors (BJT, MOSFET, IGBT) (f↑)

Signal Electronics - deals with control of low power applications.

Signal Devices - handle low power & very high switching frequencies.
eg Signal diodes → Zener diode
→ LEDs
→ Varacter diode
Signal transistors → BJT
→ MOSFET
→ UJT etc.

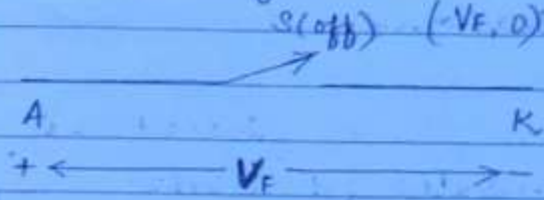
* In the fabrication of semiconductor devices we must sacrifice one quality in order to improve the other quality.
eg If the device operates at very high switching frequency the power rating is reduced.

* A switch can be utilized in 4 different modes but all the devices need not operate in all the 4 modes

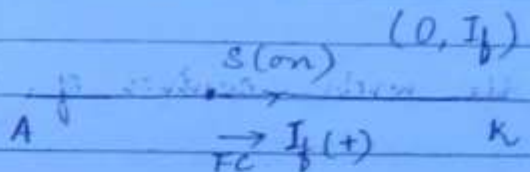
FOUR MODES OF A SWITCH (Ideal)

(3)

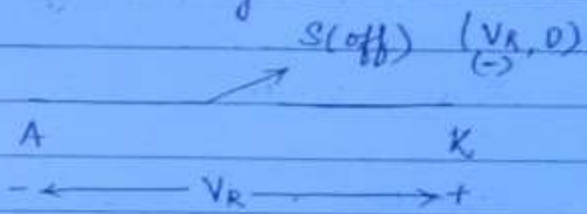
1. Forward Blocking Mode



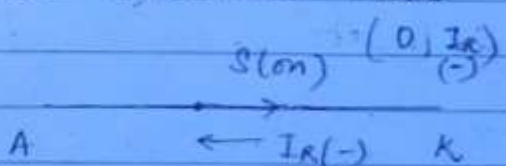
2. Forward Conduction Mode

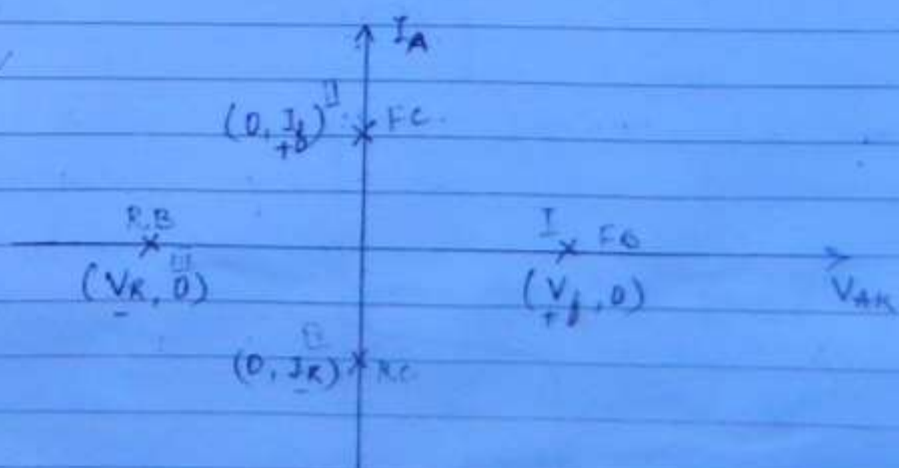


3. Reverse Blocking Mode



4. Reverse Conduction Mode





* TRIAC will support all four modes of operation.
∴ Its treated as an AC switch. (AC → AC)
Applications → AC voltage controllers

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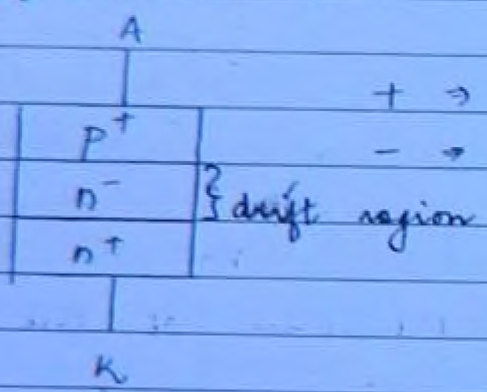
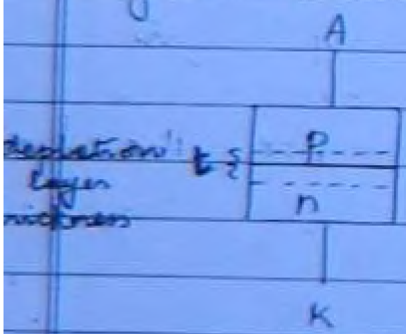
* SCR is a DC switch because it will not support reverse conduction.

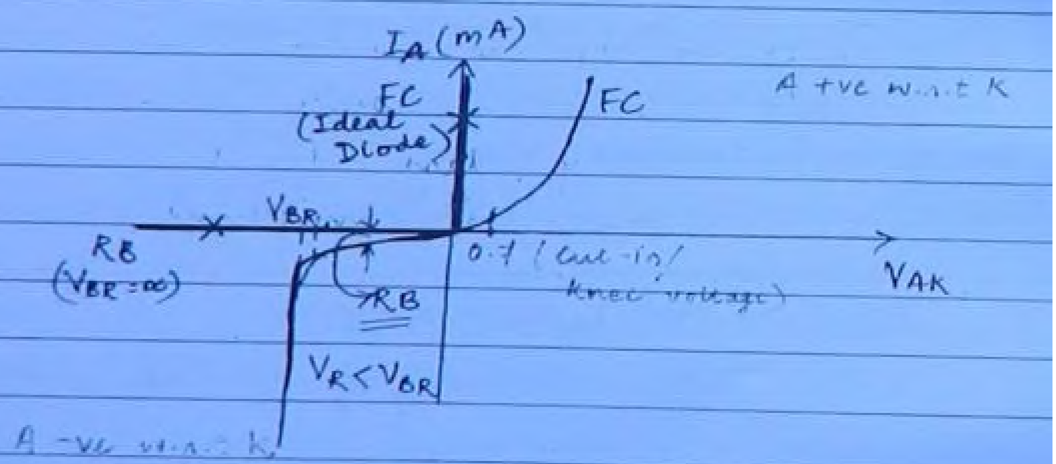
POWER DIODE. –



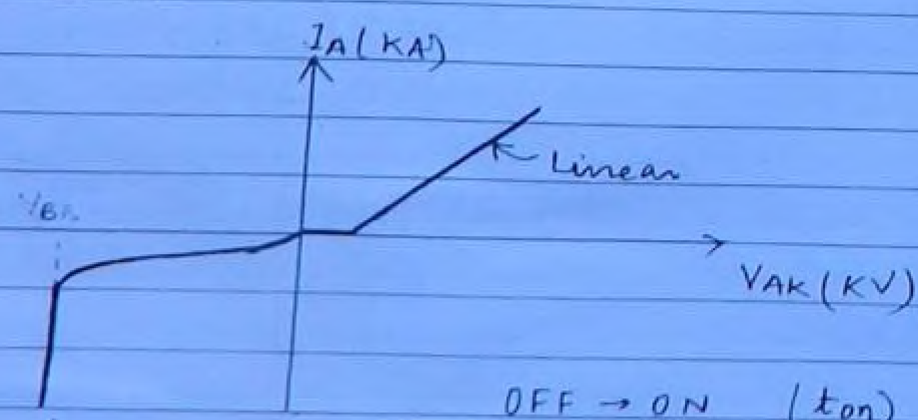
+ → heavily doped layer
– → lightly doped layer

Signal Diode –





Power Diode –



OFF → ON (t_{on})
ON → OFF (t_{off})

Significance of drift region -

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* The thickness of the depletion layer decides the reverse blocking voltage capability.

* The thickness of the depletion layer ↑ due to the n^- region (depletion layer penetrates more deeper into the lightly doped layer to equilise the charge) This ↑ the reverse blocking capability of the diode.

~~input~~ t ↑ ~~input~~ → V_R ↑

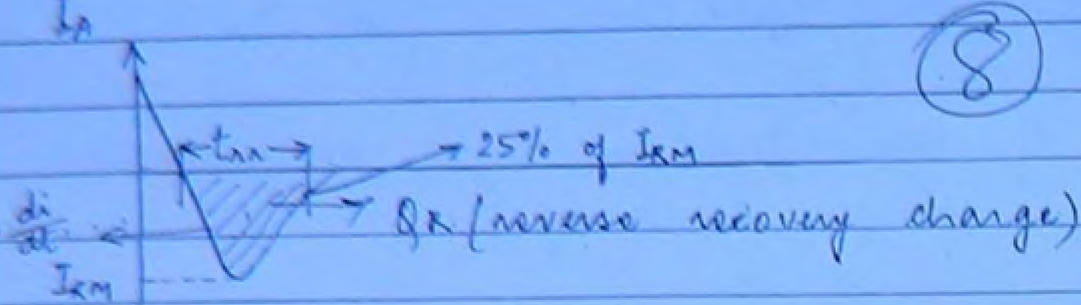
REVERSE RECOVERY CHARACTERISTICS -

* Explains the switching behaviour of the diode from ON time to OFF time.

* When diode is conducting some excess is stored in the device. These excess charge carriers are mainly due to the minority carriers. When diode is switching from ON → OFF the excess charge carriers are still present in the device after anode current becomes 0.

* In order to remove these excess charge carriers and ~~aq~~ acquire equilibrium state, recombination process takes place & hence reverse current flows in the device until all the excess charge carriers are removed ~~from~~ in the device

* This process is known as Reverse Recovery Process & the transition time during this process is known as Reverse Recovery Time (t_{rr})



$t = 0$

ON $\rightarrow$ OFF

$t_{rr} \rightarrow (I_A = 0)$ to $(\downarrow I_A = 25\%\ I_{RM})$

$$I_{RM} = \left[2 Q_R \left(\frac{di}{dt}\right)\right]^{1/2}$$

$$t_{rr} = \left[\frac{2 Q_R}{di/dt}\right]^{1/2}$$

* Q_R depends on I_A.

$$I_A \uparrow \Rightarrow Q_A \uparrow$$

$$\therefore I_{RM} \uparrow \text{ \& thus } t_{rr} \uparrow$$

* The t_{rr} decides the switching frequency of the dio[de]

$$t_{rr} \uparrow \Rightarrow f_s \downarrow$$

Classification of Power Diodes based on Reverse Recovery Time (t_{rr})

1. General Purpose Diode
2. Fast Recovery Diode
3. Schottky Diode

(Slow) ← (high speed) →

	General Purpose Diodes	Fast Recovery Diode	Schottky Diode
1.	$t_{rr} \rightarrow 25\,\mu s$	$t_{rr} \rightarrow 5\,\mu s$ (Less)	$t_{rr} \rightarrow$ nano secs
2.	$I_{rating} \rightarrow$ 1A to several 1000s of A	$I_{rating} \rightarrow$ 1A to several 100 of A	$I_{rating} \rightarrow$ limited to 300 A
*	$V_{rating} \rightarrow$ 50 V to 5 kV	$V_{rating} \rightarrow$ 50 V to 3 kV	V_{rating} = 100 V

* In fast recovery diodes, the layers are doped with gold / platinum.
* Gold / platinum doping reduces the lifetime of charge carriers & increases the speed of recombination. This reduces the reverse recovery time.
* Used in choppers & Inverters
* Schottky diode is a metal to semiconductor junction diode. Here the conduction is only due to majority carriers.
* Since there is no minority charge carriers the t_{rr} delay is very much reduced. ∴ it operates at very high switching frequency.
* Due to the absence of drift region, the thickness of depletion layer is reduced ∴ it can block a small reverse voltage limited to 100 V.
* Can be used in low power high switching frequency applications

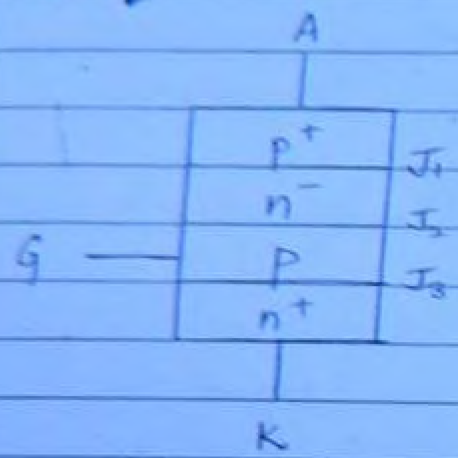
  eg Switch Mode Power Supply (SMPS)

* Used in uncontrolled rectifiers, free wheeling diodes for rectifiers.

* Diode is an uncontrolled device because there is no control terminal to decide its on off state.

(10)

SCR -



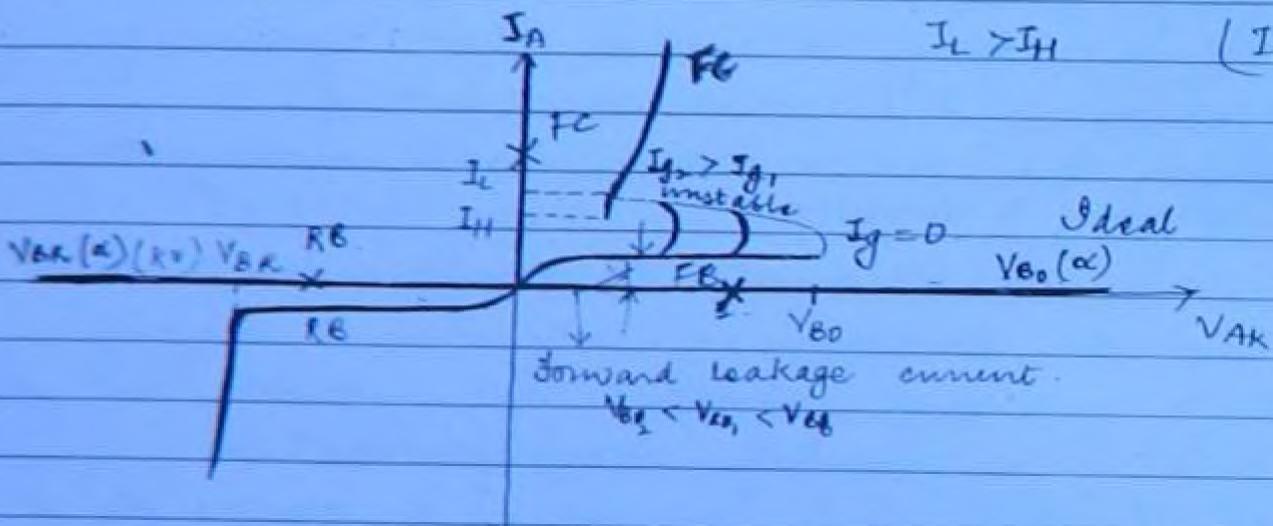
A, K ⇒ main terminals

G ⇒ control terminals

(or) there is no control of gate when SCR is ON.

Partially controlled device

$I_L > I_H$ $\quad (I_L = 2.4\, I_H)$



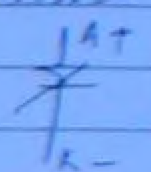
Forward Blocking Mode - A +ve w.r.t K

$J_1, J_3 = FB$

$J_2 = RB$.

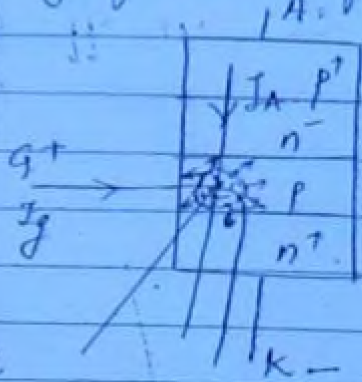
$\therefore$ SCR → OFF

Forward Conduction Mode



$V_{AK} \uparrow \Rightarrow V_{BO}$ (forward breakdown voltage) then breakdown occurs at J_2 $\therefore$ SCR → ON

Significance of gate signal -



When gate signal is applied, charge gets accumulated in depletion region p, $I_A \uparrow$ and $\uparrow$ the charge accumulation as it gets a conduction path. This leads to $\uparrow$ of charge & thus breakdown of depletion region turning on the SCR.

(initial conduction Area)

If $I_g \uparrow$ or $\frac{dI_g}{dt} \uparrow$ Initial conduction Area $\uparrow$

$\Rightarrow \frac{dI_A}{dt} \uparrow$ and lesser V_g is need for breakdown ie less V_{BO}

Reverse Blocking Mode -

A ~~must~~ -ve w.r.t K

$J_2 \Rightarrow FB$

$J_1, J_3 \Rightarrow RB$

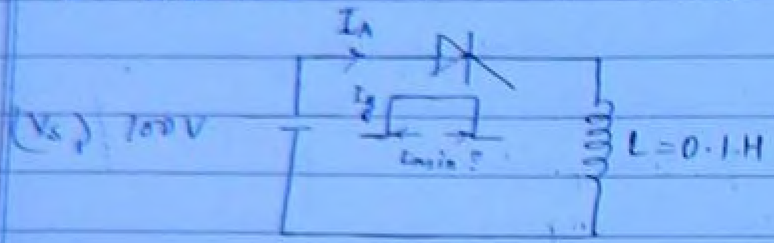
$SCR \rightarrow OFF$

Significance of Latching Current -

- Latching current is related to turn on process
- When SCR is in the ON state, gate signal is removed to avoid the continuous gate power loss
- If we remove the gate signal when $I_A < I_L$ then SCR fails to turn ON ie We must maintain the gate pulse width until I_A reaches just above certain minimum value (Latching current)
- When we remove gate signal, when $I_A > I_L$ then

SCR ~~continuous~~ SCR continues to be the ON state.

Q What is the minimum gate pulse width required to turn on the SCR in the following circuit.
$I_L = 100\ mA$



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Sol
$$V_s = L\frac{di_A}{dt}$$

$$\int di_A = \int \frac{V_s}{L}\, dt$$

$$I_A = \frac{V_s}{L}\, t \Rightarrow t_{min} = \frac{I_A \times L}{V_s}$$

$$t_{min} = \frac{I_L \times L}{V_s} = \frac{100 \times 10^{-3} \times 0.1}{100}$$

$$t_{min} = 100\ \mu sec.$$

** The minimum gate pulse width requirement to turn on the SCR depends on the load parameters

eg. $L \uparrow \Rightarrow t_{min} \uparrow$

Q If Load is $R = 20\ \Omega$ $L = 0.1\ H$ in ~~pa~~ prev ques.

$$V_s = L\frac{di_A}{dt} + R I_A$$

$$I_A = \frac{V_s}{R}\left(1 - e^{-t/T}\right) \qquad T = \frac{L}{R} = \frac{0.1}{20} = \frac{1}{200}$$

$$I_A = 5\left(1 - e^{-200t}\right)$$

$$I_L = 5\left(1 - e^{-200 t_{min}}\right)$$

$$t_{min} = 101\ \mu s$$

Significance of Holding current -

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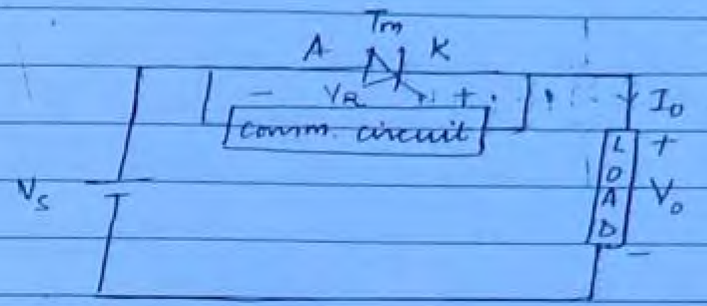
Holding current is related to turn OFF process. Gate has no control to turn OFF SCR. In some cases we require commutation circuit to turn OFF SCR.



Comm ckt forces anode current to reduce below I_H. After that it applies reverse voltage to remove all charge carriers in the device.

* Procedure to turn OFF SCR using a commutation ckt -

Commutation circuit forces anode current to reduce below a certain minimum value & then applies a reverse voltage across the SCR atleast [illegible] for a period of device turn off time or greater than that.

Circuit turn off time t_c
It is the time for which the comm[illegible] circuit applies a reverse voltage acros[illegible] after the anode current becomes 0. [illegible]

Device turn off time
It is the time taken to remove [illegible] charge carriers present in the devic[illegible] provides the tq.

will turn on before applying gate (behaves as diode)

Page

$\boxed{\text{For successful commutation } t_c > t_q \text{ always.}}$

If $t_c < t_q$ commutation fails

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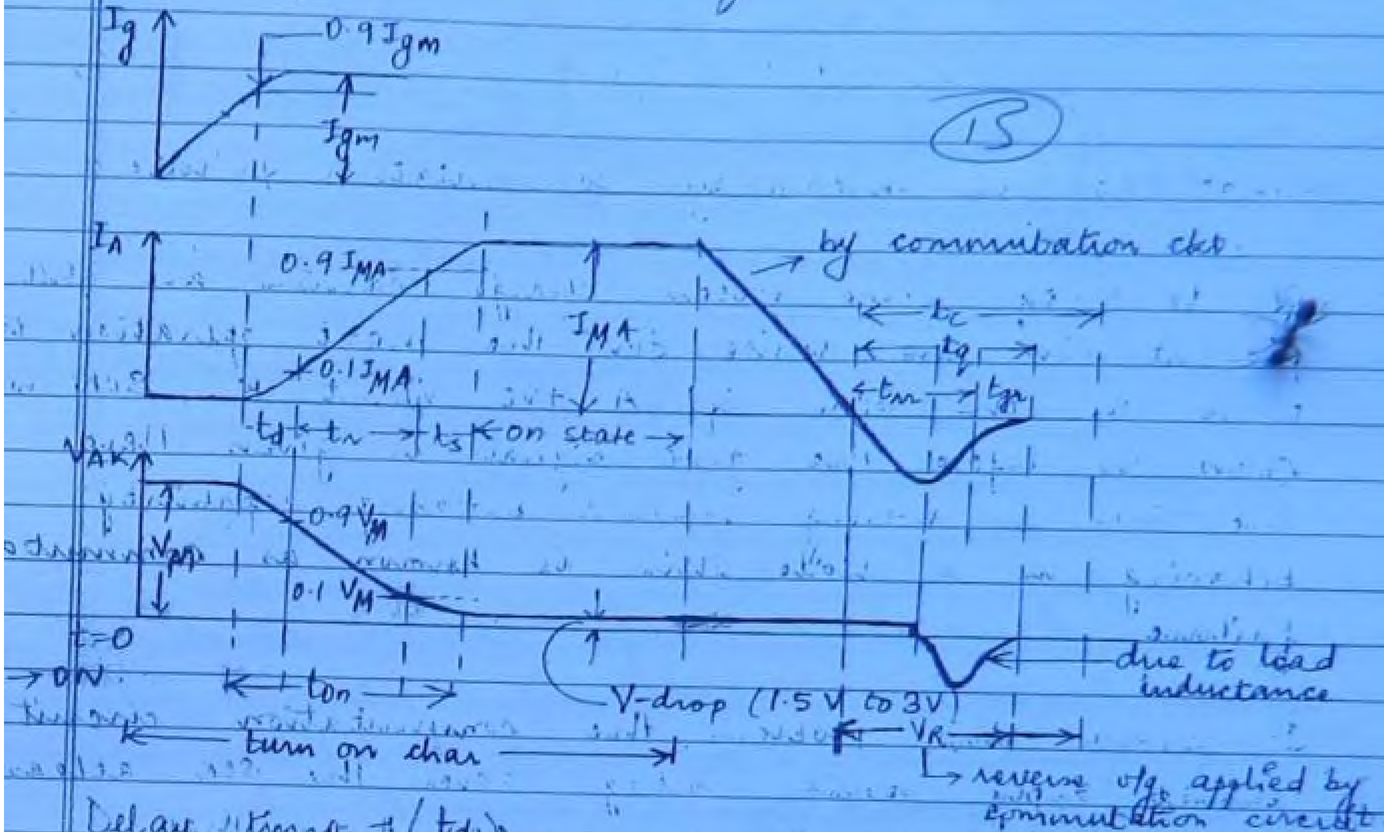
Q What do you mean by commutation failure?

A If $t_c < t_q$ some excess charge carriers are still present in the device. For the next operation to turn on the SCR if A +ve w.r.t K, SCR will turn on before the gate signal is given. Here the SCR is loosing forward blocking capability behaving as a diode. This is known as commutation failure.

To avoid this problem, the commutation circuit should apply reverse voltage across the SCR atleast for a period of t_q or greater than that

Holding current is the minimum I_A below which the SCR becomes off and regains the forward blocking capability if a reverse voltage is applied across SCR atleast for a period of t_q or more than that

Switching Characteristics of SCR –



Delay time (t_d)
depends on gate signal magnitude & d/dt of gate signal magnitude.

Delay time ⇒ $0.9\ I_{gm}$ to $0.1\ I_{MA}$.
$0.9\ I_{gm}$ to $0.9 V_m$

Delay time depends on ⇒ $I_g \uparrow$ & $\frac{dI_g}{dt} \uparrow$ ⇒ $t_d \to$ ON
∴ $t_{ON} \downarrow$

⇒ (initial conduction area) ↑
$\frac{di_A}{dt}$ ↑ initial state
$t_d \downarrow$

Rise time (t_r) 0.1 I_{MA} to 0.9 I_{MA}

0.9 V_M to 0.1 V_M

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Rise time depends on Load parameters

eg $L \uparrow$ $\frac{di_A}{dt} \downarrow$ $\therefore$ $t_r \uparrow$, $t_{on} \uparrow$

Spread time (t_s) 0.9 I_{MA} to I_{MA}.

0.1 V_M to (ON state V-drop)

(1.5 - 3 V)

Reverse recovery time (t_{rr})

During t_{rr} the excess charge carriers present in the outer layers is reduced.

Gate Recovery time (t_{gr})

During t_{gr} the excess charge carriers present in the inner layers near the gate junction is removed.

Device turn off time (t_q)

$t_q = t_{rr} + t_{gr}$

The device turn off time is generally very much greater than turn on time. Therefore the device turn off time decides the switching characteristics of the SCR.

t_q → Slow thyristors (converter grade thyristors)

$t_q \rightarrow 50\mu s$ to $200\mu s$

→ Fast thyristors (inverter grade thyristors)

$t_q \rightarrow 3\mu s$ to $50\mu s$

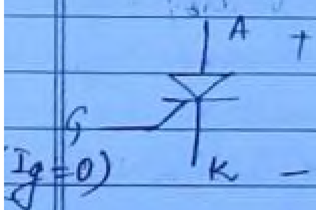
For successful commutation $t_c > t_q$

$$\boxed{t_c = SF\ t_q}$$

$SF > 1$ for successful comm.
(Safety factor)

TURN-ON / TRIGGERING METHODS OF SCR -

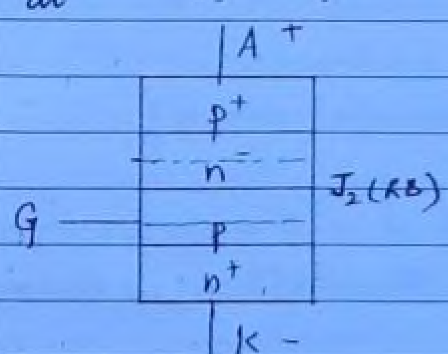
1. Forward voltage triggering -

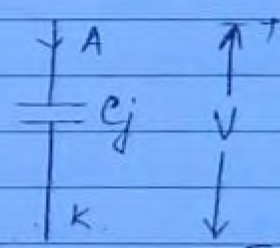


$V_{AK} \uparrow \Rightarrow V_{BO}$, Breakdown occurs at J_2 & SCR $\to$ ON.

This method is generally not preferred because the SCR may get destroyed due to high power loss when triggered at high voltage.

2. $\frac{dv}{dt}$ triggering -





$$\uparrow I_c = C_j \frac{dV}{dt} \uparrow$$

At high dv/dt the charging current increases. If the increase in charging current is more than the latching current then SCR is turned on

Critical dv/dt → It's the dv/dt at which SCR will turn on. At critical dv/dt charging current is equal to latching current ie $I_c = I_L$.

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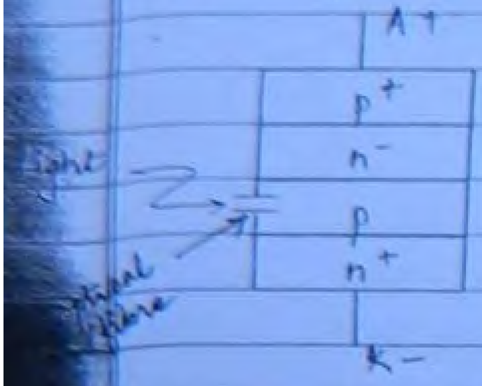
3. Light triggering -



When light radiation is incident near the depletion layer then more number of e^- hole pairs are produced by absorbing the light energy in the depletion layer, & this initiates the turn on process.

Application → Used in LASCRs for HVDC applications.

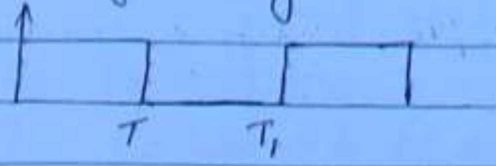
Thermal triggering -
When temperature is increased near the reverse biased gate junction, then the device is initiated to turn on by e^- - hole pairs produced in depletion layer absorbing the thermal energy.

Semiconductors are very sensitive & the characters of the device may change as the temperature changes, so this method is not used.

4. Gate triggering -

(a) Continuous Gate Triggering -
Continuous gate signal is applied until the SCR is expected to be in the ON state. This is not efficient triggering due to continuous gate power loss.

(b) Pulse gate signal -



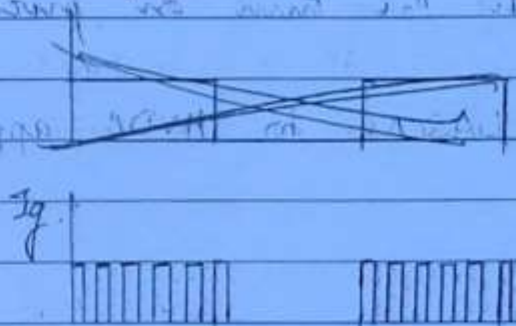
T = gate pulse width

$T > t_{min}$

$T_1 \rightarrow$ time period

$\delta = \frac{T}{T_1}$ = duty ~~d~~ cycle.

(c) High frequency gate signal



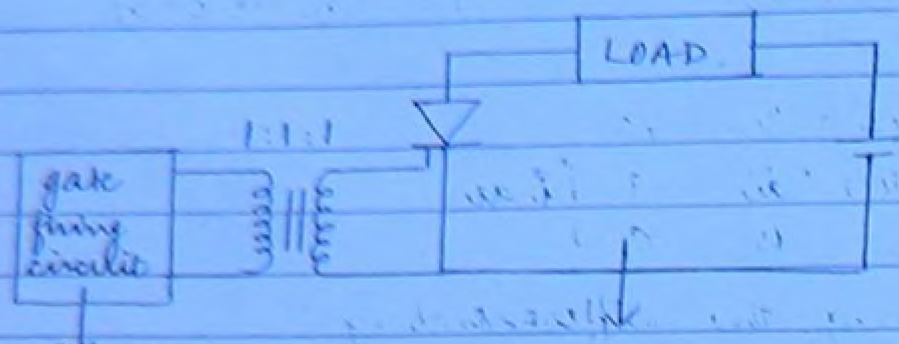
Adv.

→ reduce the size of pulse transformer

↓

provides electrical isolation b/w high power main circuit & low power gate firing circuit.

→ we can trigger more than one SCR using a pulse transformer.



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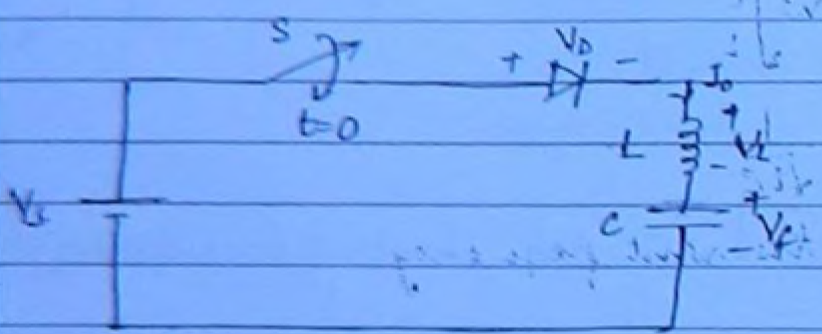
Q1



$V_c(t=0) = 0\,V$

i) When switch is closed at $t=0$ then diode conducts f

a) $\frac{\pi}{2}\sqrt{LC}$

b) $\pi\sqrt{LC}$

c) $\pi/3\sqrt{LC}$

d) $2\pi\sqrt{LC}$

ii) $V_c = ?$ when diode stops conducting

a) V_s

b) $2V_s$

c) $-V_s$

d) $-2V_s$

Sol. i) $S \to ON$ $(t=0)$ ~~Fo~~ $D = FB$ (on at $t=0$)

$$V_s = V_D + V_L + V_C$$

$$V_s = 0 + L\frac{di}{dt} + \frac{1}{C}\int i\,dt$$ 21

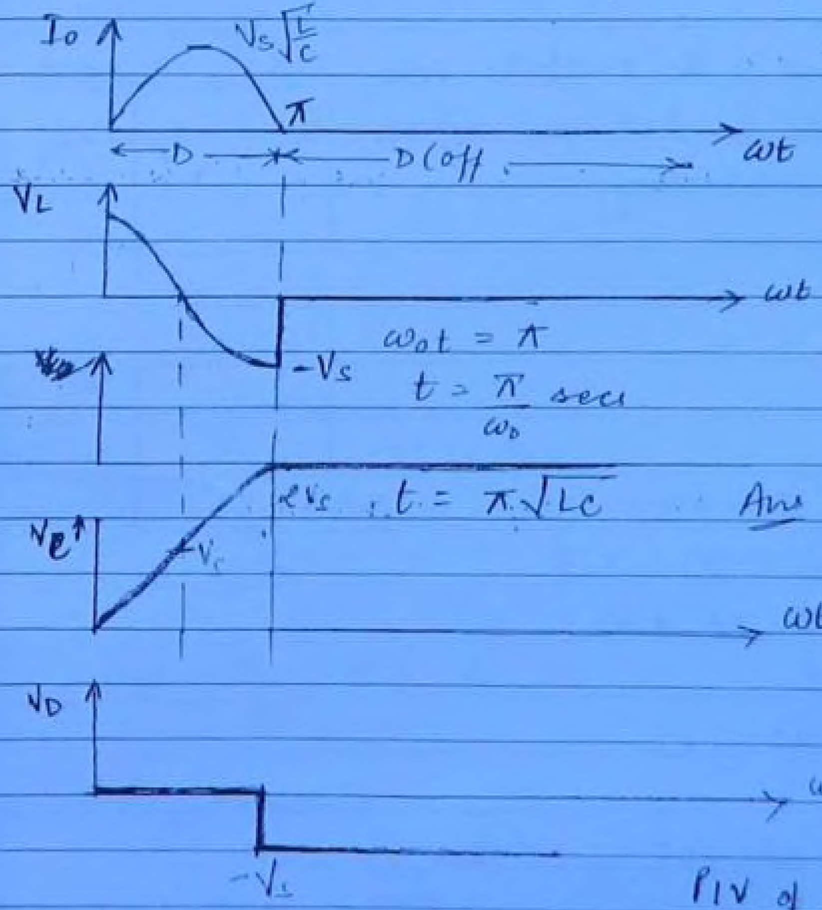
on solving the differential eq.

$$I_o = V_s\sqrt{\frac{C}{L}}\sin\omega_o t$$

$$\boxed{I_o = I_p \sin\omega_o t}$$

where $I_p = V_s\sqrt{\frac{C}{L}}$

$\omega_o = \frac{1}{\sqrt{LC}}$ → Resonant frequency



$\omega_o t = \pi$

$t = \frac{\pi}{\omega_o}$ sec

$t = \pi\sqrt{LC}$ Ans (b)

PIV of diode $= V_s$

$$V_L = L\frac{dI}{dt}$$

$$= L\frac{d}{dt}(I_p \sin \omega_0 t)$$

$$\boxed{V_L = V_s \cos \omega_0 t}$$

$$V_C = V_s - V_L$$

$$= V_s - V_s \cos \omega_0 t$$

$$\boxed{V_C = V_s(1 - \cos \omega_0 t)}$$

$$-V_s + V_D + 0 + 2V_s = 0$$

$$\therefore \boxed{V_D = -V_s}$$

(22)

from 0 to 90°

Source $\rightarrow \frac{1}{2}LI^2 + \frac{1}{2}CV^2$

from 90 to 180°

$\frac{1}{2}LI^2 \rightarrow \frac{1}{2}CV^2$

Q2 In prev. ques assume $V_C(t=0) = V_0$ volts where $V_0 < V_s$.

i) when switch is closed at $t=0$ secs whats the cap. v/g (V_C) after the diode stops conducting

a) $2(V_s + V_0)$

b) $2(V_s - V_0)$

c) $2V_s + V_0$

d) $2V_s - V_0$

$S \rightarrow ON$ $(t=0)$ $D = FB$

$$V_s = 0 + L\frac{di}{dt} + \frac{1}{C}\int i\,dt + V_0$$

$$V_s - V_0 = L\frac{di}{dt} + \frac{1}{C}\int i\,dt$$

$$I_0 = (V_s - V_0)\sqrt{\frac{C}{L}} \sin \omega_0 t \qquad \omega_0 = \frac{1}{\sqrt{LC}}$$

$$I_0 = I_p\sqrt{\frac{C}{L}} \sin\omega_0 t$$

$$V_L = L\frac{d}{dt}(I_p \sin \omega_0 t)$$

$$V_L = (V_s - V_0)\cos \omega_0 t$$

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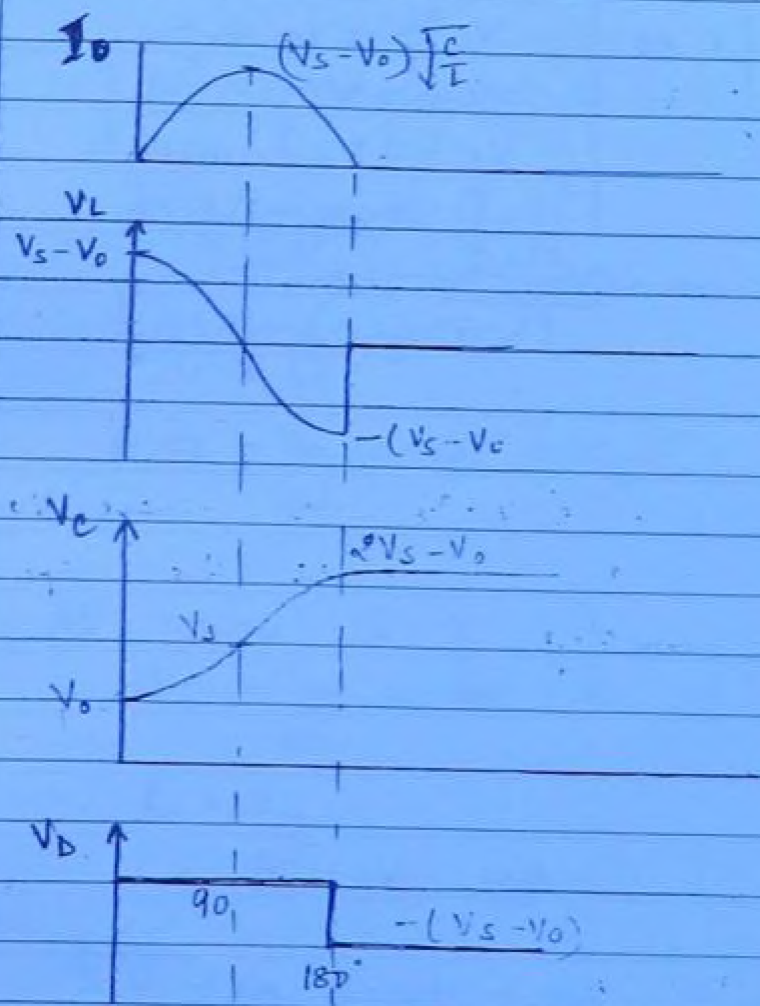
$$V_c = V_s - V_L$$

$$= V_s - (V_s - V_0)\cos \omega_0 t$$

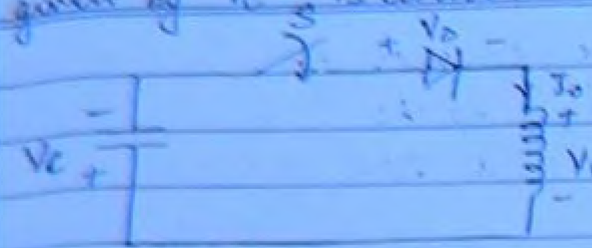
$$V_c = V_s(1 - \cos \omega_0 t) + V_0 \cos \omega_0 t$$



$$PIV = V_s - V_0$$

Ans $V_c = 2V_s - V_0$

then charging current of cap behaves as an sine function given by $I_p = I_p \sin \omega_0 t$ & charging vlg of cap behaves as ~~positive~~ function given by $V_c = V_s \cos \omega_0 t$.

Q3.


V_c when D stops conducting
a) $2V_s$
b) $-2V_s$
c) $-V_s$
d) V_s

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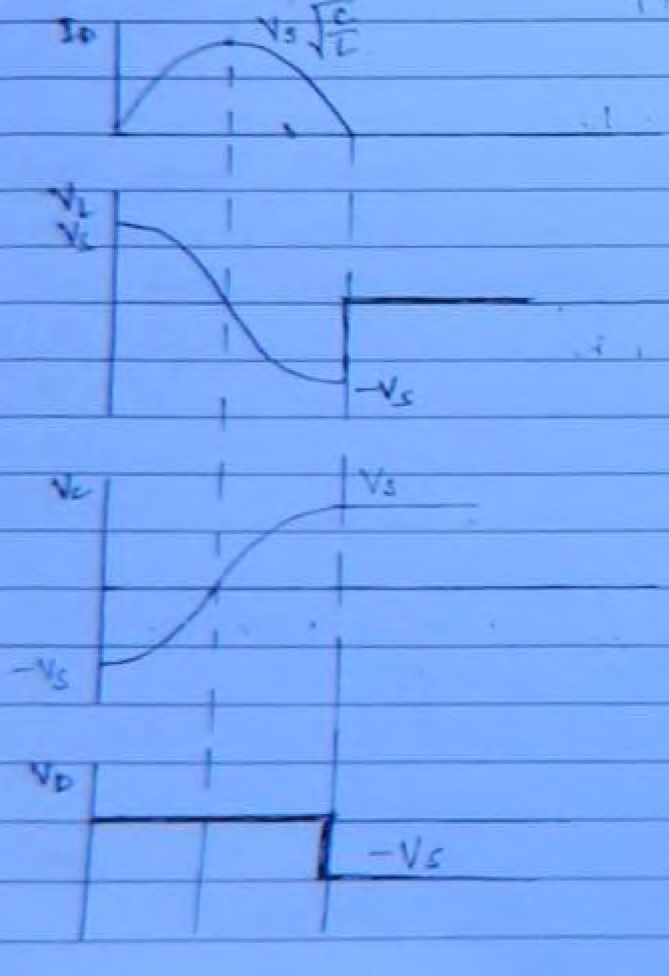
$V_c(t=0) = -V_s$

Sol
$V_c + V_D + V_L = 0$
$V_L + (V_c - V_s) = 0$
$V_s = V_L + V_c$
$V_s = L\dfrac{di}{dt} + \dfrac{1}{C}\int i\,dt$
$I_o = \underbrace{V_s\sqrt{\dfrac{C}{L}}}_{I_p} \sin \omega_0 t$

$V_L = L\dfrac{d}{dt}(I_p \sin \omega_0 t)$
$V_L = V_s \cos \omega_0 t$
$V_C = -V_L$
$V_c = -V_s \cos \omega_0 t$

~~Ans same as Q1 so $2V_s - V_c$ (a)~~



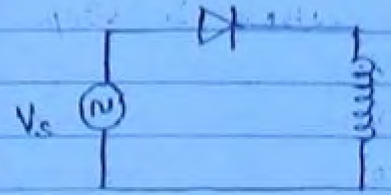
Ans $V_c = V_s$

PIV = V_s

If $V_c(t=0) = V_s$ then cap vlg reverses & $V_c = -V_s$

GATE Q4

half bridge rectifier



Diode conducts for

a) 90° b) 180°

b) 270° d) 360°

25

sol.

$$V_s = L\frac{di}{dt}$$

$$V_m \sin \omega t = L\frac{di}{dt}$$

$$\int di = \int \frac{V_m \sin\omega t}{L}\, dt$$

$$\cancel{I = \frac{V_m}{\omega L}\cos\omega t \Big|_0^t}$$

$$\cancel{I = \frac{V_m}{\omega L}(\cos\omega t - 1)}$$

$$I = -\frac{V_m}{\omega L}\cos\omega t + K$$

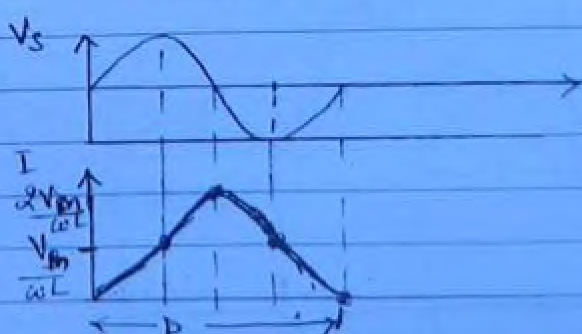
At $\omega t = 0$ $I = 0$

$$0 = -\frac{V_m}{\omega L}\cos 0 + K$$

$$K = \frac{V_m}{\omega L}$$

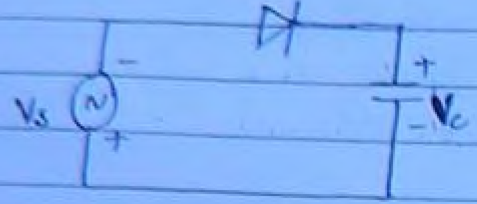
$$I = -\frac{V_m}{\omega L}\cos\omega t + \frac{V_m}{\omega L} = \frac{V_m}{\omega L}(1 - \cos\omega t)$$

(*) In 180° – 360° cycle the inductance energy forces the D to maintain the conduction until it releases the complete energy



Ans 360°

Q5.



Diode conducts for

a) 90° b) 180°

c) 270° d) 360°

26

sol

$$V_c + V_s = 0$$

$$V_s = -V_c$$

$$V_m \sin \omega t = -\frac{1}{C}\int i \, dt$$

$$I = -V_m C \times \omega \times \cos \omega t$$

$$I = -V_m \omega C \cos \omega t$$

COMMUTATION TECHNIQUES

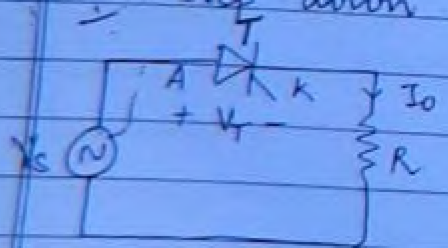
1. Natural / Line Commutation –

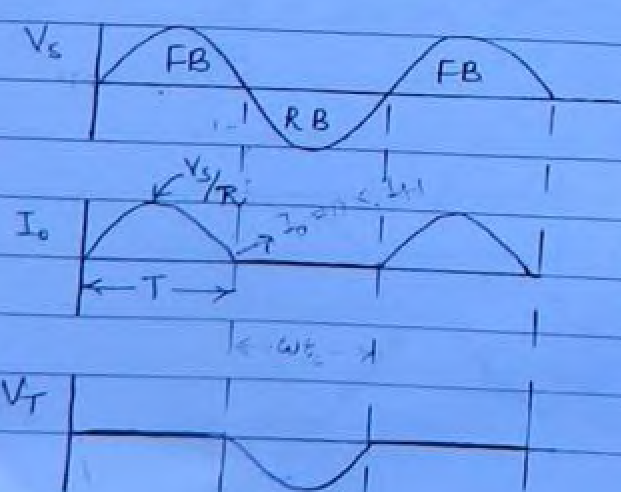
If nature of supply supports commutation process its called natural commutation.

eg Rectifiers

AC vlg controllers

Step down cyclo converters





$$\omega t_c = \pi$$

$$\boxed{t_c = \frac{\pi}{\omega}} \text{ secs}$$

$T \rightarrow ON$

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

$$I_o = \frac{V_m \sin \omega t}{R}$$

2 Forced commutation :-

DC supply will not support the commutation process. We need a separate forced commutation circuit to turn off the SCR.

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eg choppers

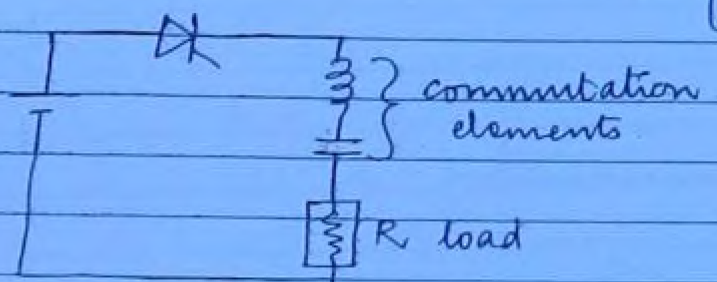
inverter

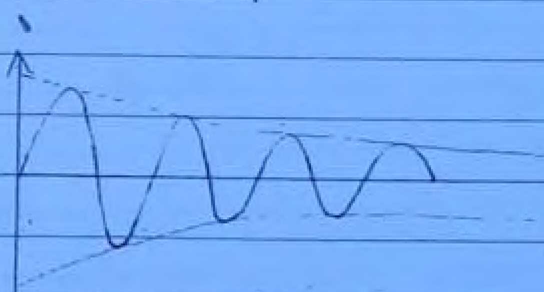
step up cycloconverter

a) Class A commutation circuits -

RLC should satisfy underdamped condition

$$\left(R^2 < \frac{4L}{C}\right)$$





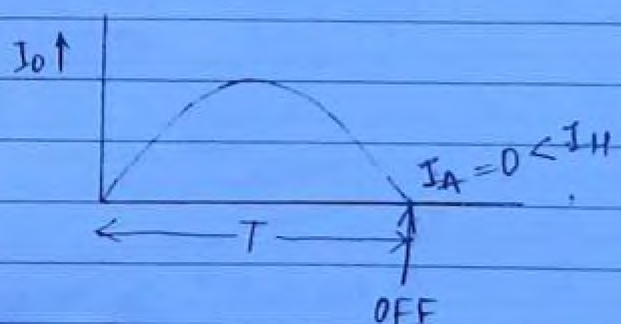
$$I = \frac{V_s}{\omega_r L} e^{-\delta t} \sin \omega_r t$$

$$\delta = \frac{R}{2L}$$

$$\omega_r = \sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$$

$\downarrow$

singing frequency



$$\omega_r t = \pi$$

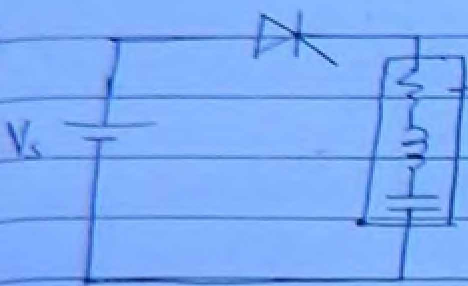
$$t = \frac{\pi}{\omega_r}$$

conduction time of thyristor $= \frac{\pi}{\omega_r}$ sec.

Load commutation -

If the load elements support the commutation process then it is known as load commutation.

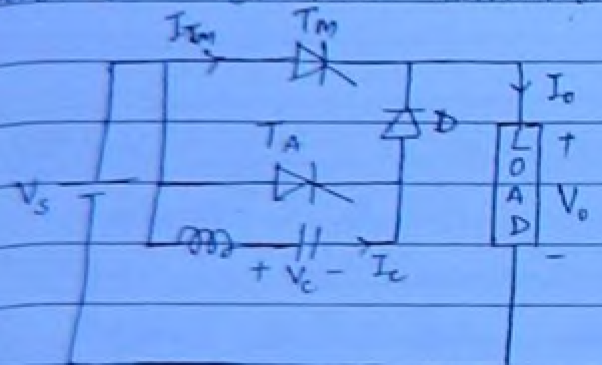
eg If load is RLC load satisfying underdamped condition as shown in figure.



→ RLC load satisfying underdamped condition

$$R^2 < \frac{4L}{C}$$

Class B - Current Commutation -



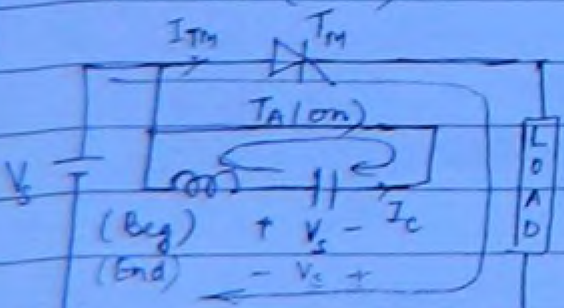
Assume → $V_C(t=0) = V_s$

→ I_o = constant
(highly inductive load)

→ T_M → ON $(t<0)$

① Mode

$T_A \rightarrow ON\ (t=0)$



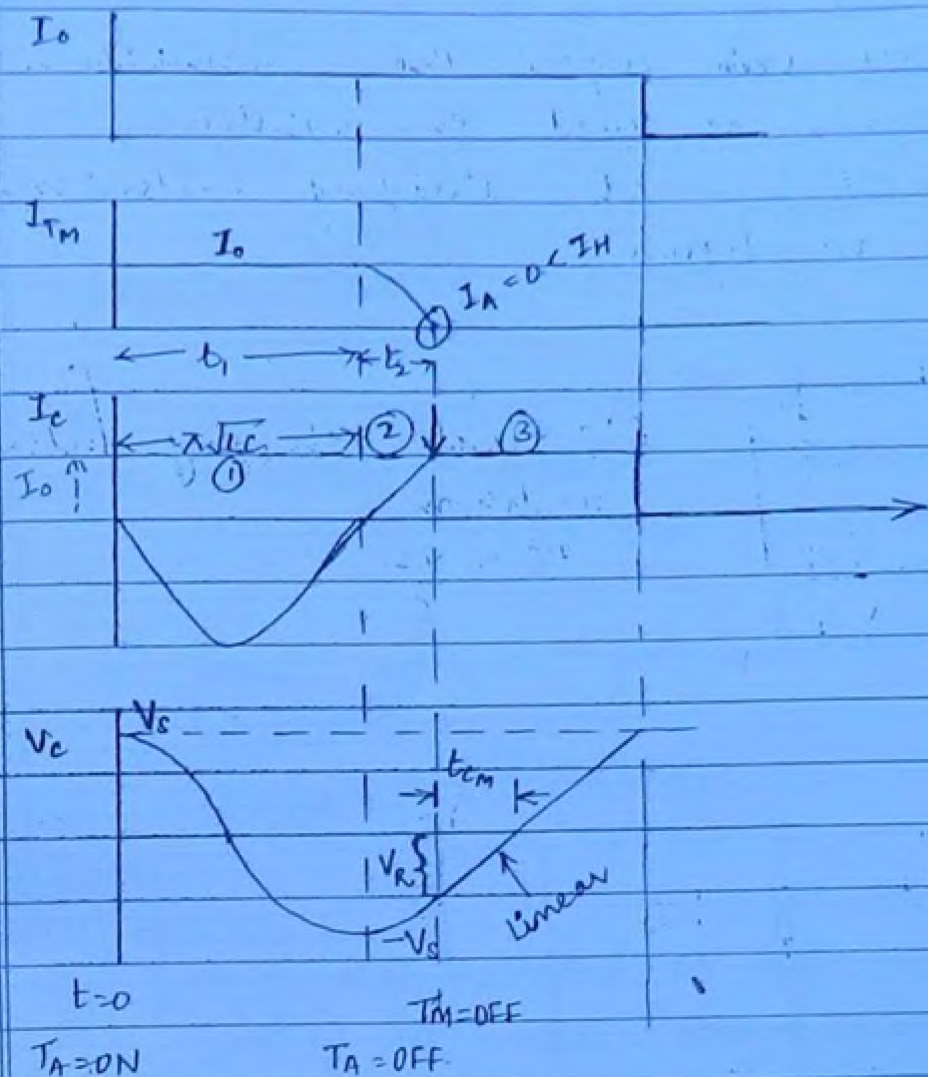
$I_{T_M} = I_o$

$I_c = -I_p \sin\omega_0 t$ [∵ dir^n is opp to reference I_c]

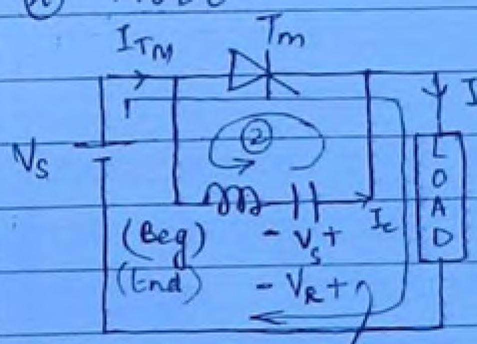
$V_C = V_S \cos\omega_0 t$

End → $V_c = -V_s$

$I_c = 0$



(2) MODE



for this also reference initial is the vlg V_c

$$\downarrow I_{TM} = I_0 - I_c \uparrow$$

END $\rightarrow$ when $I_c = I_0$ $\quad I_{TM} = 0$

$\therefore T_M \rightarrow$ OFF

$$I_p \sin \omega_0 t_2 = I_0$$

$$\omega_0 t_2 = \sin^{-1} \frac{I_0}{I_p}$$

$$\boxed{t_2 = \sqrt{LC} \sin^{-1} \frac{I_0}{I_p}}$$

V_c at the end of mode ②

$$V_c = V_s \cos(\pi + \omega_0 t_2)$$

$$V_c = -V_s \cos(\omega_0 t_2)$$

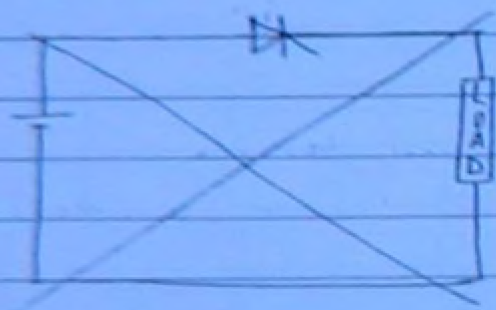
$$V_c = -V_s \cos\left[\sin^{-1}\frac{I_o}{I_p}\right]$$

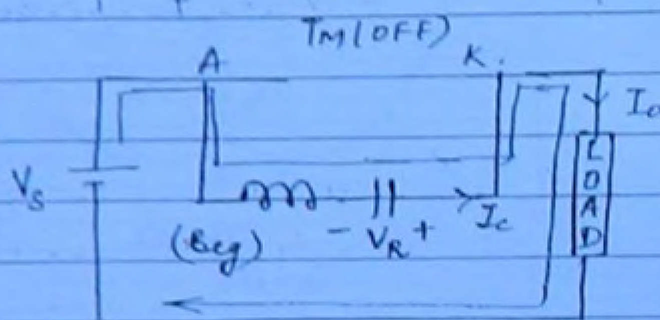
At the end of mode ②

V_R is the reverse voltage of cap.

$$\underline{V_R} = V_s \cos\left[\sin^{-1}\frac{I_o}{I_p}\right]$$

③ MODE





$$I_c = I_o$$

$$V_c = \frac{1}{C}\int i\,dt$$

$$V_c = \frac{I_o}{C} t$$

$$V_R = \frac{I_o}{C} t_{cq}$$

$$\boxed{t_{cq} = \frac{V_R C}{I_o}}$$

circuit turn off time for main thyristor

- $(I_{T_M})_{peak} = I_0$

- $(I_{T_A})_{peak} = V_s\sqrt{\frac{C}{L}}$ (I_p)

- conduction time of $T_A = \pi\sqrt{LC}$

- Min^m time required to turn OFF the main Thyristor after auxiliary thyristor is switched ~~on~~ ON

  $t = \pi\sqrt{LC}$ (for low values of load current I_0)

- Max^m time required to turn OFF T_M

  $t = t_1 + t_2$

  $= \pi\sqrt{LC} + \sqrt{LC}\,\sin^{-1}\left(\frac{I_0}{I_p}\right)$

- If $I_0 > I_p$, commutation is not possible

  $\therefore$ $I_0 \leq I_p$ to make commutation possible

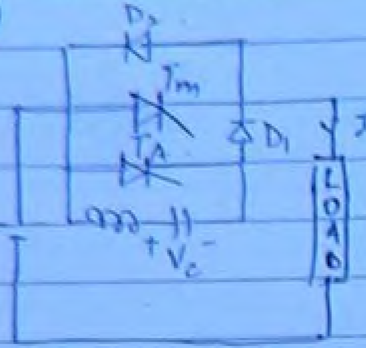
- $t_{C_M} = \frac{CV_R}{I_0}$

- Max reverse vlg applied across the T_M when its in off state is V_R.

  $V_R = V_s \cos\left[\sin^{-1}\frac{I_0}{I_p}\right]$

- If diode is not present, T_A has ~~there is~~ no control on commutation.

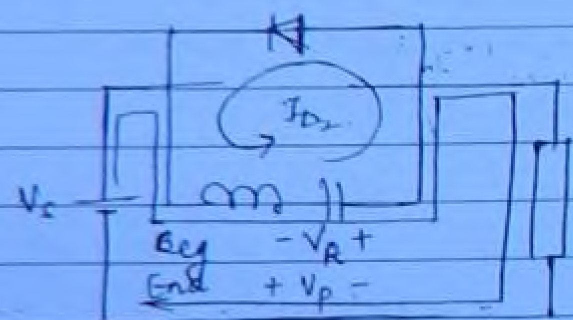
Q. Check if commutation is possible. If possible, calculate the circuit turn-off time of T_M



(32)

A. End of mode 1 reverses polarity of $+V_c-$ to $-V_c+$
In mode 2, T_M is ON, reverses biases D_2
D_1 is FB and D_2 is RB so commutation is possible
D_2 is RB, the ckt becomes same as class B.
1st two modes are same as prev case.

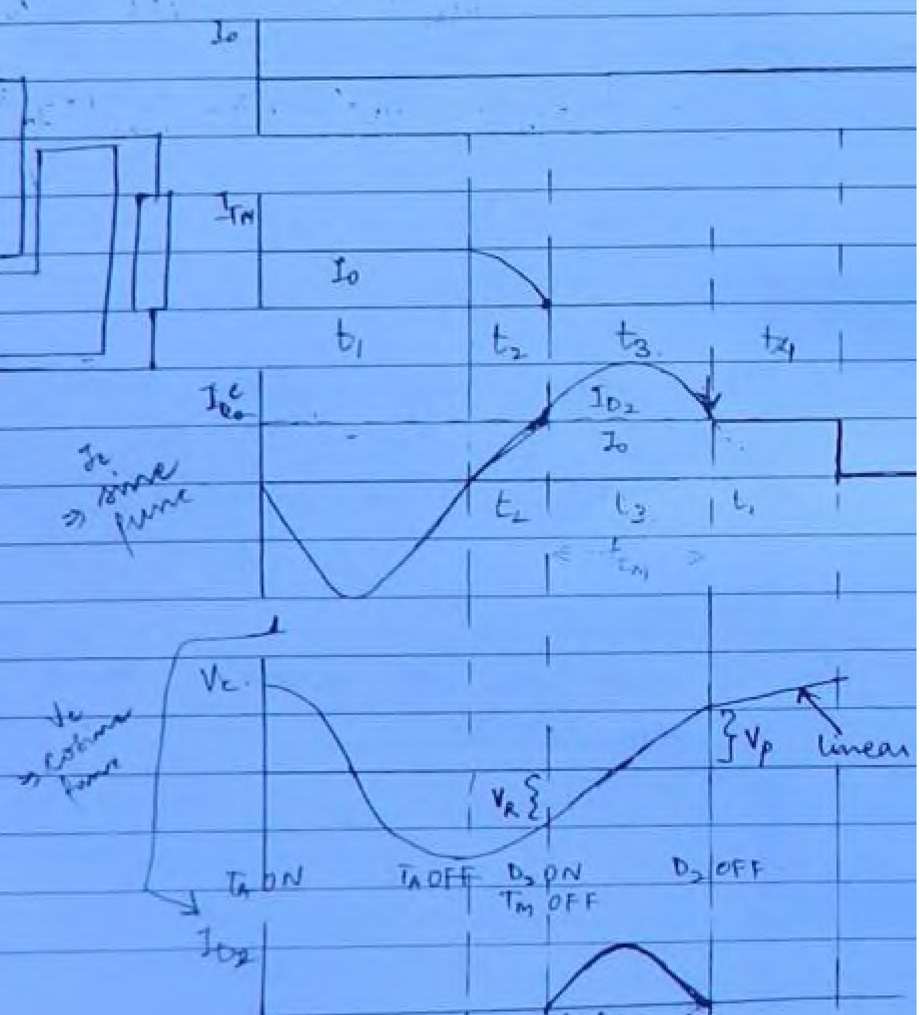
Mode (3)



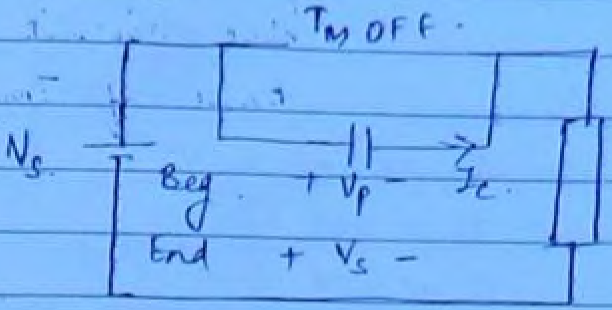
$$I_c = I_0 + I_{D_2}$$

$$\downarrow I_{D_2} = \downarrow I_c - I_0$$

END → when $I_c = I_0$
$I_{D_2} = 0$
D_2 = OFF



Mode (4)



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$I_c = I_o$

In mode (3): when D_2 is in the ON state v/g drop of D_2 applies ~~Rev~~ Reverse v/g across the main thyristor ∴ conduction time of D_2 is equal to ckt turn off time of ~~m~~ T_M

$$t_{cm} = t_3 = \pi\sqrt{LC} - 2t_2$$

$$t_{cm} = \pi\sqrt{LC} - 2\sqrt{LC}\sin^{-1}\left(\frac{I_o}{I_p}\right)$$

Applications -

This type of commutation technique is used in step down chopper ∴ its also known as current commutation chopper.

(c) Class C - Complimentary Commutation

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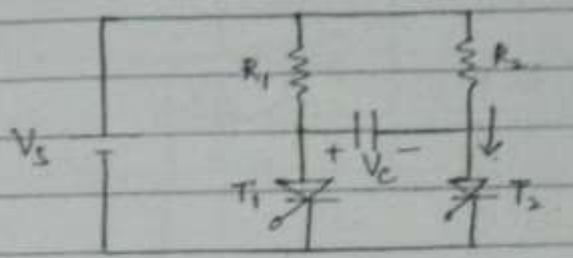
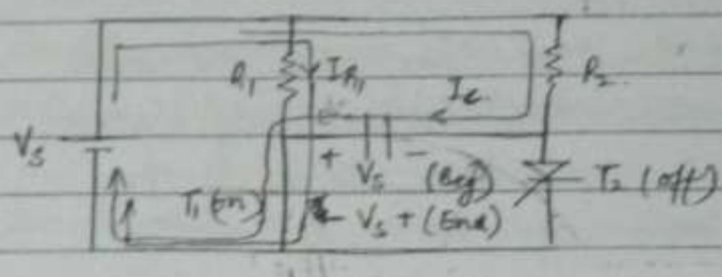


Assume
→ $V_c(t=0) = V_s$
→ $T_2 \to ON$ } $(t<0)$
$T_1 \to OFF$

Mode (1)

At $t=0$, $T_1 \to ON$



$$I_{T_1} = I_{R_1} + I_c$$

$$= \frac{V_s}{R_1} + \underset{\uparrow}{K} e^{-t/R_2C}$$

initial current

$$I_{T_1} = \underbrace{\frac{V_s}{R_1}}_{\text{steady state current}} + \underbrace{\frac{2V_s}{R_2} e^{-t/R_2C}}_{\text{transient current}}$$

$$V_c = \frac{1}{C}\int i\,dt = \frac{1}{C}\int 2\frac{V_s}{R_2} e^{-t/R_2C}$$

$$= 2V_s e^{-t/R_2C} - V_s$$ by trial & error form V_c graph

$$V_c = V_s\left(2e^{-t/R_2C} - 1\right)$$

t_{c_2} = ckt turn off time of T_2

$$V_s\left(2e^{-t/R_2C} - 1\right) = 0\ V_c$$

At $t = t_{c_2}$ $V_c = 0$

$$V_s\left[2e^{-t_{c_2}/R_2C} - 1\right] = 0$$

$$t_{c_2} = R_2 C \ln 2$$

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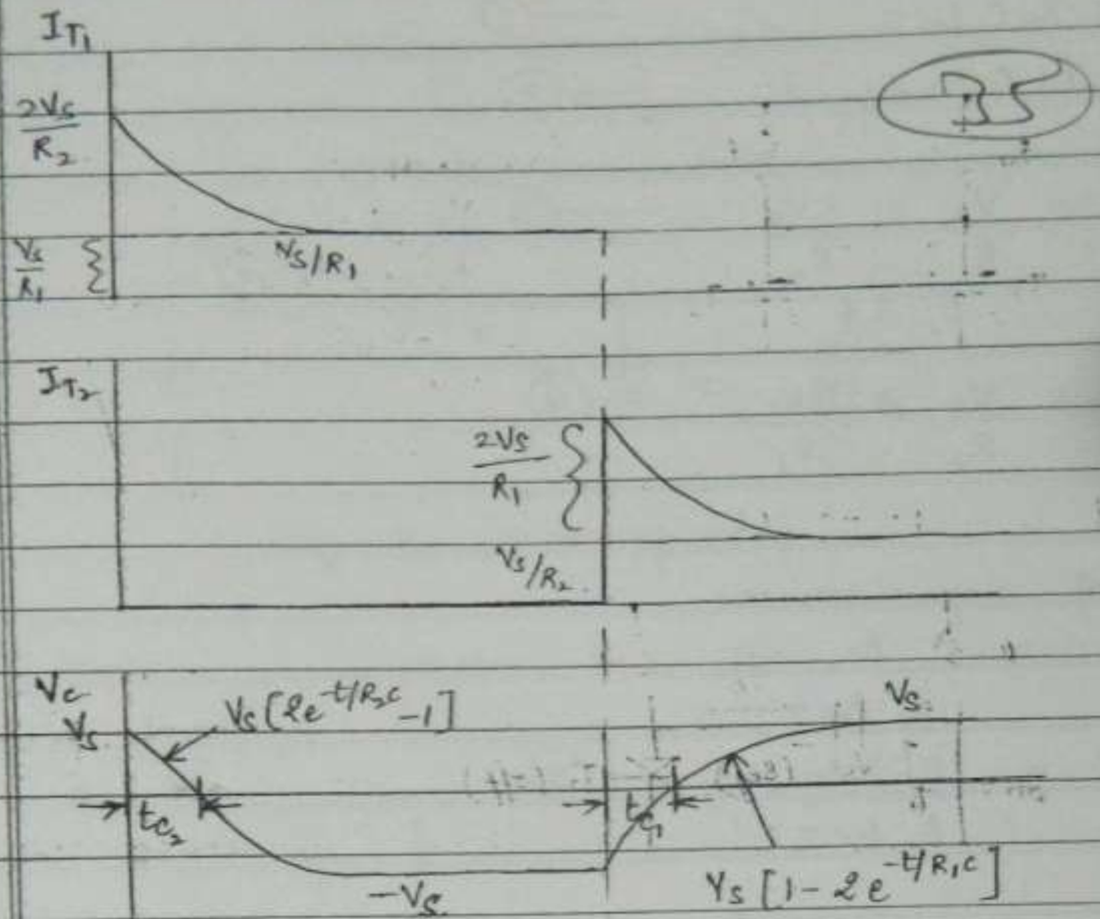


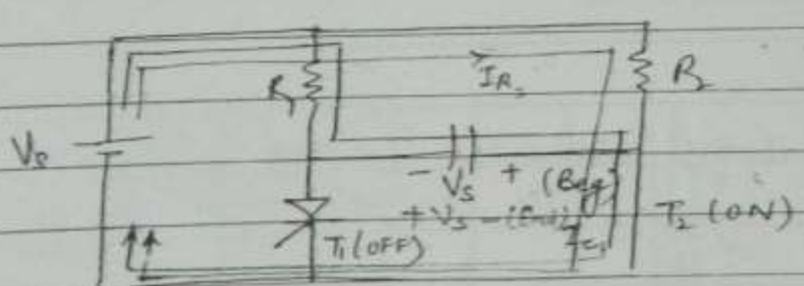
$t=0$ t_1 $t=t_2$ t_2

$T_1 \to ON$

$T_2 \to OFF$ $T_2 = ON$

Mode 2

At $t = t_2$ $T_2 \to ON$



$$I_{T_2} = I_{R_2} + I_C$$

$$I_{T_2} = \underbrace{\frac{V_S}{R_2}}_{\text{steady state current}} + \underbrace{\frac{2V_S}{R_1} e^{-t/R_1C}}_{\text{transient current}}$$

$$\boxed{t_{c_1} = R_1 C \ln 2}$$

* $t_{c_2} = R_2 C \ln 2$ —— (1)
* $t_{c_1} = R_1 C \ln 2$ —— (2)

* $$(I_{T_1})_{peak} = \frac{V_s}{R_1} + \frac{2V_s}{R_2} \quad \text{—— (3)}$$

* $$(I_{T_2})_{peak} = \frac{V_s}{R_2} + \frac{2V_s}{R_1} \quad \text{—— (4)}$$

* Desired value of capacitance

from (1) $C = \dfrac{t_{c_2}}{R_2 \ln 2}$

$$C = \frac{(SF)\, t_q}{R_2 \ln 2} \quad \text{—— (5)}$$

from (2), $$C = \frac{(SF)\, t_q}{R_1 \ln 2} \quad \text{—— (6)}$$

From eq (5) & (6) we get 2 different values for capacitance We must consider the highest value of capacitance to make commutation possible

Applications -

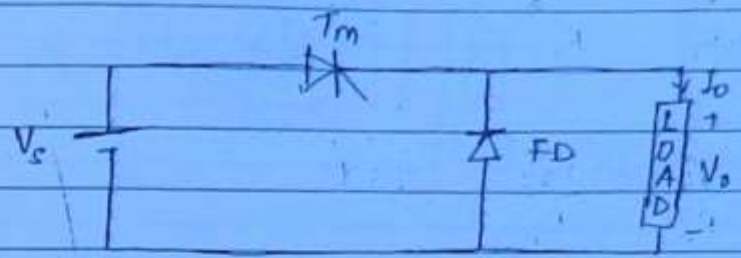
→ Current source ~~interv~~ inverter (CSI)

→ Parallel Inverter

(d) Class D – Voltage Commutation

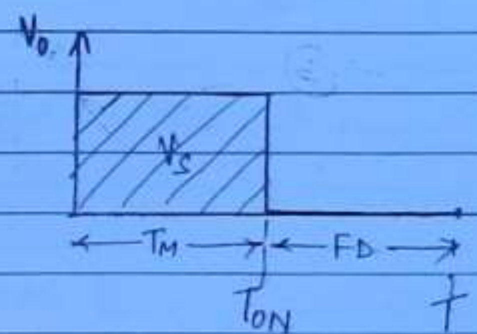
Its used in step down choppers. i.e its also known as voltage commutation chopper.

(37)



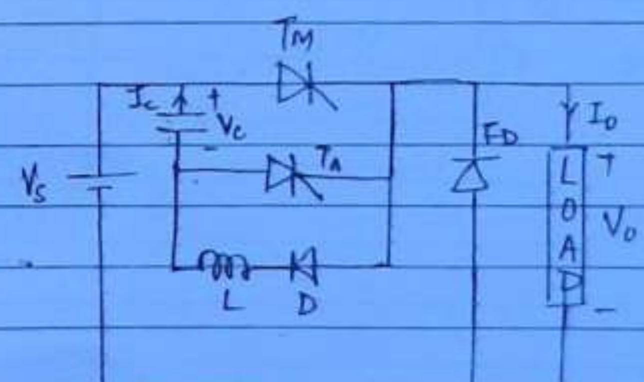
Step down choppers.

without commutation circuit



duty cycle $\alpha = \frac{T_{ON}}{T}$, $V_0 = \frac{\text{Area}}{\text{Time period}} = \frac{V_S T_{ON}}{T}$

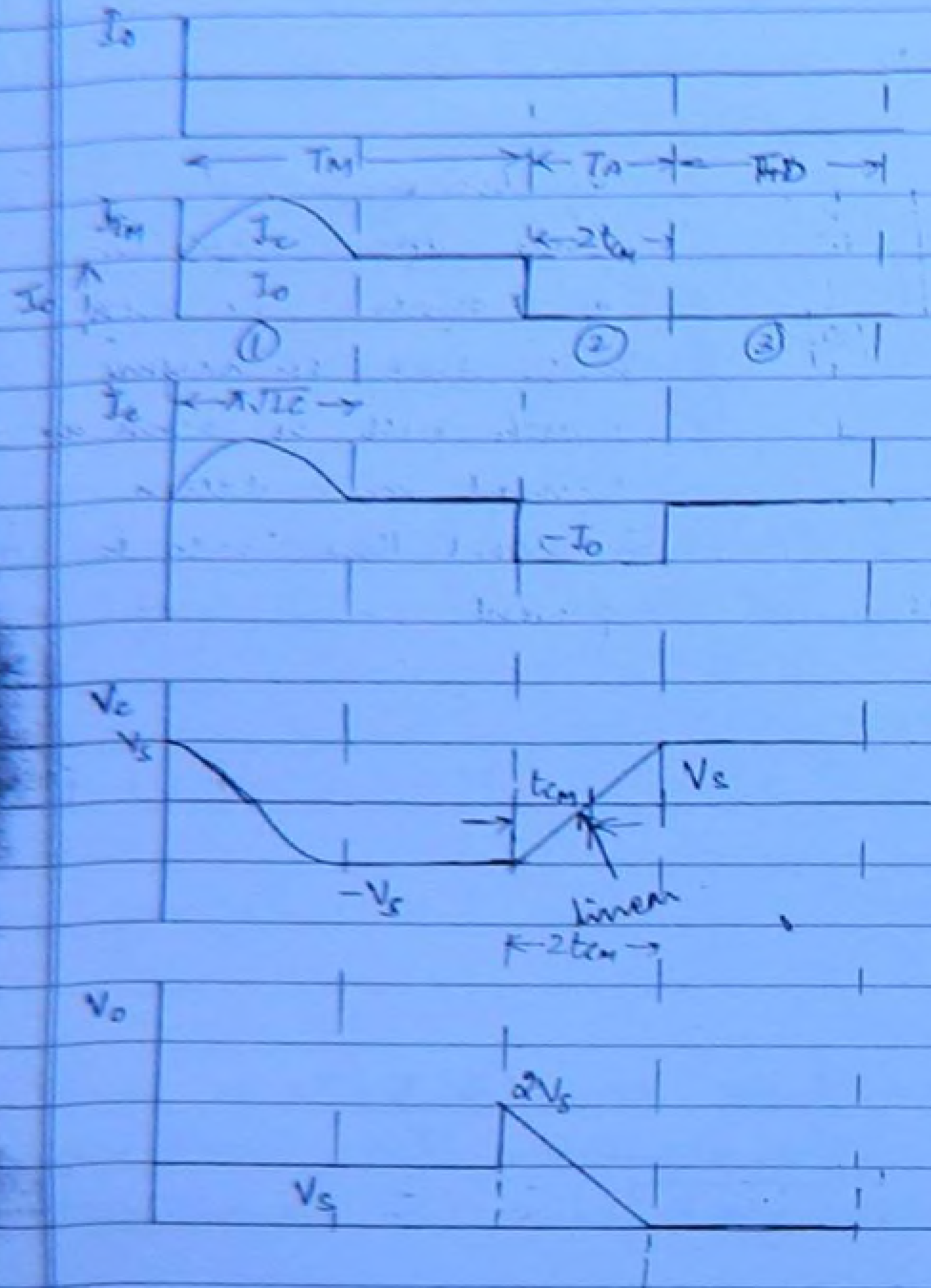
$$\boxed{V_0 = \alpha V_S}$$



Assume

→ $V_c(t=0) = V_s$

→ Consider highly inductive load so that load current I_0 = constant



$t=0$ $\qquad$ T_{ON} $\qquad$ $T_A \to OFF$ $\qquad$ T $\qquad$ $\alpha = \dfrac{T_{ON}}{T}$

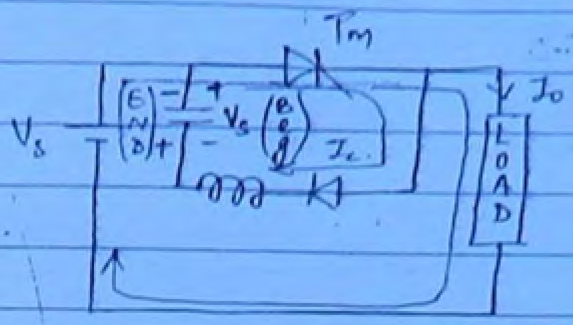
$T_M \to ON$ $\qquad$ $T_A \to ON$

$T_M \to OFF$

Mode (1)

At $t=0$ $T_M \to ON$

(39)



(*) If diode is not there, after completion of mode 1 capacita will start discharging which will be same as current commutation. To avoid this Diode is present.

$I_{T_M} = I_0 + I_c$

$I_c = I_p \sin \omega_0 t$

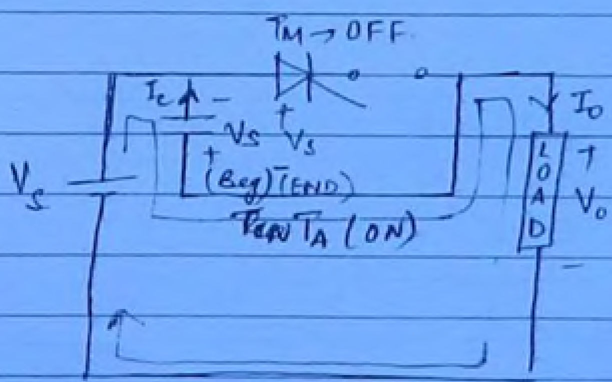
$V_c = V_s \cos \omega_0 t$.

Gnd $\to$ $V_c = -V_c$

$I_c = 0$

Mode (2)

At $t = T_{ON}$ $T_A \to ON$.



$I_c = -I_0$

$V_c = -V_s$

$V_0 = -2V_s$

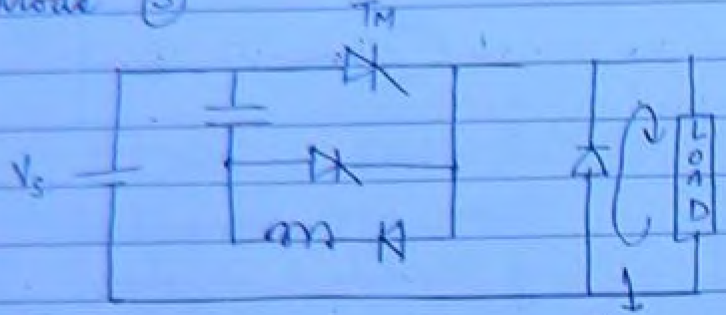
$\to$ $V_c = V_s$

$V_0 = 0$

$I_c = 0$

$T_A \to$ OFF

Mode ③



$V_0 = 0$, FD → ON.

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$$(I_{TM})_{peak} = I_0 + V_S\sqrt{\frac{C}{L}}$$

$$(I_{TA})_{peak} = I_0$$

Without completion of the 1st mode we cannot turn off the TM i.e the minm turn ON time of the thyristor is $\pi\sqrt{LC}$ secs
(this is because the polarities will not change before the completion of 1st mode)

Minm duty cycle of the chopper

$$\alpha = \frac{(T_{ON})_{min}}{T} = \pi\sqrt{LC}\, f$$

Circuit turn off time of TM

$$V_c = \frac{1}{C}\int i\,dt \Rightarrow V_c = \frac{I_0 t}{C} \text{ (linear)}$$

$$V_S = \frac{I_0}{C} t_{CM} \Rightarrow \boxed{t_{CM} = \frac{C V_S}{I_0}}$$

Conduction time of $T_A = 2 t_{CM}$

Commutation interval = time taken to disconnect the load from the source once TM is off.
$= 2t_{r}$.

* PIV of FD is $2V_s$
since its RB by the load (when its off)
so max v/g at load is $2V_s$

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* PIV of $T_M = V_s$

* Average value of voltage

$$V_0 = \frac{V_s T_{ON} + \frac{1}{2} \times 2t_{CM} \cdot 2V_s}{T}$$

$$V_0 = V_s \left[\frac{T_{ON} + 2t_{CM}}{T} \right] = \frac{V_s (T_{ON})_{eff}}{T}$$

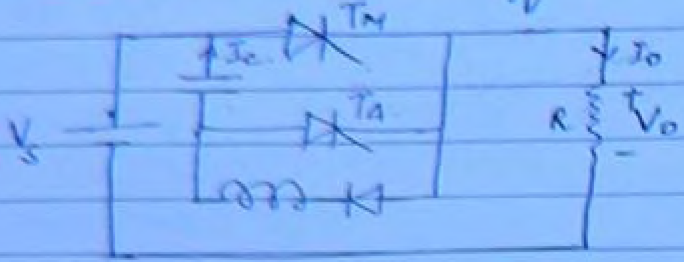
* Effective turn on time of chopper
ie $(T_{ON})_{effective}$

$$T_{on(effective)} = T_{ON} + 2t_{CM}$$

* Minimum possible average voltage of chopper is

$$(V_0)_{min} = V_s \left[\frac{\pi\sqrt{LC} + 2t_{CM}}{T} \right]$$

Find the circuit turn off time of the main thyristor



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Mode 1 is same as prev. case

Mode 2 → is not same as I_0 ↑ crnt due to R load.
(which forms RC ckt with cap C)

$$I_c = -I_0 = \frac{-2V_s}{R}\, e^{-t'/RC}$$

$$I_0 = \frac{2V_s}{R}\, e^{-t'/RC}$$

At $t' = 0$

$$I_0 = \frac{2V_s}{R}$$

& load current is exponentially reducing.

$$I_c = \frac{-2V_s}{R}\, e^{-t'/RC}$$

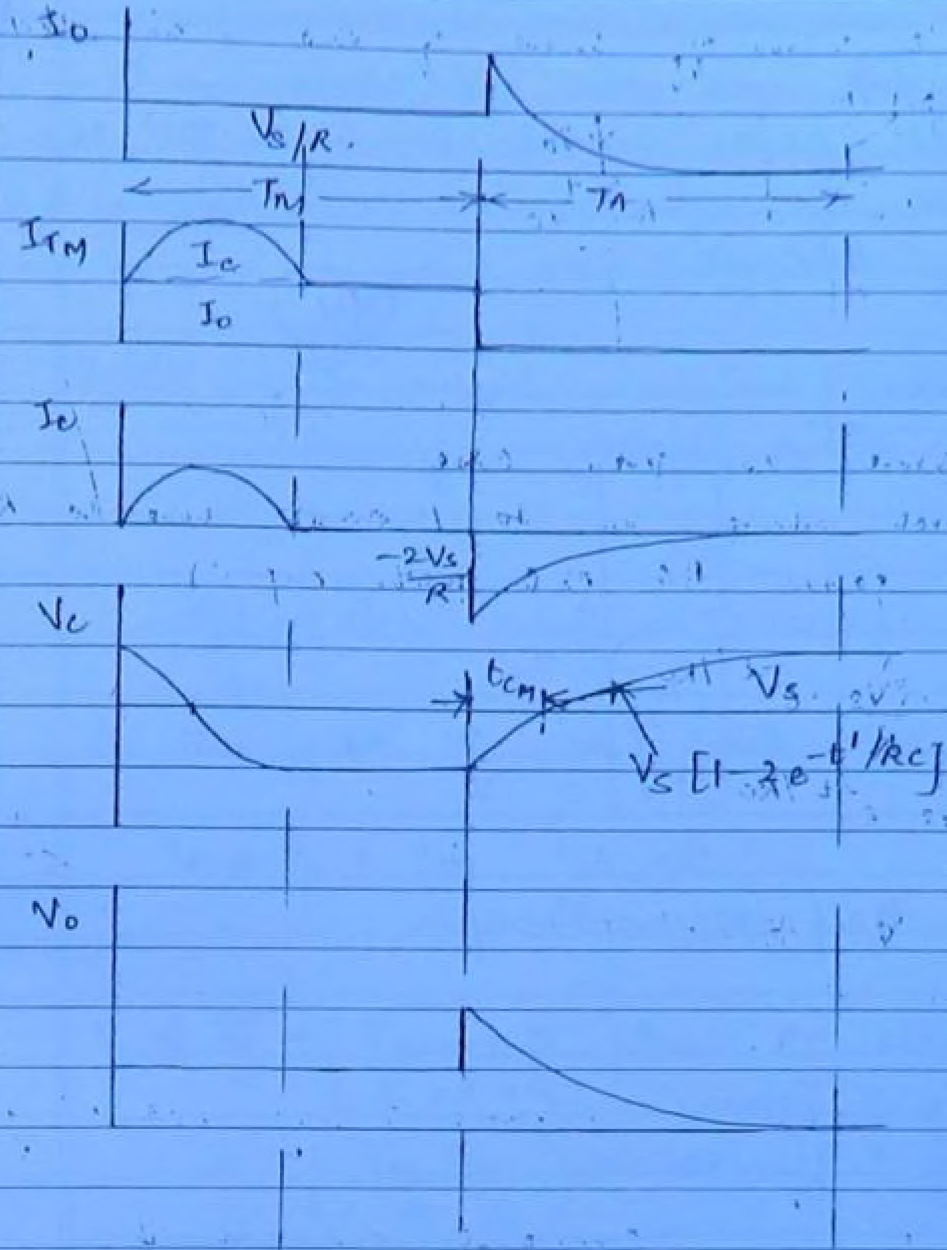
current current ↓ exponentially

V_c also

$$V_c = V_s\left[1 - 2e^{-t/RC}\right]$$

At $t = t_{cm}$ $\quad V_c = 0$

$$t_{cm} = RC \ln 2$$

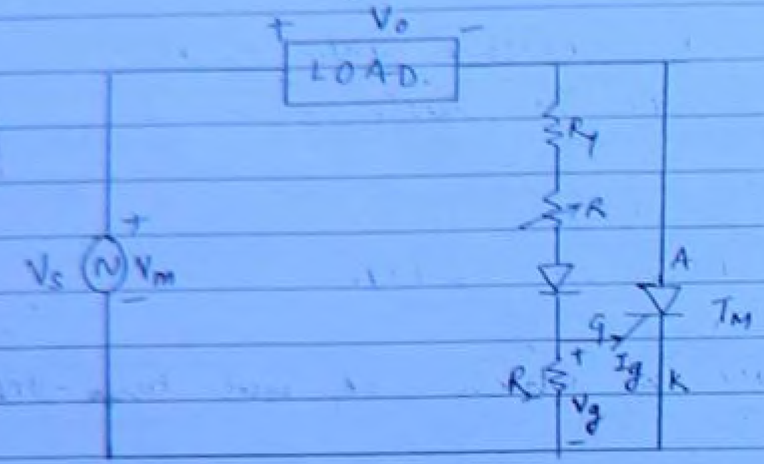

Io
Vs/R
TM
TA
ITM
Ic
Io
Ic
-2Vs/R
Vc
tcm
Vs
Vs [1-2e^{-t'/RC}]
Vo

FIRING CIRCUITS

gives the required gate signal to turn ON the SCR.

1 Resistance firing circuit -

(44)



Main ckt : 1 ϕ Half Wave Rectifier

Gate specifications →

$$I_{g\,min} \le I_g \le I_{g\,max}$$
$$V_{g\,min} \le V_g \le V_{g\,max}$$

R_1 → To limit gate current I_g within max^m value ($I_{g\,max}$)

For worst condition.

Maximum gate current $= \frac{V_M}{R_1} \le I_{g\,max}$

$$\therefore R_1 \geq \frac{V_M}{I_{g\,max}}$$

R_2 → To limit gat vlg V_g within max^m value ($V_{g\,max}$)

For worst condition.

Maximum gate vlg $= \left(\frac{V_M}{R_1 + R_2}\right) R_2 \le V_{g\,max}$

From above eq^n we can design value of R_2.

Variable R → To change the timing of gate signal ie α

(45)

Diode → To avoid negative gate signal during negative cycle of source.

$\boxed{V_{gt}}$ → Gate turn on voltage

Its the gate Vg at which SCR will turn-ON

ie at $V_g = V_{gt}$ SCR → ON ($\omega t = \alpha$)

$$V_g = \left(\frac{V_m \sin \omega t}{R_1 + R + R_2}\right) R_2$$

$$V_g = \left(\frac{V_m R_2}{R_1 + \uparrow R + R_2}\right) \sin \omega t$$

$$\boxed{V_g = V_{gm} \sin \omega t} \quad \text{where } \downarrow V_{gm} = \frac{V_m R_2}{R_1 + \uparrow R + R_2}$$

$V_g = 0$

At $V_g = V_{gt}$ SCR → ON

$V_{gm} \sin \alpha = V_{gt}$

$$\boxed{\uparrow \alpha = \sin^{-1} \frac{V_{gt}}{\downarrow V_{gm}}}$$

$\uparrow R \quad V_{gm} \downarrow \quad \alpha \uparrow$

for example

(I) $R = R_a$

$\alpha = \alpha_a$

$$V_{gm_a} = \frac{V_m R_2}{R_1 + R_a + R_2}$$

$V_{g_a} = V_{gm_a} \sin \omega t$

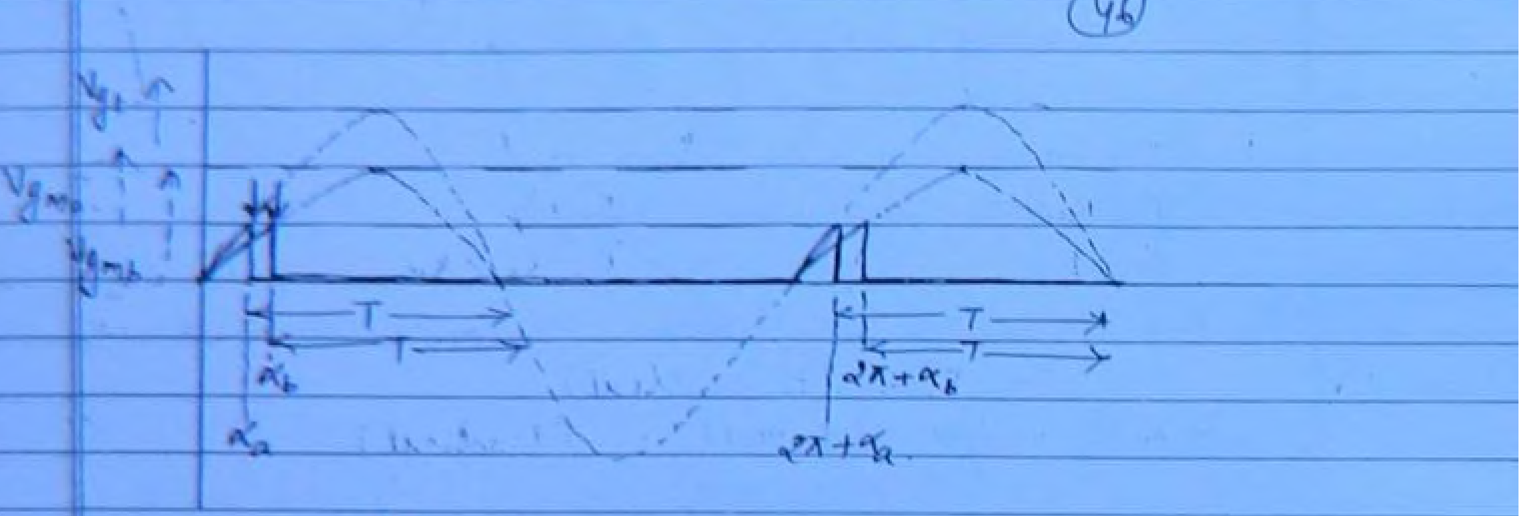
II $\uparrow R = R_b$

$\alpha = \alpha_b$

$V_{gm_b} < V_{gm_a}$

$V_{g_b} = V_{gm_b} \sin \omega t$

(4b)

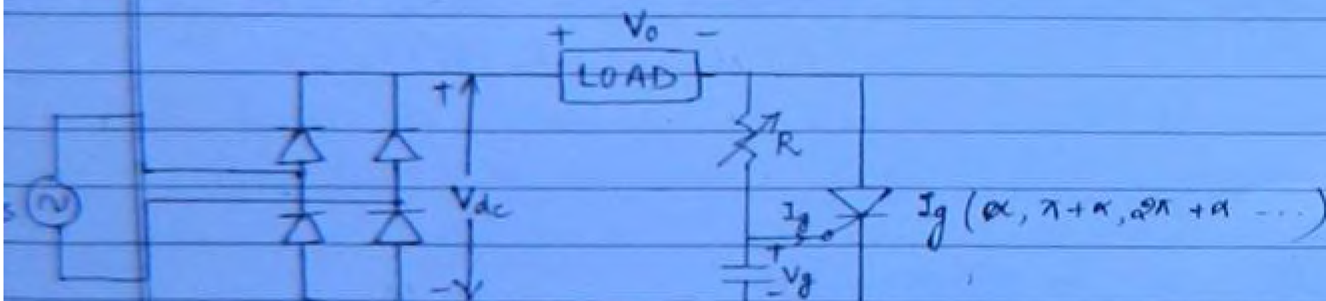


$\uparrow R \quad V_{gm} \downarrow \quad \therefore \alpha \uparrow$

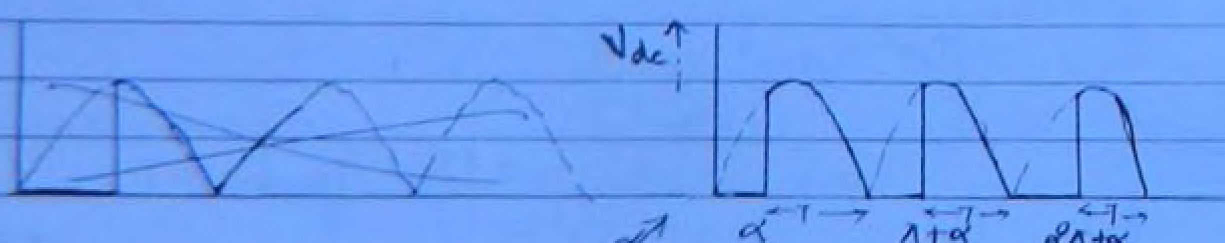
Limitation of R firing circuit

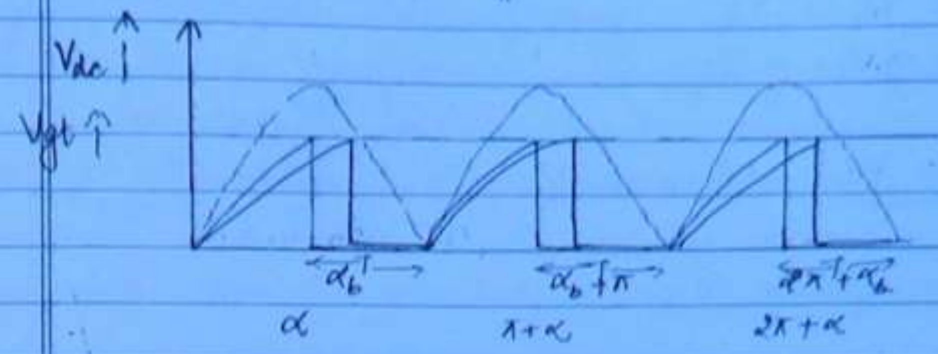
The maximum firing angle is limited to 90°

2 RC firing circuit -



Main circuit : Full wave Rectifier





I $R = R_a$

$T = R_a C$

$V_{g_a} \uparrow$

II $V_{gb} \uparrow$

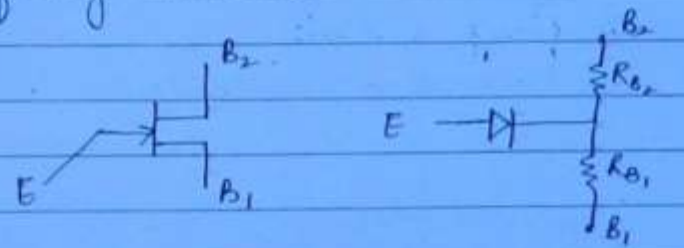
$T = R_b C$

$R = R_b > R_a$.

$$\boxed{\begin{array}{c} 0 < \alpha < 180^\circ \ (\text{Ideal}) \\ (5 \text{ to } 7^\circ) \leq \alpha \leq (165 \text{ to } 175^\circ) \ (\text{Practical}) \end{array}}$$

3. UJT firing circuit –

UJT



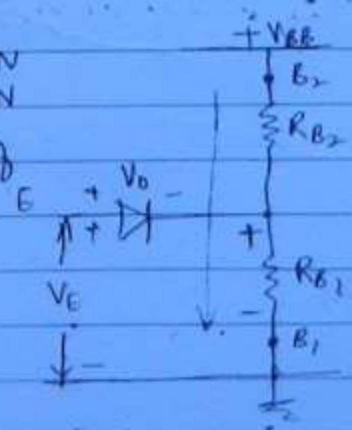
$B_1, B_2 \rightarrow$ Base terminals

$R_{B_1}, R_{B_2} \rightarrow$ Base resistances

$E \rightarrow$ Emitter terminal

When diode is ON UJT ON

When diode is off UJT OFF



$$V_{R_{B_1}} = \left(\frac{R_{B_1}}{R_{B_1} + R_{B_2}} \right) V_{BB}$$

$$\boxed{V_{R_{B_1}} = \eta V_{BB}}$$

$$\eta = \frac{R_{B_1}}{R_{B_1} + R_{B_2}}$$

Intrinsic stand off ratio

$V_E = V_{R_{B_1}} + V_D$
$= \eta V_{BB} + V_D$

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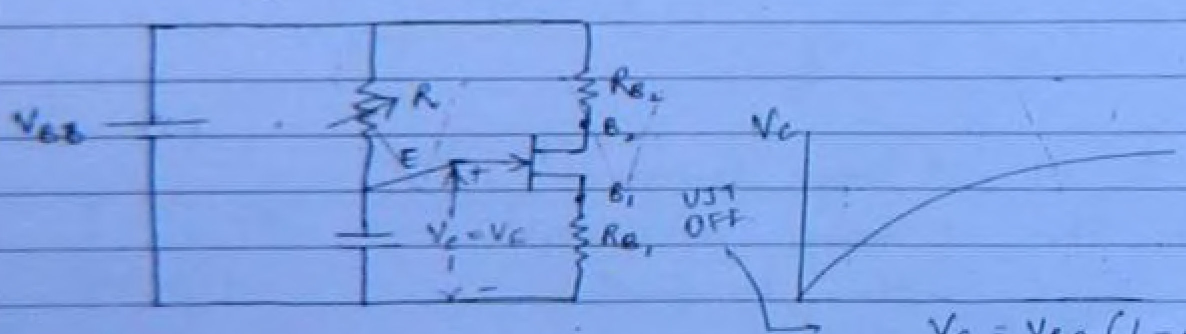
$\boxed{\uparrow V_E \Rightarrow V_P}$ then UJT → ON. (when ↑ V_E reaches V_P UJT → ON)

$V_P = \eta V_{BB} + V_D$
↓
peak point voltage

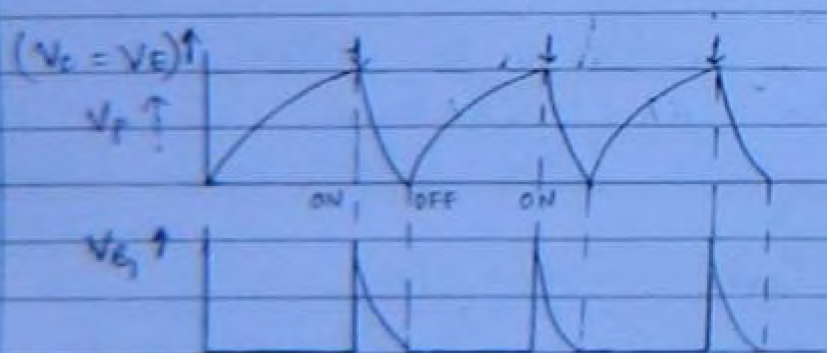
* When UJT is switching from OFF to ON state, the base resistance R_{B_1} starts decreasing i.e UJT exhibits negative resistance behaviour. This reduces emitter vlg.

$\boxed{V_E \downarrow \Rightarrow V_V}$ ⟶ UJT OFF. (when ↓V_E reaches V_V UJT → OFF)
↓
Valley voltage

UJT working as Relaxation Oscillator –

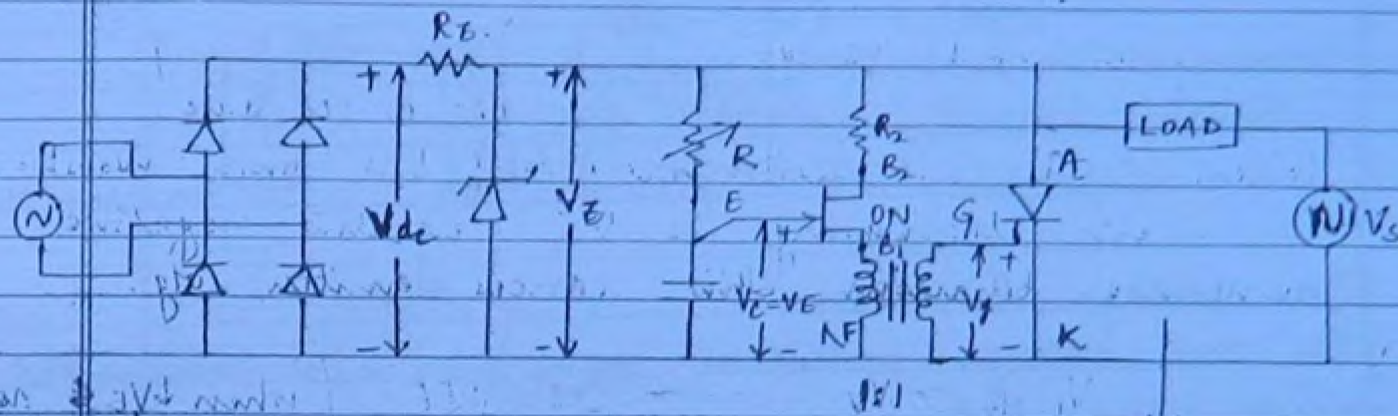


$V_C = V_{BB}(1 - e^{-t/RC})$

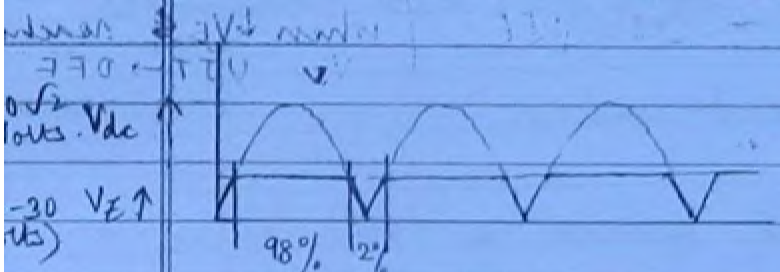


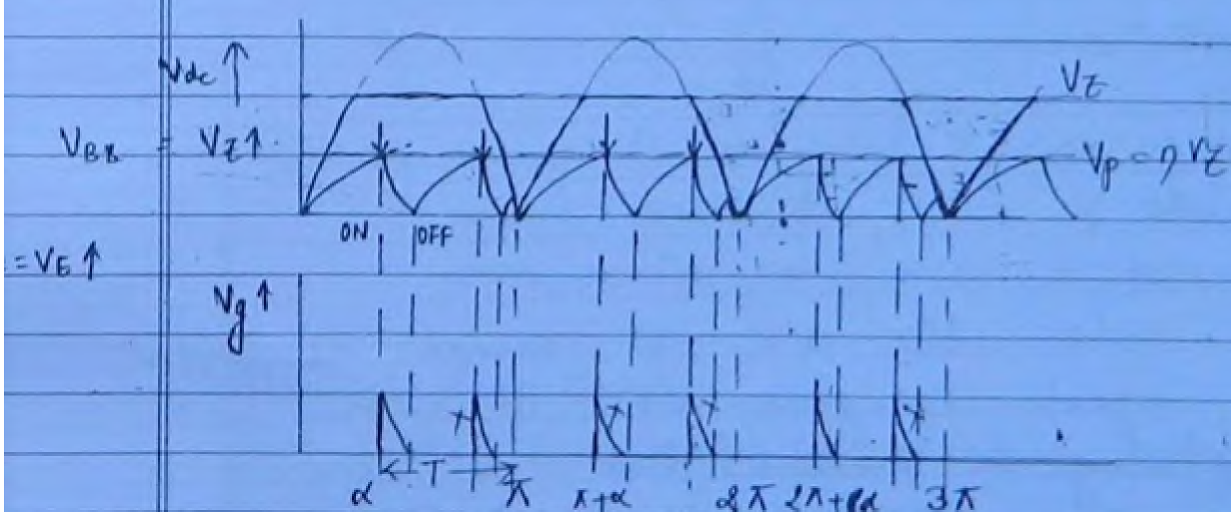
Synchronized UJT firing circuit -

We must synchronise the firing circuit with the main circuit in order to match the timing of gate pulse in both the circuits. Here we must use same power supply in the main & firing circuits for the purpose of synchronisation.



main circuit
Half wave rectifier
$I_g(\alpha, 2\pi+\alpha, ——)$



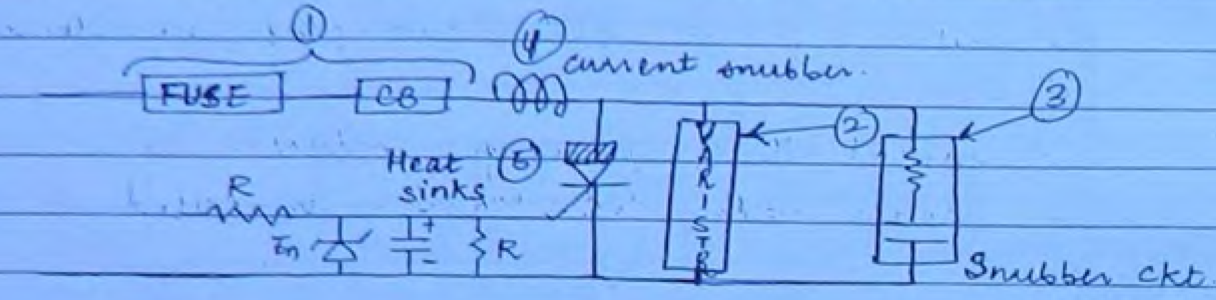


PROTECTION OF THYRISTORS -

50

1 **Over Current Protection -**

For over current protection, we must connect fuse or circuit breaker in series with the SCR.



2 **Over Voltage Protection -**

For over voltage protection, varistors are connected across the SCR

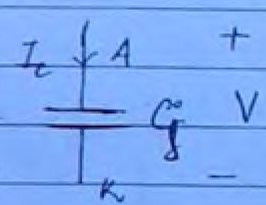
Varistors → Non-linear resistor

↓

All metal oxide resistors behave as non-linear R.

3 **dv/dt Protection -**

$$\uparrow I_c = C_j \frac{dV}{dt} \uparrow$$

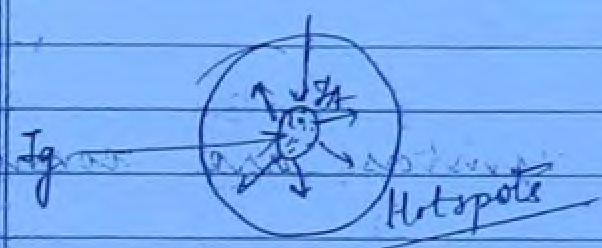


At high dV/dt the SCR is turned ON before the gate signal is given. This is known as false triggering. To prevent this a capacitor is connected across SCR to limit dV/dt. A resistor is connected in series with the C, to reduce the discharge current magnitude. This is called Snubber circuit

4 di/dt Protection

when di/dt > (spread velocity of charge carriers) the charge accumulation increases cummulatively in a small conduction area & leads to the formation of hot spots damaging the device.

To prevent this, a large inductor is connected in series with the SCR. This is called current snubber



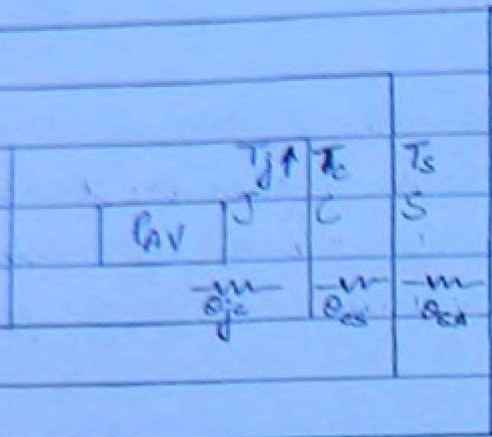
Initial conduction area ↑

(*) di/dt capability of thyristor can be improved by:-

i) → by increasing Ig or
→ by increasing $\frac{dI_g}{dt}$

ii) → by using Centre Gated Thyristor
(initial conduction area is increased when centre gated SCR is prefered)

5 Thermal Protection -
Heat sinks are used for thermal protection.



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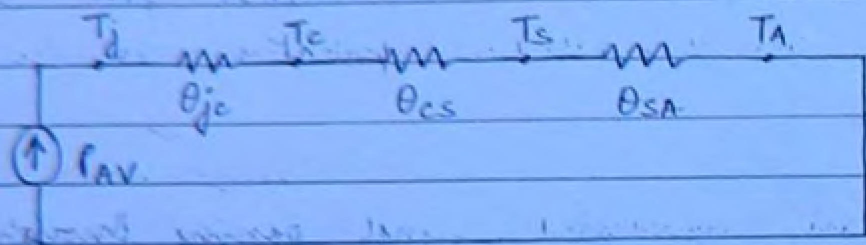

$$P_{AV} = \frac{T_j - T_c}{\theta_{jc}} = \frac{T_c - T_s}{\theta_{cs}} = \frac{T_s - T_A}{\theta_{SA}} = \frac{T_j - T_s}{\theta_{jc} + \theta_{cs}} = \frac{T_c - T_A}{\theta_{cs} + \theta_{SA}}$$

$$= \frac{T_j - T_A}{\theta_{jc} + \theta_{cs} + \theta_{SA}}$$

* Rating of SCR is $\propto \sqrt{P_{AV}} \propto \sqrt{\dfrac{T_j - T_A}{\theta_{jA}}}$

* Rating of SCR is decided by cooling methods in the heat sink. Lesser the ambient temperature (T_A) higher the rating of the SCR.

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6 Gate Protection.-

a) Over Current Protection
A resistance is ~~an~~ connected in series with the gate to limit the gate current within the permissible value.

b) Over Voltage Protection
Zener diode is connected across gate cathode terminals for overvoltage protection in the gate.

c) Protection against noise signals-
Noise is an unwanted signal passing through the gate terminal. It will false turn ON the SCR. To prevent it, ~~am~~ connect a parallel RC across gate cathode terminals, to protect the SCR against noise signals.

CWB chapter 1

Q1 (b) $T_j = 125°C$
$T_s = 70°$
$\theta_{jc} = 0.16$
$\theta_{cs} = 0.08$

$$P_{AV_1} = \frac{T_j - T_s}{\theta_{jc} + \theta_{cs}} = \frac{125 - 70}{0.16 + 0.08} = 229.167 \text{ W}$$

$$P_{AV_1} = \frac{125 - 60}{0.16 + 0.08} = 270.83 \text{ W}$$

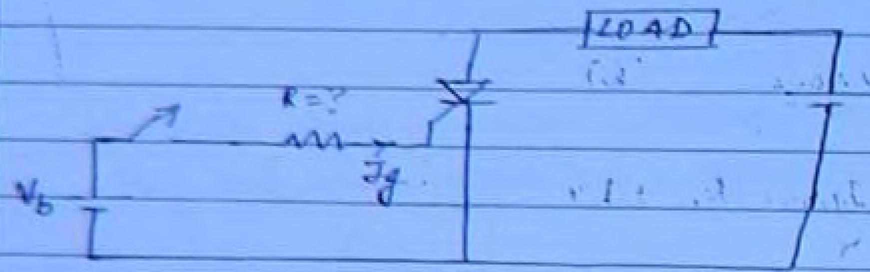
% Increase in Rating of SCR $= \dfrac{\sqrt{P_{AV_2}} - \sqrt{P_{AV_1}}}{\sqrt{P_{AV_1}}} \times 100$

(54)

$$= \frac{\sqrt{270.83} - \sqrt{229.16}}{\sqrt{229.16}} \times 100$$

$$= 8.7\%$$

Q2



$V_b = 12 \pm 4$ $\qquad I_{g\,min} = 100\ mA$

For worst condition

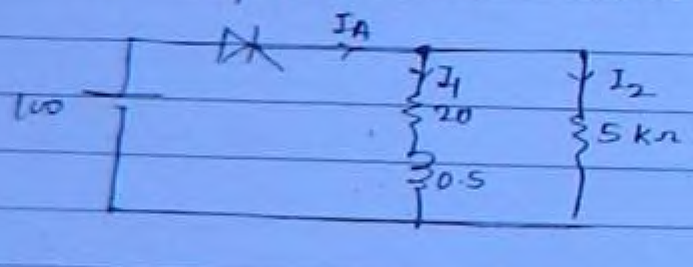
Minimum possible $I_g = \dfrac{12-4}{R}$

$$\frac{12-4}{R} \geq 10\ mA$$

$$R \leq 800\ \Omega \qquad (d)$$

Q5 $\quad C = \dfrac{SF \times tq}{R \ln 2} = \dfrac{2 \times 50 \times 10^{-6}}{50 \times \ln 2} = 2.88\ \mu F \quad (a)$

Q6 $\quad T_{ON} = 5\ \mu sec \qquad I_L = 50\ mA \qquad I_H = 40\ mA$



$$I_A = I_1 + I_2$$

$$= \frac{V_S}{R_1}\left(1 - e^{-t/T_1}\right) + \frac{V_S}{R_2}$$

$$T_1 = \frac{L_1}{R_1} = \frac{0.5}{20} = \frac{1}{40}$$

$$I_A = \frac{100}{20}\left(1 - e^{-40t}\right) + \frac{100}{5 \times 10^3}$$

$I_A = 5(1-e^{-40t}) + (20\times10^{-3})$

$\downarrow$

till I_L

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$50\times10^{-3} = 5(1-e^{-40t}) + (20\times10^{-3})$

$\frac{30\times10^{-3}}{5} = 1-e^{-40t}$

$t = 150\,\mu\text{secs}$ (b)

Q7

$9V = 1V + I_{gmax} R + 1V$

$I_{gmax} = 150\ mA$

$R \geq 46.67\,\Omega$

$9V = 1V + I_{gmin} R + 1V$

$I_{gmin} = 100\ mA$

$R \leq 70\,\Omega$

$46.6 \leq R \leq 70$

Ans $47\,\Omega$ (c)

Q8. Volt sec rating of pulse transformer

$= 10V \times t_{gpw}$ (gate pulse wid

$t_{gpw} > t_{min}$

$I_A = \frac{200}{1}(1-e^{-t/T})$ $\frac{L}{R} = 150\times10^{-3} = 0.15$

$I_A = 200(1-e^{-t/0.15})$

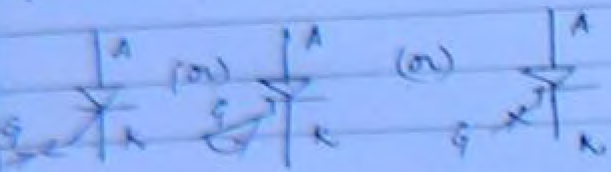
$I_L \rightarrow 250\,mA = 200(1-e^{-t/0.15})$

$t_{min} = 187\,\mu s$

$t_{pgw} > 187\,\mu s$

$= 200\,\mu s$ Ans (a)

GTO (Gate Turn OFF Thyristor)

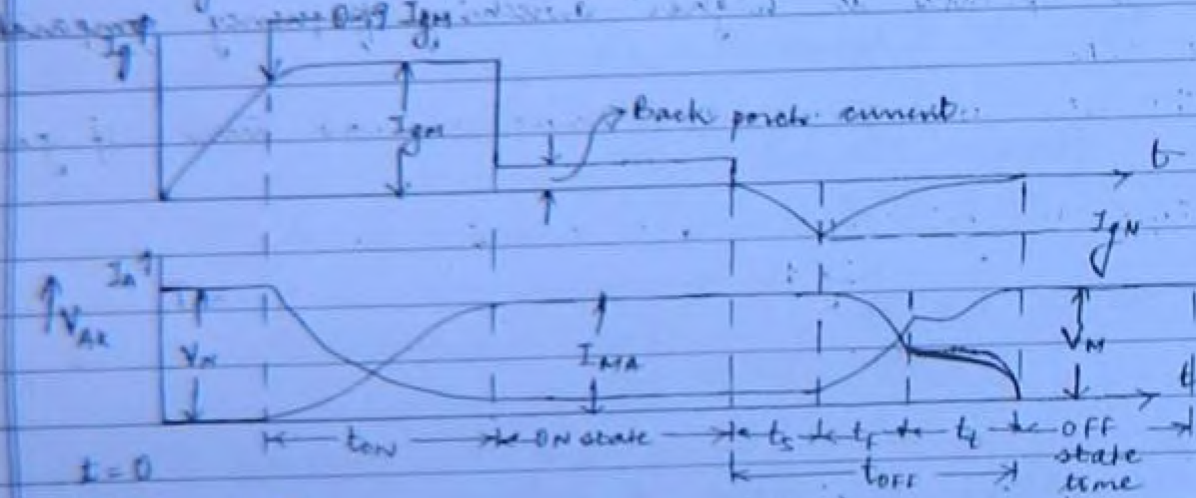


5f

To turn ON → + I_g (when A +ve w.r.t K).
To turn OFF → − I_g (when I_{gN} = 20-25% I_{MA}]

The V-I characteristics of GTO & conventional thyristor (SCR) are similar.

Switching Characteristics of GTO –



t = 0
$\underbrace{\text{OFF} \longrightarrow \text{ON}}$
turn ON char.

* During storage time the stored charge carriers are removed ~~from~~ in the device

t_s = storage time

* During fall time rate of reduction of anode current is fast t_f = fall time

* During tail time rate of reduction of anode current is slow.

t_t = tail time

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Compare GTO with conventional thyristor (SCR)

1. I_L & I_H are higher in GTO.
2. On state vlg drop is higher in GTO.
3. Gate signal requirement is higher in GTO.
4. Reverse vlg blocking capability is lesser than forward vlg blocking capability in GTO.
5. GTO unit is more efficient & compact compared to SCR unit.
6. GTO has fast t_{ON} & ∴ faster t_{OFF}
∴ it operates at higher switching frequency compared to SCR.
7. GTO has low turn-on gain & low turn-off gain.

$\downarrow$ turn-on gain $= \dfrac{I_{MA}}{(+)I_g \uparrow}$ $\downarrow$ turn-off gain $= \dfrac{I_{MA}}{-I_{gM}\uparrow}$

Applications -

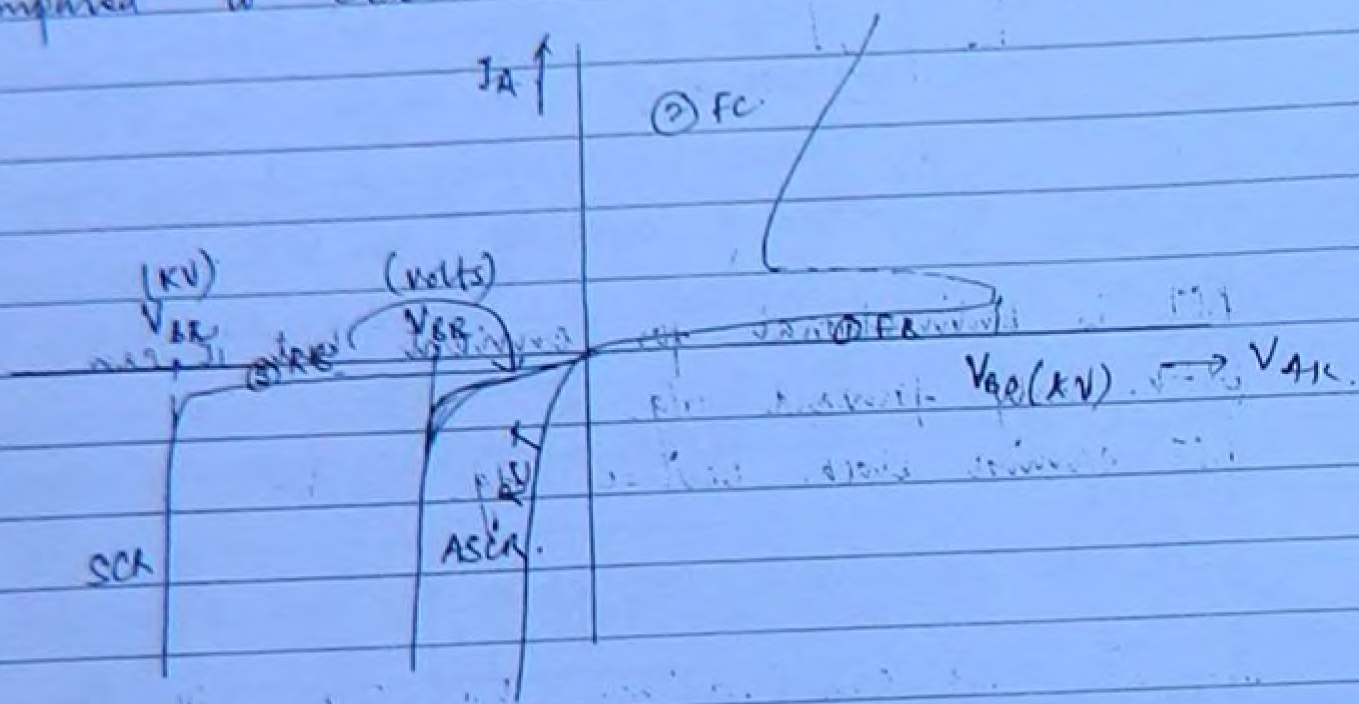
* In inverters & choppers we can replace the SCR by using a GTO to avoid commutation circuit.

ASCR (Asymetrical SCR)

Its a special thyristor with reduced reverse v/g blocking capability.

(58)

ASCR has fast turn-on & turn off times ∴ it operates at higher switching frequency as compared to SCR.



Applications -

In v/g source inverters (VSI) we can replace the SCR by ASCR.

RCT (Reverse Conducting Thyristor)

An antiparallel diode is inbuilt across the SCR within the same structure.

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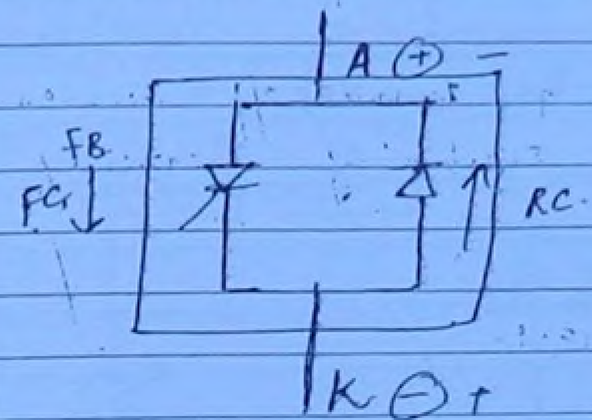
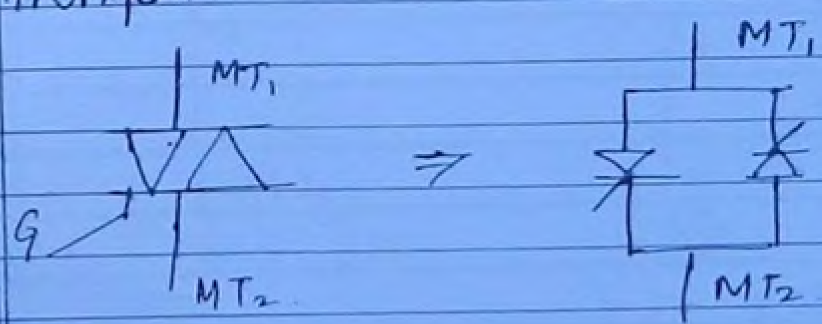
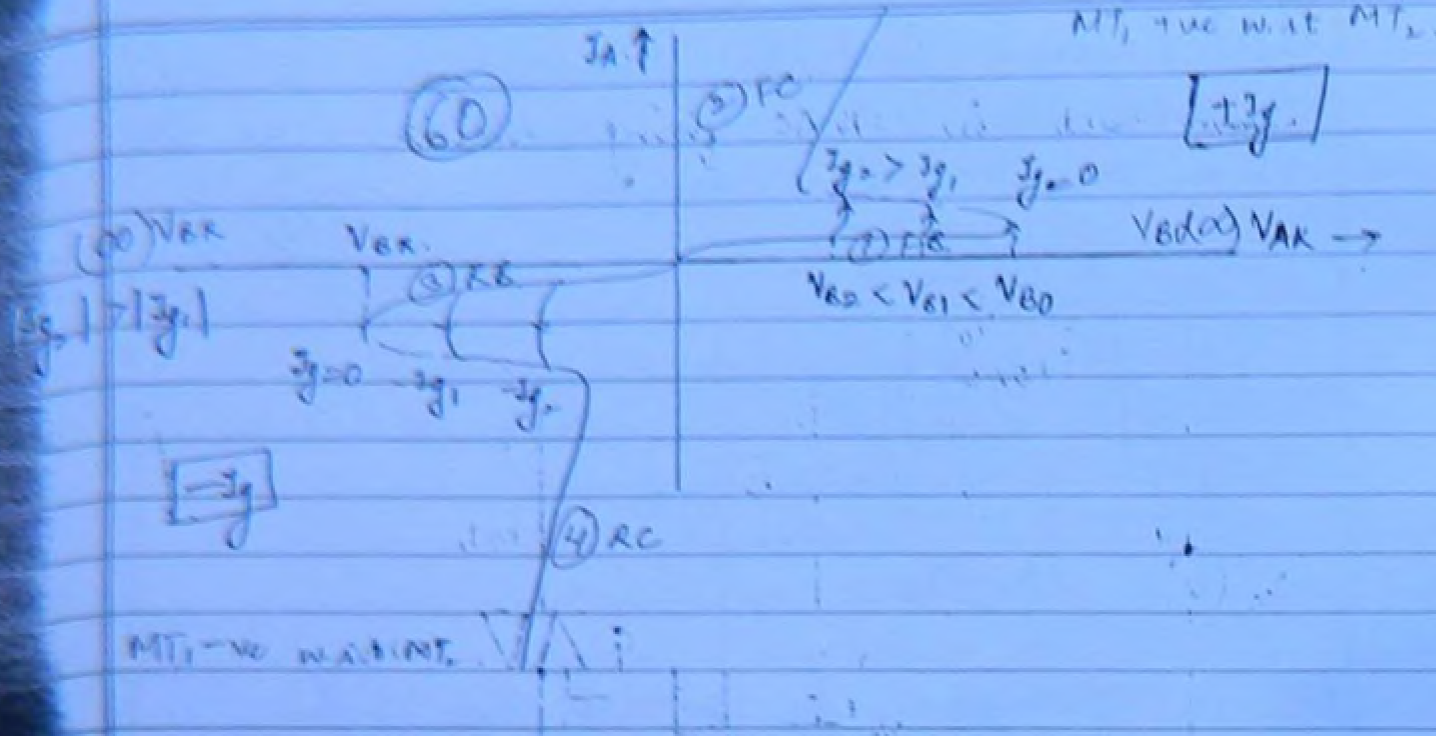


RCT is bidirectional for current but it can block (only) forward vlg.
RCT cannot block reverse vlg.

Applications –
In VSI, we can replace the SCR with an antiparallel diode by RCT.

TRIAC –

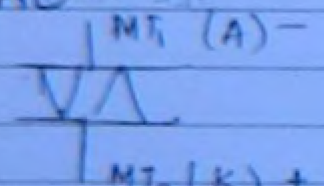


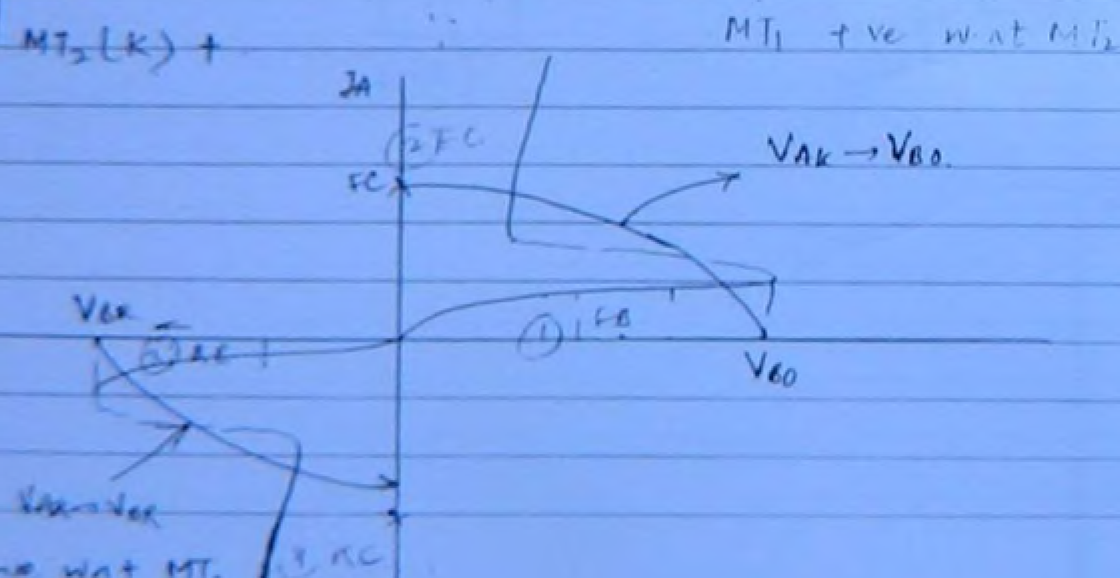


Applications –
used in AC v/g controllers as AC switch

Limitations of Triac –
In AC v/g controllers, its used only for resistive loads and low inductive loads. Its not preferred for high inductive loads with high time constant.

DIAC

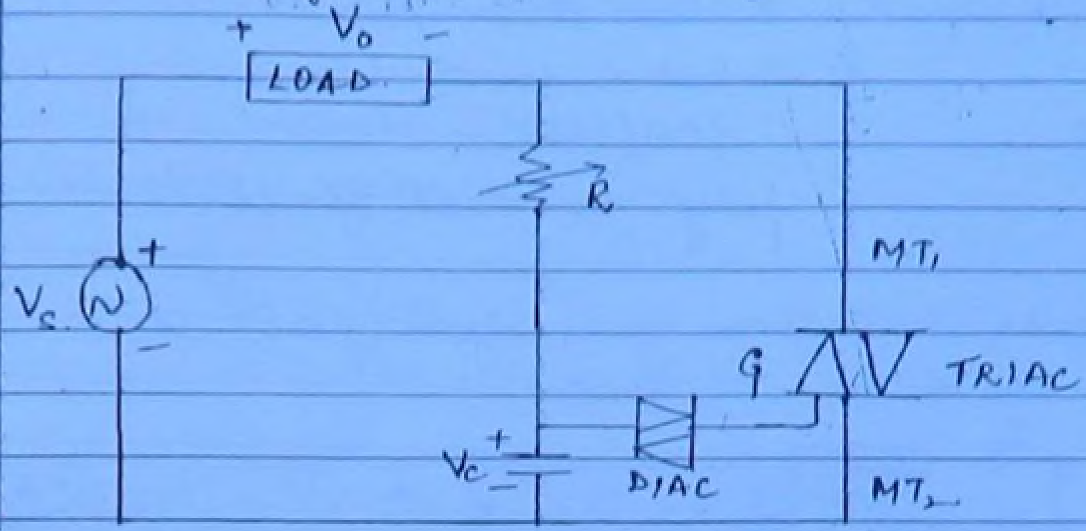




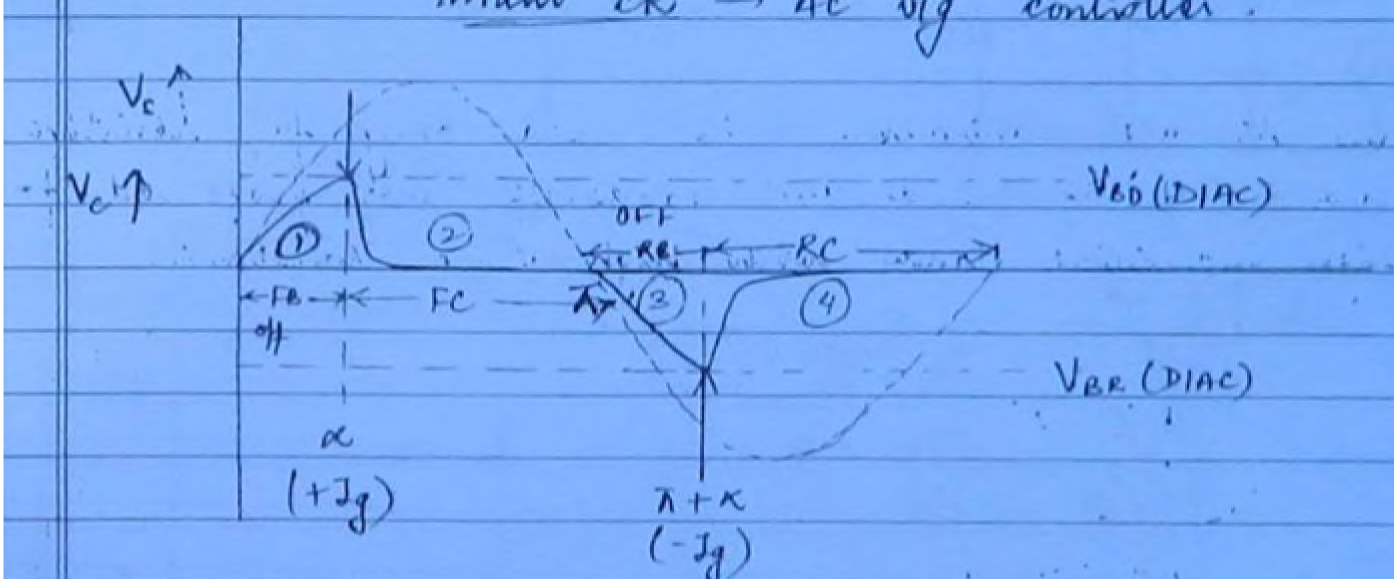
Applications -

~~Main~~ used in TRIAC firing circuit. (61)

TRIAC firing circuit using DIAC →



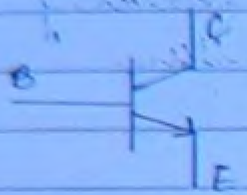
main ck → AC v/g controller.



POWER TRANSISTORS –

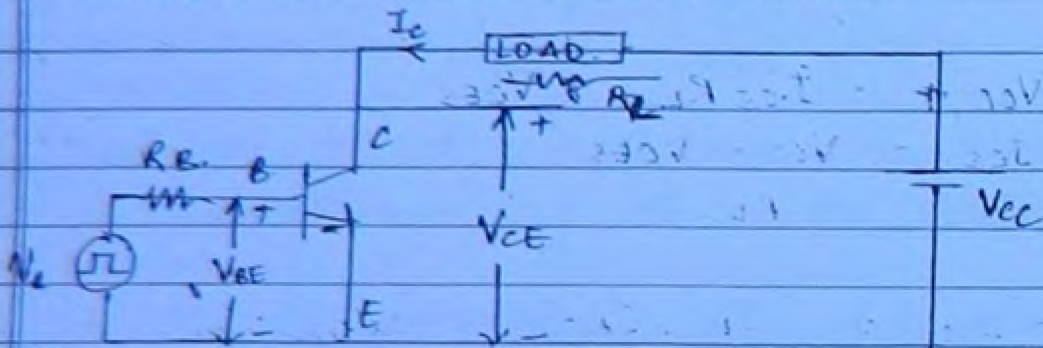
(62)

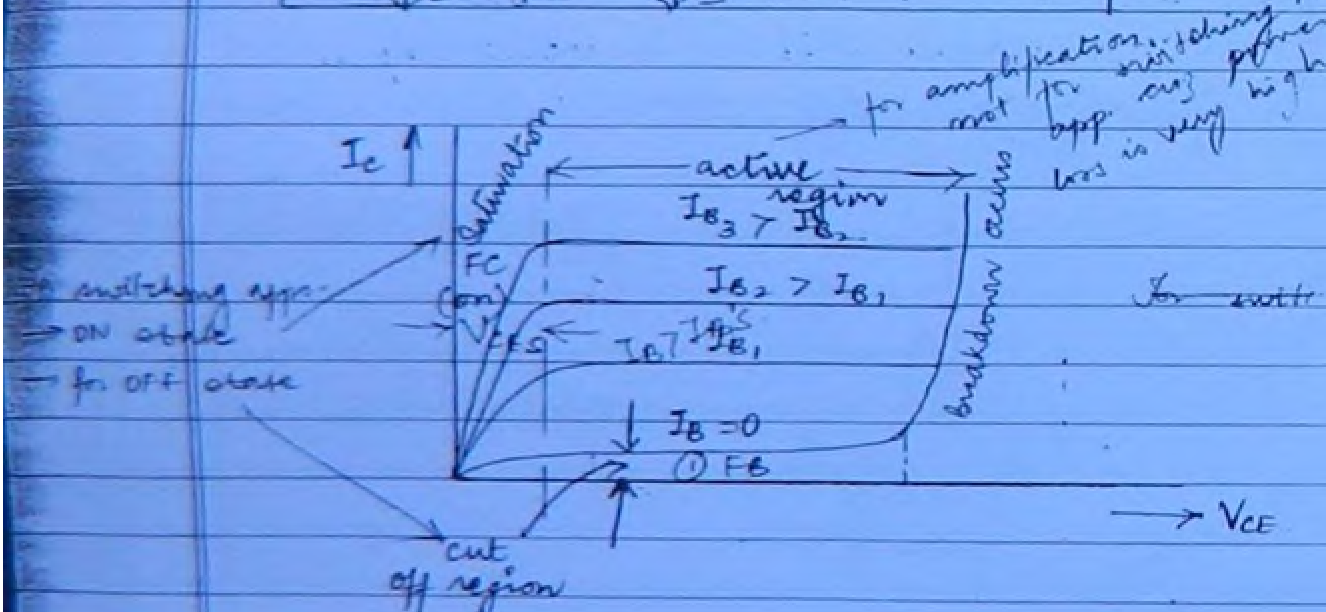
1 POWER BJT –



C, E ⇒ Major terminals
B ⇒ Control terminal
(ON/OFF)
∴ fully controlled device

* Here we require continuous gate signal (base) to maintain the device in ON state.





I_{BS} → min^{m} base current required to drive the transistor into saturation

* For switching applications in PE, the transistor should be operated in the cut-off region for OFF state & saturation region for ON state. Active region is not preferred for switching applications. Its only preferred in amplifiers. 63

$I_C = \beta I_B$

Consider transistor in :

Saturation region:

$$V_{CE} = V_{CES}$$
$$I_C = I_{CS}$$

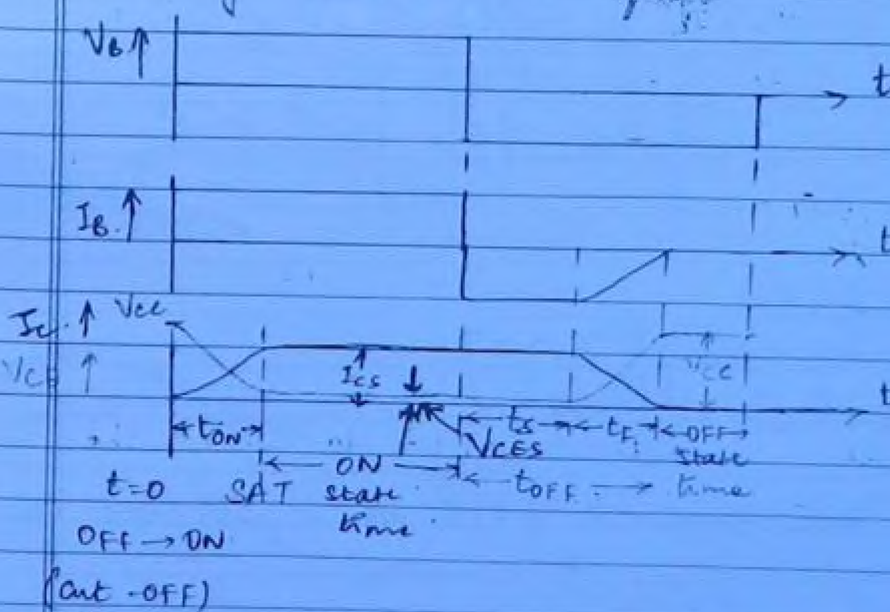
KVL $\rightarrow$ $V_{CC} = I_{CS} R_L + V_{CES}$

$$I_{CS} = \frac{V_{CC} - V_{CES}}{R_L}$$

$I_{BS} = \frac{I_{CS}}{\beta}$ & $I_B \geq I_{BS}$ $\rightarrow$ saturation

$I_B \leq I_{BS}$ $\rightarrow$ active region

Switching characteristics of power BJT -



t_s During storage time, stored charges present in the base region is removed.

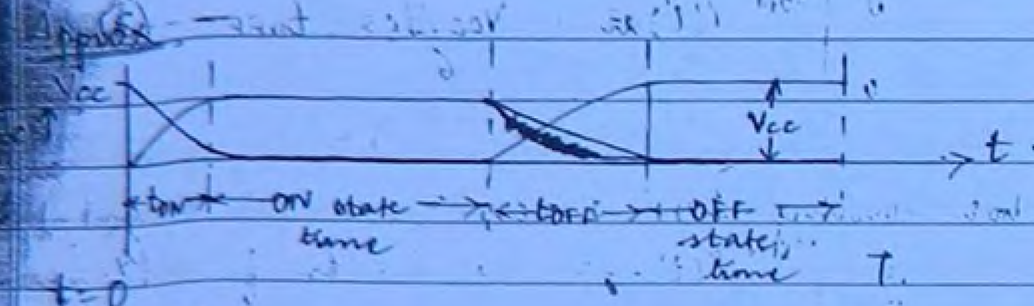
(64)

Instantaneous power loss during t_{ON} process $P(t)$ $\Big] = V_{CE}(t)\, I_c(t)$

Energy lost during turn-on process $= \int_0^{t_{ON}} P(t)\, dt.$

Avg power lost during turn-on process $= \frac{1}{T} \int_0^{t_{ON}} P(t)\, dt.$

Approx.



t_{ON} process

$$I_c = \left(\frac{I_{cs}}{t_{ON}}\right) t$$

$$V_{CE} = \left(\frac{-V_{cc}}{t_{ON}}\right) t + V_{cc}$$

t_{OFF} process

$$I_c = \left(\frac{-I_{cs}}{t_{OFF}}\right) t + I_{cs}$$

$$V_{CE} = \left(\frac{V_{cc}}{t_{OFF}}\right) t$$

$$P(t) = V_{CE}(t)\, I_c(t)$$
$$= \left[\left(\frac{-V_{cc}}{t_{ON}} t\right) + V_{cc}\right]\left[\left(\frac{I_{cs}}{t_{ON}}\right) t\right]$$

Energy lost during t_{ON} process $= \int_0^{t_{ON}} P(t)\, dt \xrightarrow{\text{solve}} = \frac{V_{cc}\, I_{cs}}{6} t_{ON}$ — (1)

Avg power lost during ton process = $\frac{1}{T_0}\int_0^{t_{ON}} P(t)\,dt$

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$$= \frac{V_{cc} I_{cs}}{6} t_{ON} f \quad -(2)$$

Instantaneous power loss during toFF process

$$P(t) = V_{CE}(t) \cdot I_c(t)$$

$$= \left[\left(\frac{V_{cc}}{t_{OFF}}\right)t\right]\left[\left(-\frac{I_{cs}}{t_{OFF}} t + I_{cs}\right)\right]$$

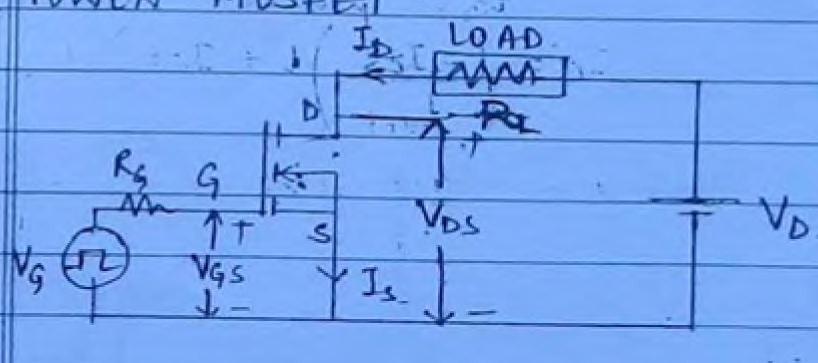
Energy lost during toFF process =

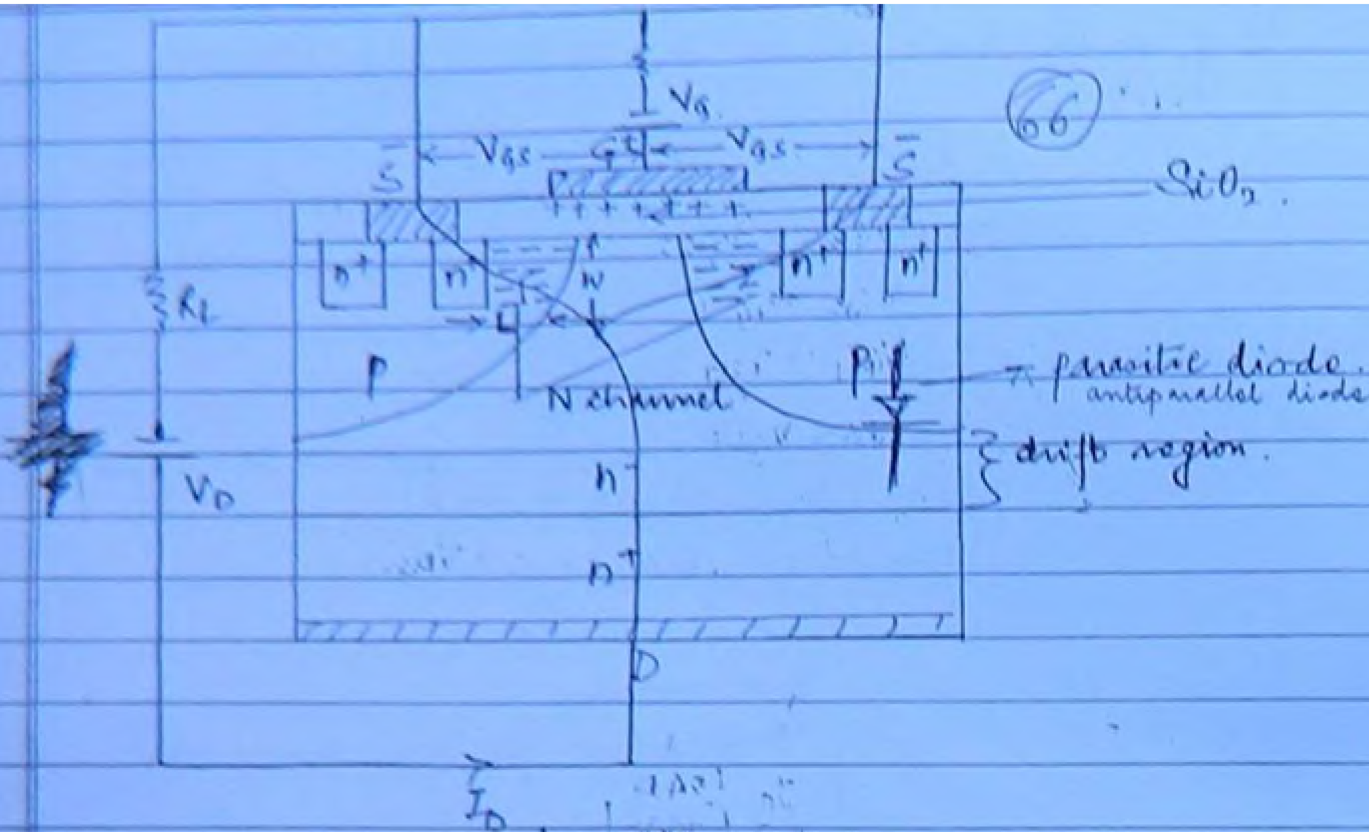
$$\int_0^{t_{ON}} P(t)\,dt = \frac{V_{cc} \cdot I_{cs}}{6} t_{OFF} \quad -(3)$$

Avg power lost during toFF process =

$$\frac{1}{T}\int_0^{t_{ON}} P(t) \cdot dt = \frac{V_{cc} I_{cs}}{6} t_{OFF} f \quad -(4)$$

2 POWER MOSFET -





Here the MOSFET starts conducting only after the formation of N-channel when possible gate signal is given. Here the conduction is only due to majority carriers, since there is no charge carriers f ∴ the reverse & recovery time delay is very much reduced. Hence Mosfet operates at high switching frequency.

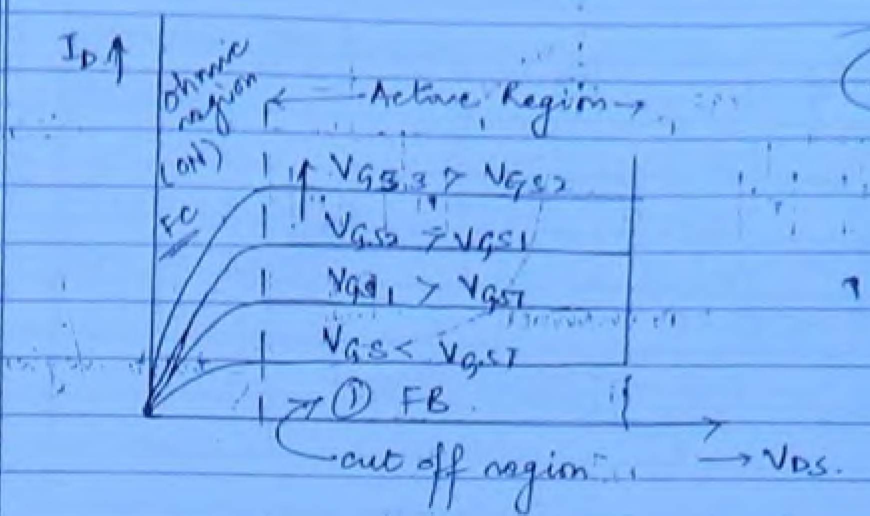
$V_{GS} \uparrow \quad \therefore W \uparrow \quad \therefore R_{CH} \downarrow \quad \therefore I_D \uparrow$

$V_{GS} \uparrow , \quad I_D \uparrow$

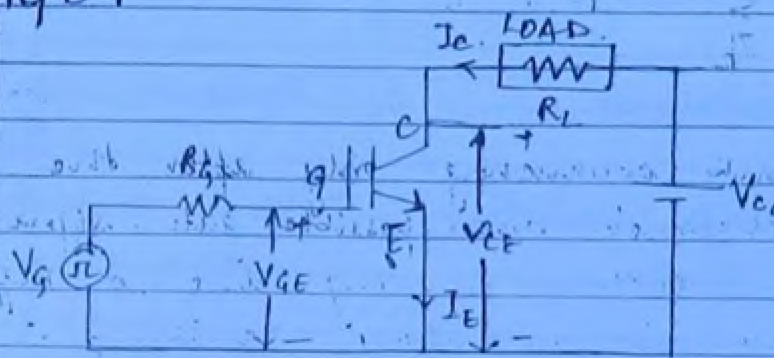
$I_B \uparrow \quad I_C \uparrow$

channel resistance $\downarrow$ $R_{CH} \propto \frac{L}{W \uparrow}$

$V_{GS} \uparrow \quad W \uparrow$



3 IGBT –

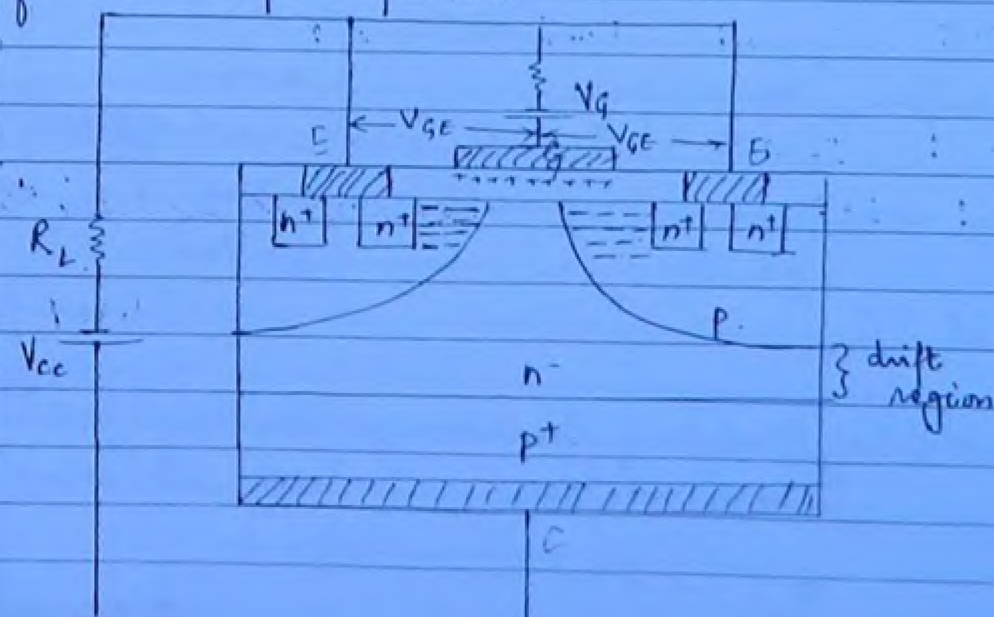


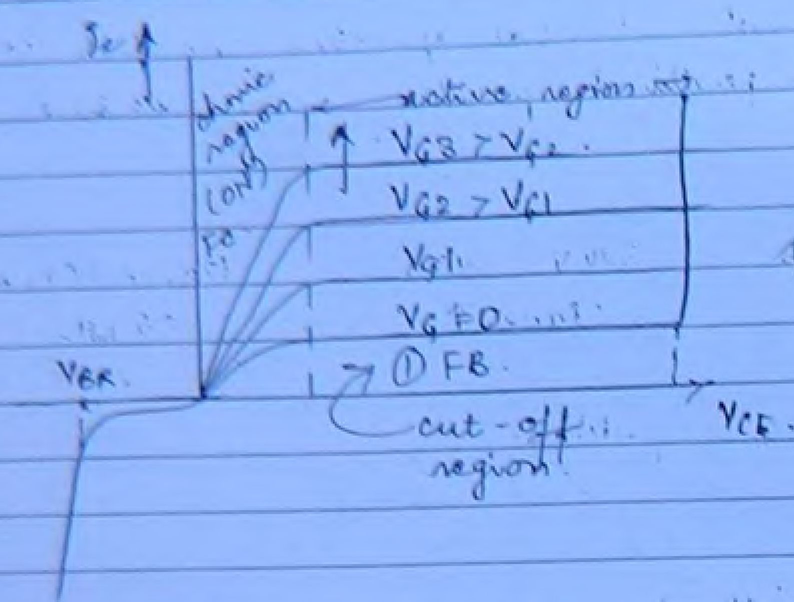
C, E → main terminals

G → control terminals

IGBT is a hybrid device that gives the advantages of both Mosfet & BJT.

IGBT switching is like MOSFET like conduction like BJT





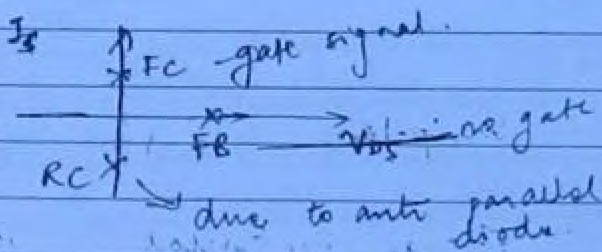
(6B)

POWER BJT	POWER MOSFET	IGBT
1. Bipolar device	Unipolar	Bipolar
2. Current controlled device	Voltage controlled device	Voltage controlled device
3. Low i/p impedance	high i/p impedance	high i/p impedance
4. on state v/g drop & conduction loss is less	more	less
5. Switching loss higher	~~losses~~ low	~~losses~~ low
DISADV 6. Negative Temp co-eff for Ron Temp↑ Ron↓ I↑ P↑	Positive Temp co-eff for Ron	Positive Temp co-eff for Ron
∴ Secondary Breakdown occurs	Secondary Breakdown will not occur	Secondary Breakdown will not occur

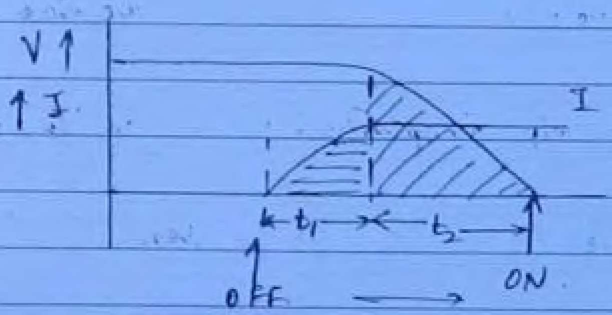
7.	BJTs are not advisable for parallel operation.	Parallel operation is possible	Parallel operation is possible. 69
Ratings 8.	1200V, 800A 10 - 20 KHz	500V, 140A 1MHz ↑	1200V, 500A, 50 kHz
App		SMPS.	

chapter 1 CWB.

1 (b)



3



turn-on process

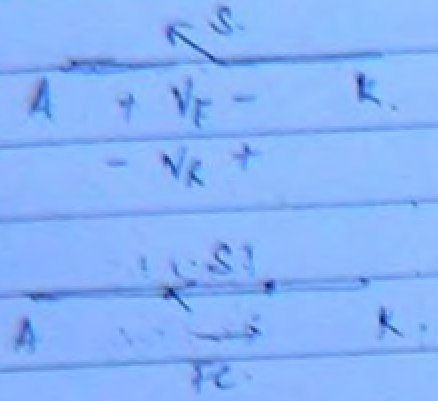
Energy lost during $t_1 = V\int_0^{t_1} I\,dt$

$= V \cdot \frac{1}{2} I t_1 = \frac{1}{2} V I t_1$

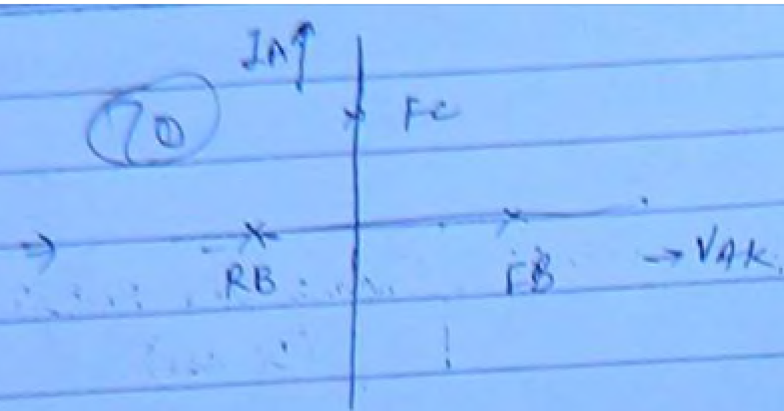
Energy lost during $t_2 = I\left(\int_0^{t_2} V\,dt\right) = I \cdot \frac{1}{2} V t_2 = \frac{1}{2} V I t_2$

Total $= \frac{1}{2} V I (t_1 + t_2)$ (a)

4

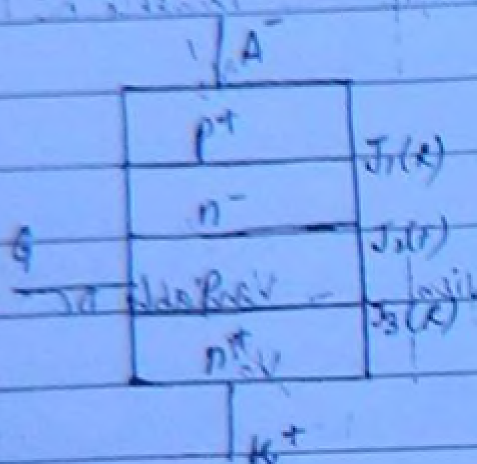


(D)



Ans (c)

17



P↑ temp ↑ I_c↑.

RECTIFIERS

Fixed AC → **Uncontrolled Rectifiers (Diodes)** → Fixed DC, V_0 = const

Full converter (Two quadrant operation)

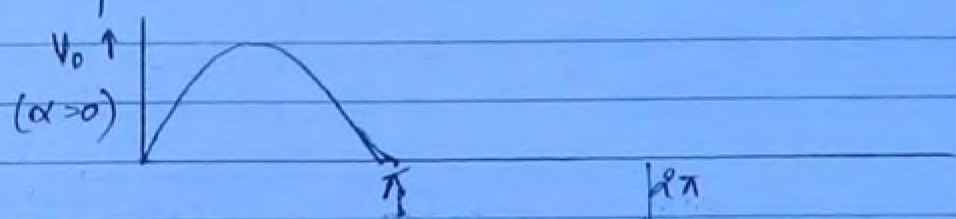
Fixed AC → **Fully controlled rectifier, Thyristor (α)** → Variable DC, $V_0 \nearrow$

Semi Converter (One quadrant)

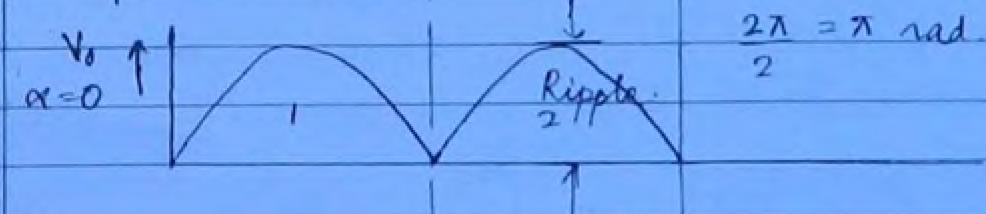
Fixed AC → **Half controlled rectifiers, Thyristors, Diodes** → Variable DC, $V_0 \nearrow$

Classification of Converters based on Pulse number (m)

1. One pulse converter



2. Two pulse converter

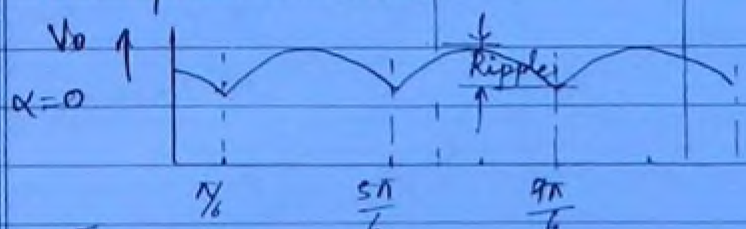


$\frac{2\pi}{2} = \pi$ rad

3. Three pulse converter

$\frac{2\pi}{3} = 120^\circ$ or $\frac{4\pi}{6}$

starts at $\pi/6$

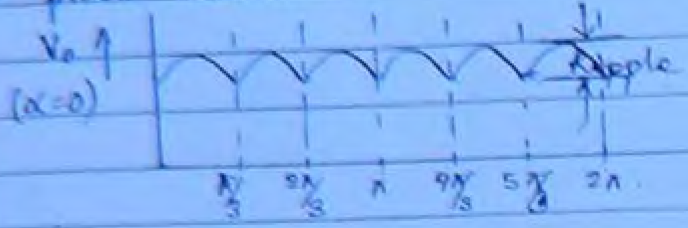


so $\frac{\pi}{6} + \frac{4\pi}{6} = \frac{5\pi}{6}$

Output ripple frequency = $f_o = mf_s$.
peak to peak

(Q2)

4. 6 pulse converter.



$\frac{2\pi}{6} = \frac{\pi}{3}$ $f_o = 6f_s$

$\boxed{m \uparrow \text{ ripple} \downarrow \therefore \text{ harmonics} \downarrow}$

That's why we classify converters based on pulse number.

Applications

Performance of a DC motor fed with converters -

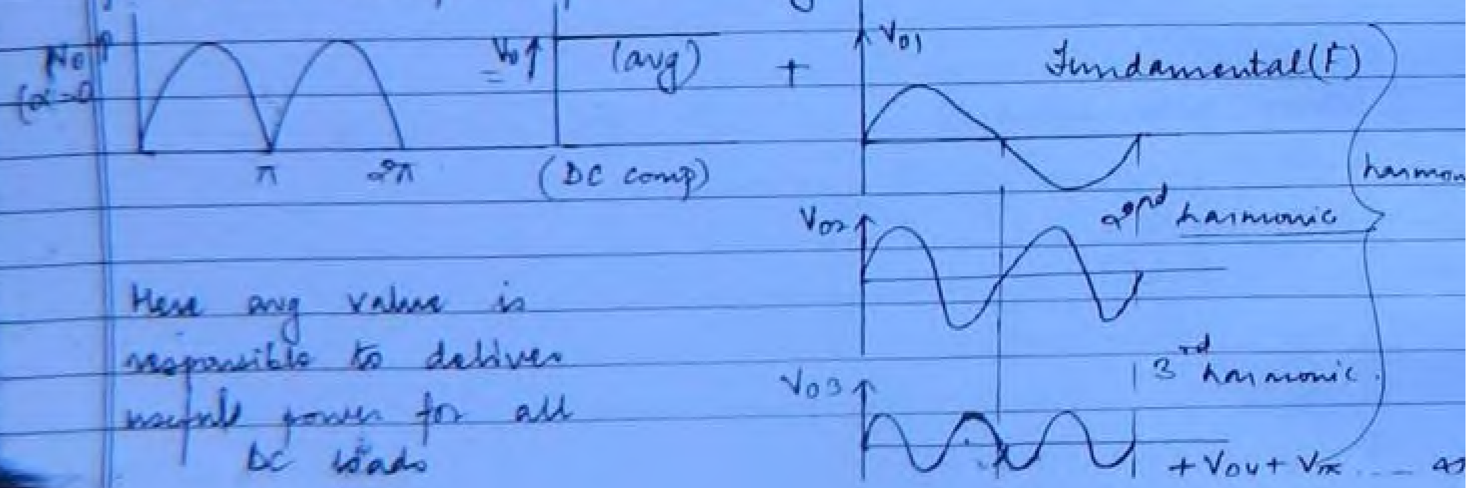
for R loads (eg heater) harmonics are also useful so take V_{OR} not V_0

1. Harmonics will overheat machine windings
∴ we cannot utilise the m/c to its full capacity.
∴ we must derate the m/c when fed with converters.

2. Harmonics produce pulsating torque in the motor & hence smooth rotation is not possible.

Harmonic Analysis on DC side of converter (V_0)

Output vtg waveform of a two pulse converter is distorted with harmonics. To find harmonic content present in waveform, fourier analysis is done.



Here avg value is responsible to deliver useful power for all DC loads

$+ V_{0u} + V_{0\pi} \ldots$

$$V_0 = a_0 + \sum_{n=1}^{\infty} (a_n \cos n\omega t + b_n \sin n\omega t)$$

$$V_0 = a_0 + \sum_{n=1}^{\infty} C_n \sin(n\omega t + \phi_n)$$

$\uparrow$

avg where $C_n = \sqrt{a_n^2 + b_n^2}$
(DC comp)

$$V_{0r} = \sqrt{V_0^2 + (V_{01})_{rms}^2 + (V_{02})_{rms}^2 + (V_{03})_{rms}^2 + \ldots}$$

$\uparrow$ RMS $\quad \uparrow$ avg

Squaring both sides

$$V_{0r}^2 = V_0^2 + (V_{01})_{rms}^2 + (V_{02})_{rms}^2 + (V_{03})_{rms}^2 + \ldots$$

$$\sqrt{V_{0r}^2 - V_0^2} = \sqrt{(V_{01})_{rms}^2 + (V_{02})_{rms}^2 + (V_{03})_{rms}^2 + \ldots}$$

RMS value of harmonics

$\boxed{\text{VRF}}$ = Voltage Ripple Factor

→ Its the measure of harmonics on DC side of converter.

$$VRF = \frac{\sqrt{V_{0r}^2 - V_0^2}}{V_0} = \sqrt{\left(\frac{V_{0r}}{V_0}\right)^2 - 1} = \sqrt{FF^2 - 1}$$

FF = form factor $= \left(\frac{rms}{avg}\right)$

$$\boxed{VRF = \sqrt{FF^2 - 1}}$$

Quality of perfect DC -

1. $V_0 = V_{0r}$ (74)
avg rms

2. $FF = 1$

3. $VRF = 0$
∴ no harmonics. no ripple.

without harmonics FF = 1
with harmonics FF > 1
FF ↓ approaching unity.
⇒ smoothness of waveform is improved towards DC.

• Form Factor → gives the information of shape of the waveform.

Harmonic analysis on AC side of converter -

x. Consider an inverter. Output v/g waveform of inverter is not perfect AC. Its distorted with harmonics

x. For all AC loads, fundamental is responsible to deliver useful power

$$V_{0r} = \sqrt{V_0^2 + (V_{01})_{rms}^2 + (V_{02})_{rms}^2 + (V_{03})_{rms}^2 + \cdots}$$

↑ RMS ↑ avg

$$\sqrt{V_{0r}^2 - (V_{01})_{rms}^2} = \sqrt{V_0^2 + (V_{02})_{rms}^2 + \cdots}$$

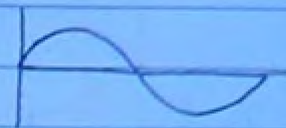
[THD]

Total Harmonic Distortion → Method of harmonics on AC side of the converter.

$$THD = \frac{\sqrt{V_{0r}^2 - V_{01}^2}}{V_{01}}$$

Distortion Factor $(g) \rightarrow \quad g = \dfrac{(V_{01})_{rms}}{V_{or}}$

w/o harmonics $g = 1$



with harmonics $g < 1$ $[\because (V_{01})_{rms} < V_{or}]$

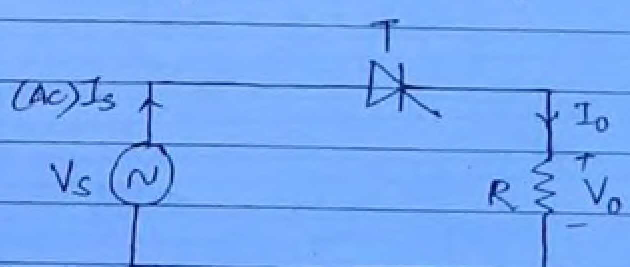
As $g \uparrow$ & approaches unity, smoothness of waveform is improved towards AC.

Qualities of perfect AC

1. $V_{or} = (V_{01})_{rms}$
2. $g = 1$
3. $THD = \left(\dfrac{1}{g^2} - 1\right)^{1/2}$

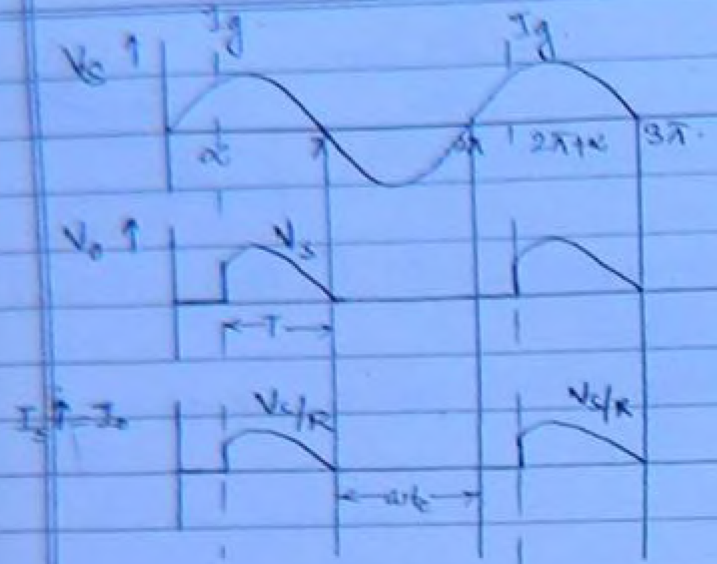
for perfect AC, THD $= 0$
$\Rightarrow$ no harmonics

1 ϕ Half Wave Rectifier (one pulse converter)



$I_g(\alpha, 2\pi + \alpha, \ldots)$

(96)



$$\omega t_c = \pi$$

$$t_c = \frac{\pi}{\omega}$$

$$V_0 = \frac{1}{2\pi}\int_{\alpha}^{\pi} V_m \sin \omega t \, d(\omega t)$$

$$V_0 = \frac{V_m}{2\pi}(1+\cos\alpha)$$

$$(I_s)_{avg} = I_0 = \frac{V_0}{R} = \frac{V_m}{2\pi R}(1+\cos\alpha)$$

DC comp

Disadvantages -

- Source current contains DC component & saturates the supply transformer core.

$$V_{rms} = V_{0r} = \left\{\frac{1}{2\pi}\int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \, d(\omega t)\right\}^{1/2}$$

$$\Rightarrow V_{0r} = \left\{\frac{1}{2\pi}\int_{\alpha}^{\pi} V_m^2\left(\frac{1-\cos 2\omega t}{2}\right) d(\omega t)\right\}^{1/2}$$

$$\Rightarrow V_{0r} = \frac{V_m}{2\sqrt{\pi}}\left\{(\omega t)_{\alpha}^{\pi} - \frac{1}{2}(\sin 2\omega t)_{\alpha}^{\pi}\right\}^{1/2}$$

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

$$V_{or} = \frac{V_m}{2\sqrt{\pi}}\left[(\pi-\alpha) + \frac{1}{2}\sin 2\alpha\right]^{1/2}$$

$$P_{in} = V_{sr} I_{sr} \cos\phi$$
$$P_0 = V_{or} I_{or}$$
$$P_{in} = P_0$$

$$\cos\phi = \frac{V_{or} I_{or}}{V_{sr} I_{sr}}$$

$$\boxed{PF = \frac{V_{or}}{V_{sr}}}$$

$$PF = \frac{1}{\sqrt{2\pi}}\left[(\pi-\alpha) + \frac{1}{2}\sin 2\alpha\right]^{1/2}$$

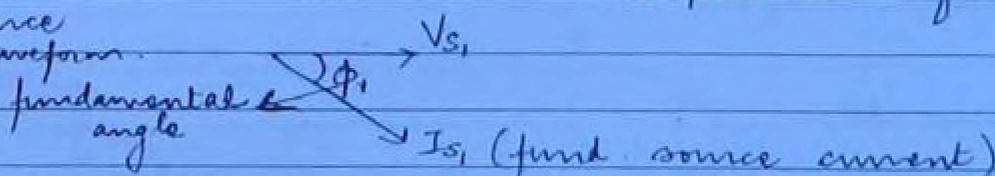
Power factor depends on –

1. Firing angle – As $\alpha \uparrow$ $PF \downarrow$
2. PF also depends on the harmonics ie it depends on the shape of the source current waveform

$$\boxed{PF = g \times FDF}$$

g – distortion factor for source current waveform

$\boxed{FDF}$ = fundamental displacement factor



$$FDF = \cos\phi_1$$

$$P_{in} = V_{s1} I_{s1} (PF) = V_{s1} I_{sh} (FDF)$$

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$$PF = \frac{I_{s1}}{I_{s2}} (FDF)$$

$$PF = g(FDF) \qquad g = \frac{I_{s1}}{I_{s2}}$$

3. PF also depends on load parameters.

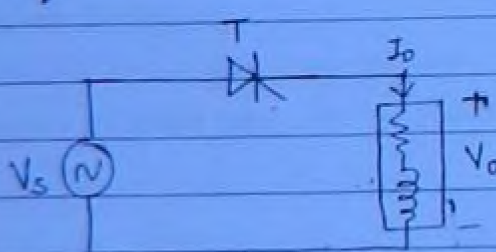
Drawback of PE converters -

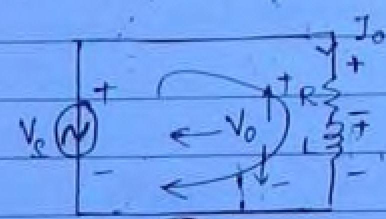
1. The converters inject harmonics into the supply system (or utility system) & reduce the quality of supply line. To rectify this problem, we must use AC filters on AC side of converters.

2. The converter draws reactive power from supply line for its operation. We must compensate the reactive power required for converter operation by using reactive power ~~supply~~ source on the AC side of the converter.

(II) 1ϕ Half Wave Rectifier $\rightarrow$ RL load.

① α to π $\quad T \rightarrow ON$





P+

$P_{source} \longrightarrow P_{load}$

$$\left(I^2 R + \frac{1}{2} L I^2 \right)$$

L stores energy.

$$V_m \sin \omega t = Ri + L\frac{di}{dt}$$

$$I_0 = I_{steady} + I_{transient}$$ (79)

$$I_{steady} = \frac{V_m}{|Z|}\sin(\omega t - \phi)$$

where $|Z| = \sqrt{R^2 + (\omega L)^2}$

$\phi = \tan^{-1}\frac{\omega L}{R}$

$$I_{transient} = Ke^{-t/\tau}$$

$I_0 = I_{steady} + I_{trans.}$

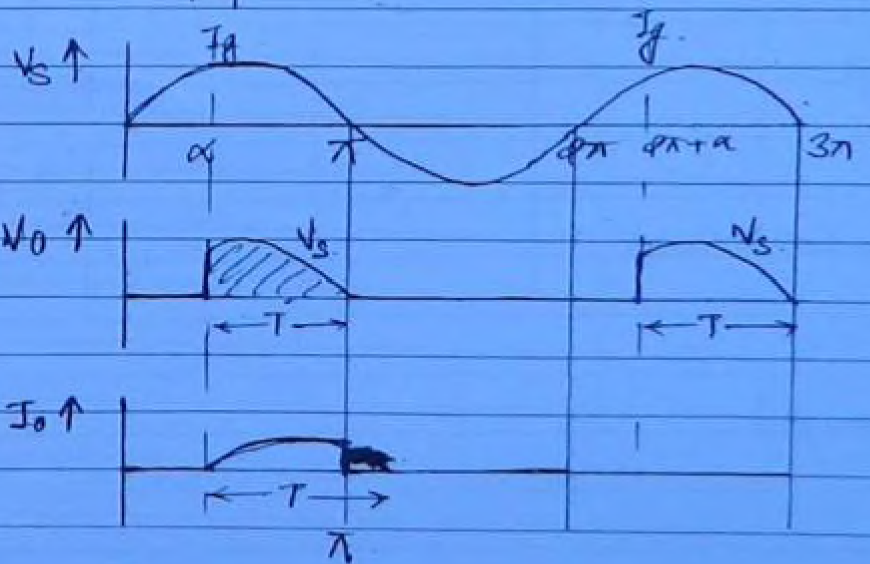
$$I_0 = \frac{V_m}{|Z|}\sin(\omega t - \phi) + Ke^{-t/\tau}$$

At $\omega t = \alpha$ $\quad I_0 = 0$

$t = \frac{\alpha}{\omega}$ $\quad \tau = \frac{L}{R}$

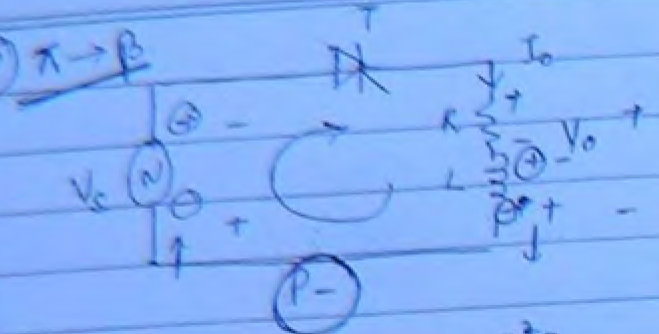
$$0 = \frac{V_m}{|Z|}\sin(\omega t - \phi) + Ke^{-R\alpha/\omega L}$$

$$K = \frac{-V_m}{|Z|}\sin(\alpha - \phi)\, e^{R\alpha/\omega L}$$



(2) $\pi \rightarrow \beta$

$I_A = 0 < I_H$

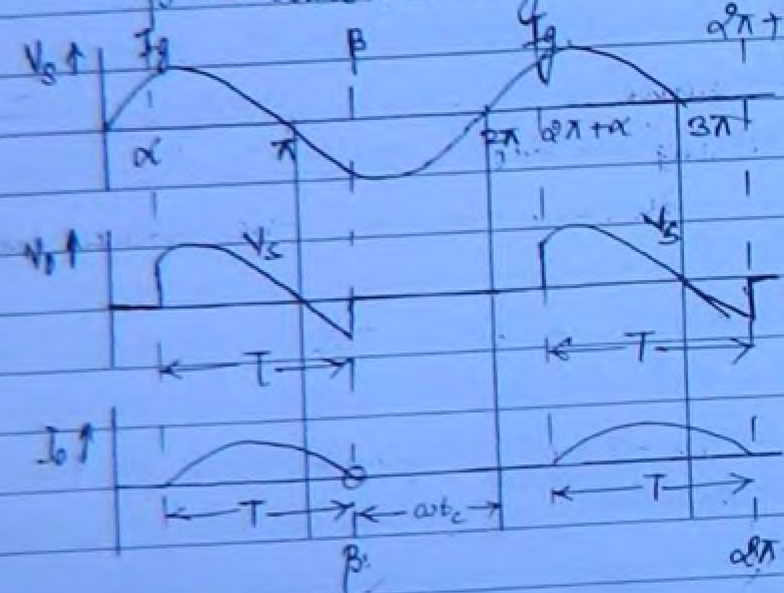


(80)

$\frac{1}{2} L I^2 \rightarrow$ source $+ I^2 R$

⇒ The inductance energy maintains conduction of thyristor even in the –ve cycle until it releases the complete energy at $\omega t = \beta$.

β = extinction angle.



At $\omega t = \pi$ reverse v/g is applied across SCR but it'L still conduct. SCR doesnt get turned OFF until $I_A = 0$.

L : does not accept sudden change in i

I_A continues to flow till β → the point where L looses its energy completely.

$$\omega t_c = 2\pi - \beta$$

$$t_c = \frac{2\pi - \beta}{\omega}$$

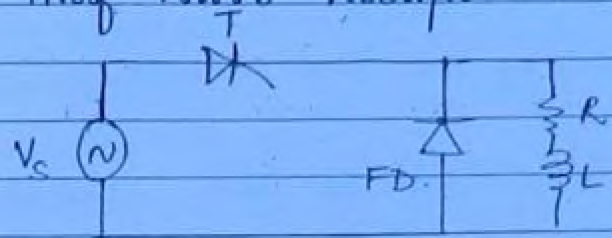
$$V_0 = \frac{1}{2\pi} \int_{\alpha}^{\beta} V_m \sin\omega t \, d(\omega t)$$

$$V_0 = \frac{V_m}{2\pi} [\cos\alpha - \cos\beta]$$

$$V_{or} = \left\{ \frac{1}{2\pi} \int_{\alpha}^{\beta} V_m^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

$$V_{or} = \frac{V_m}{2\sqrt{\pi}} \left[(\beta - \alpha) + \frac{1}{2} (\sin 2\alpha - \sin 2\beta) \right]$$ (81)

(III) 1-ϕ Half Wave Rectifier → RL load with FD.



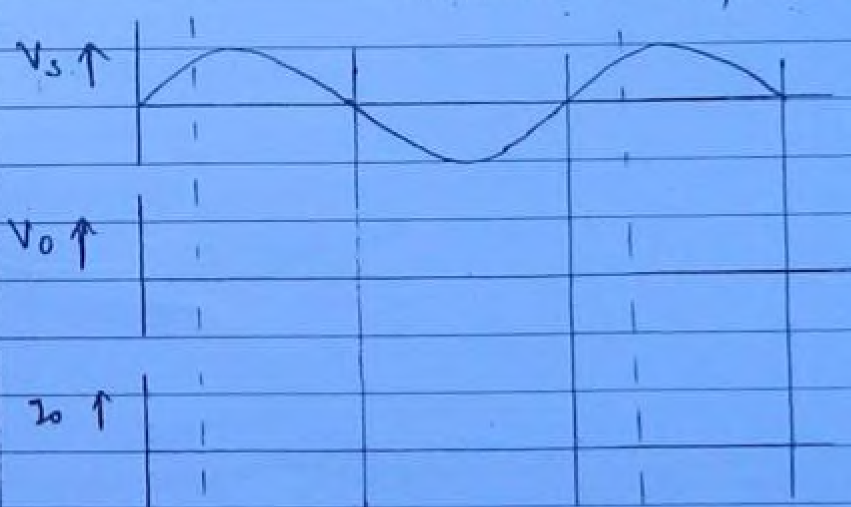
(1) mode is same.

(2) ~~FD~~ freewheeling mode.

During free wheeling action, -ve spikes are removed.

L releases energy through I^2R → takes more time $\beta \uparrow$

smoothness improves $g \uparrow$ PF $\uparrow$



Advantages of FD -

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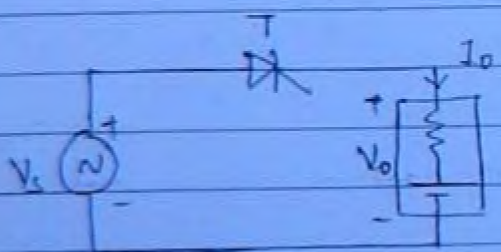
1. PF is improved
2. -ve spikes in load vlg is improved & this ↑ avg vlg.
3. Smoothness of output current waveform is improved as $\beta \uparrow$.
4. ∴ the overall perfomance of the converter is improved with FD.
5. There will be freewheeling action in semiconverters ∴ PF is better in semiconverter as compared to full converters.

∴ performance of semiconverter is superior to full converter.

12-12

IV 1-ϕ Half Wave Rectifier - charging a battery (RE load)



At $\omega t = \theta_1$

$$V_s = E$$

$$V_m \sin\theta_1 = E$$

$$\alpha_{min} = \theta_1 = \sin^{-1}\left(\frac{E}{V_m}\right)$$

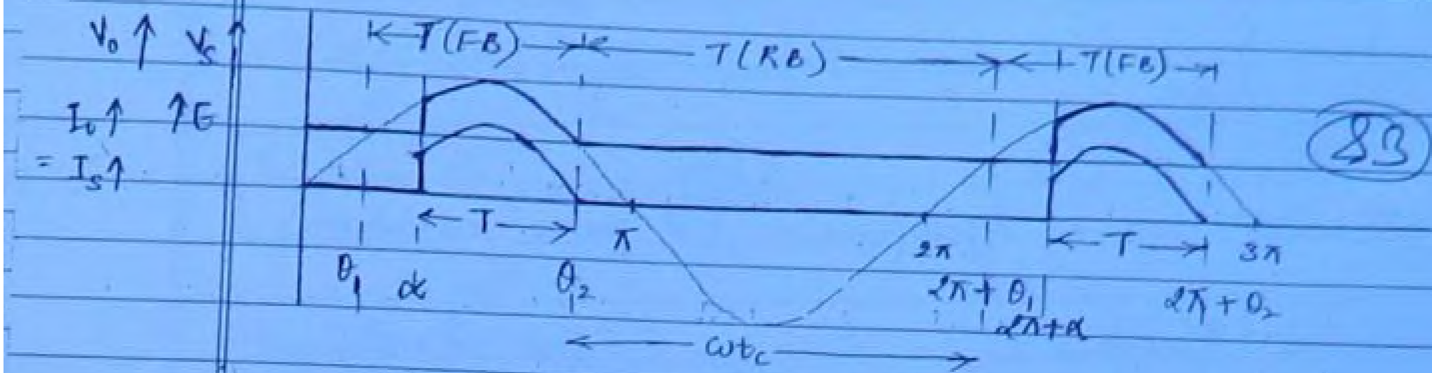
$$\alpha_{max} = \theta_2 = \pi - \theta_1$$

$$\theta_1 \le \alpha \le \theta_2$$

$$I_g(\alpha, 2\pi+\alpha, \ldots)$$

$T \rightarrow$ ON $\Rightarrow V_0 = V_s \Rightarrow V_s = I_0 R + E$

$$I_0 = \frac{V_s - E}{R} = \frac{V_m \sin\omega t - E}{R}$$



$$\omega t_c = (2\pi+\theta_1) - \theta_2$$
$$= (2\pi+\theta_1) - (\pi - \theta_1)$$

$$\boxed{t_c = \frac{\pi + 2\theta_1}{\omega}}$$

$$\boxed{PIV = V_m + E}$$

$$V_0 = \frac{1}{2\pi}\left[\int_{\alpha}^{\theta_2} V_m \sin\omega t \, d(\omega t) + \int_{\theta_2}^{2\pi+\alpha} E \, d(\omega t)\right]$$

$$V_0 = \frac{1}{2\pi}\left[V_m(\cos\alpha - \cos\theta_2) + E\underbrace{(2\pi+\alpha-\theta_2)}_{\text{radians}}\right]$$

** Avg charging current of the battery

$$I_0 = \frac{1}{2\pi}\int_{\alpha}^{\theta_2} \frac{V_m \sin\omega t - E}{R} \, d(\omega t)$$

$$I_0 = \frac{1}{2\pi R}\left[V_m(\cos\alpha - \cos\theta_2) - E\underbrace{(\theta_2 - \alpha)}_{\text{radians}}\right]$$

$$P_{in} = V_{sr} I_{sr} (PF)$$
$$P_0 = I_{or}^2 R + E I_0$$
$$PF = \frac{I_{or}^2 R + E I_0}{V_{sr} \cdot I_{sr}} \qquad PF = \frac{I_{or}^2 R + E I_0}{V_{sr} \cdot I_{sr}}$$

$$I_{or} = \left\{ \frac{1}{2\pi} \int_{\alpha}^{\theta_2} \left(\frac{V_m \sin \omega t - E}{R} \right)^2 d(\omega t) \right\}^{1/2}$$

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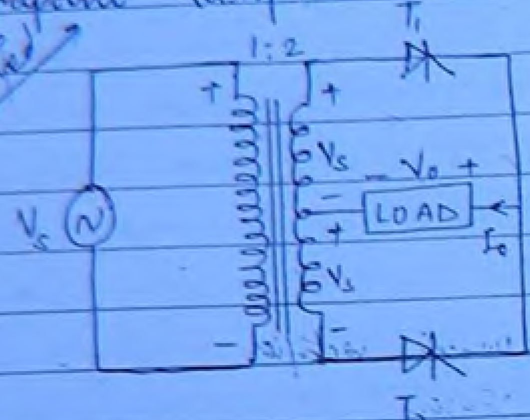
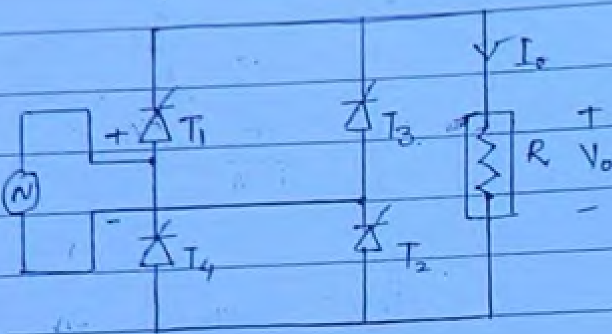
1-ϕ Full Wave Rectifiers – (2 pulse converter)

Midpoint Rectifiers –

where transformers are required



Bridge Rectifiers –



$T_1 \rightarrow ON$
$V_0 = V_s$
$I_0 = V_s/R$

$+ \; T_1 \,(FB) \quad I_g (\alpha, 2\pi+\alpha \ldots)$

$- \; T_2 \,(FB) \quad I_g (\pi+\alpha, 3\pi+\alpha \ldots)$

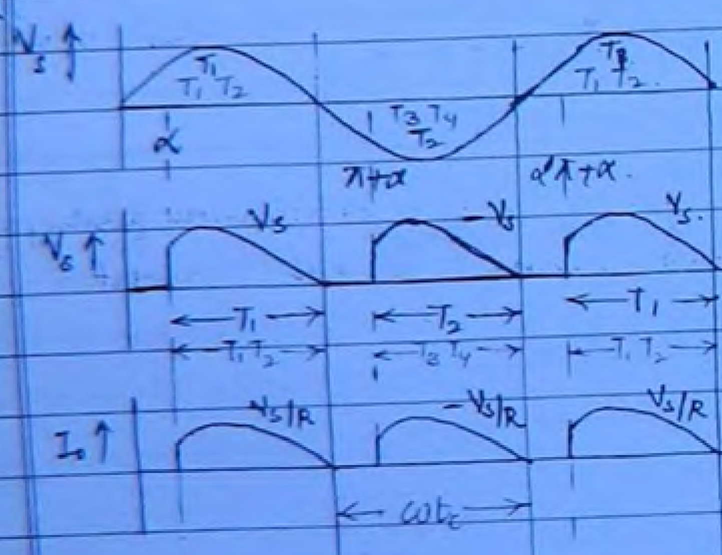
$T_2 \rightarrow ON$
$V_0 = -V_s$
$I_0 = -\frac{V_s}{R}$

$+ \; T_1 T_2 \,(FB) \quad I_g (\alpha, 2\pi+\alpha \ldots)$

$- \; T_3 T_4 \,(FB) \quad I_g (\pi+\alpha, 3\pi+\alpha \ldots)$

$T_1 T_2 \rightarrow ON$
$V_0 = V_s$
$I_0 = V_s/R$

$T_3 T_4 \rightarrow ON$
$V_0 = -V_s$
$I_0 = -V_s/R$



$\omega t_c = \pi$

$$\boxed{t_c = \frac{\pi}{\omega} \text{ secs}}$$

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi} V_m \sin \omega t \, d(\omega t)$$

$$\boxed{V_0 = \frac{V_m}{\pi}(1+\cos\alpha)}$$

$$V_{or} = \left\{\frac{1}{\pi}\int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \, d(\omega t)\right\}^{1/2}$$

$$\boxed{V_{or} = \frac{V_m}{\sqrt{2\pi}}\left[(\pi-\alpha) + \frac{1}{2}\sin 2\alpha\right]^{1/2}}$$

V↑
t↑

$PIV = 2V_m \longrightarrow$ In mid-point rectifier

adv $\rightarrow$ $PIV = V_m \longrightarrow$ bridge rectifier

Advantages of Bridge Rectifier -

1. PIV of thyristor in bridge rectifier is half that of mid point rectifier

2. If same thyristors with same specifications (same vtg & current ratings) are same in both converters then power handled by bridge rectifier is double that of mid-point rectifier

1-φ Full Wave Rectifiers (R-L Load) (Full converter)

Discontinuous Conduction

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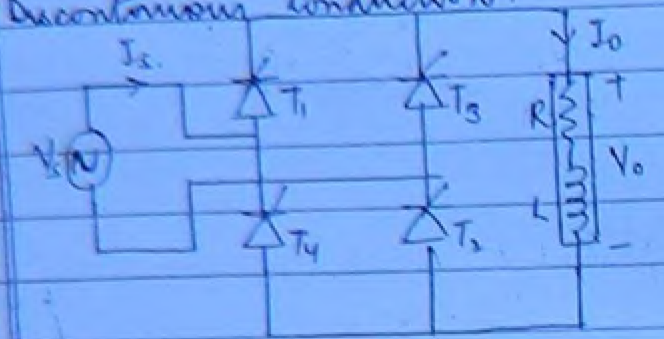
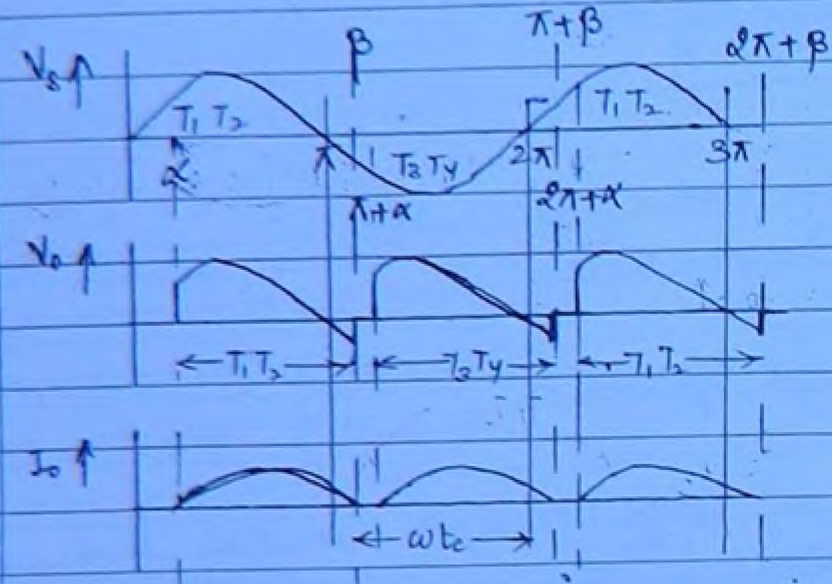




$$\omega t_c = 2\pi - \beta$$

$$\boxed{t_c = \frac{2\pi - \beta}{\omega}}$$

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\beta} V_m \sin \omega t \, d(\omega t)$$

$$V_0 = \frac{V_m}{\pi}(\cos\alpha - \cos\beta)$$

$$V_{or} = \left\{ \frac{1}{\pi}\int_{\alpha}^{\beta} V_m^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

$$V_{or} = \frac{V_m}{\sqrt{2\pi}}\left[(\beta - \alpha) + \frac{1}{2}(\sin 2\alpha - \sin 2\beta) \right]^{1/2}$$

Reasons for Discontinuous conduction -

1. $L \downarrow$ [Time constant $T \downarrow = \frac{\downarrow L}{R}$] or $R \uparrow$
   $T \downarrow \therefore \beta \downarrow$ $[\beta < (\pi + \alpha)]$

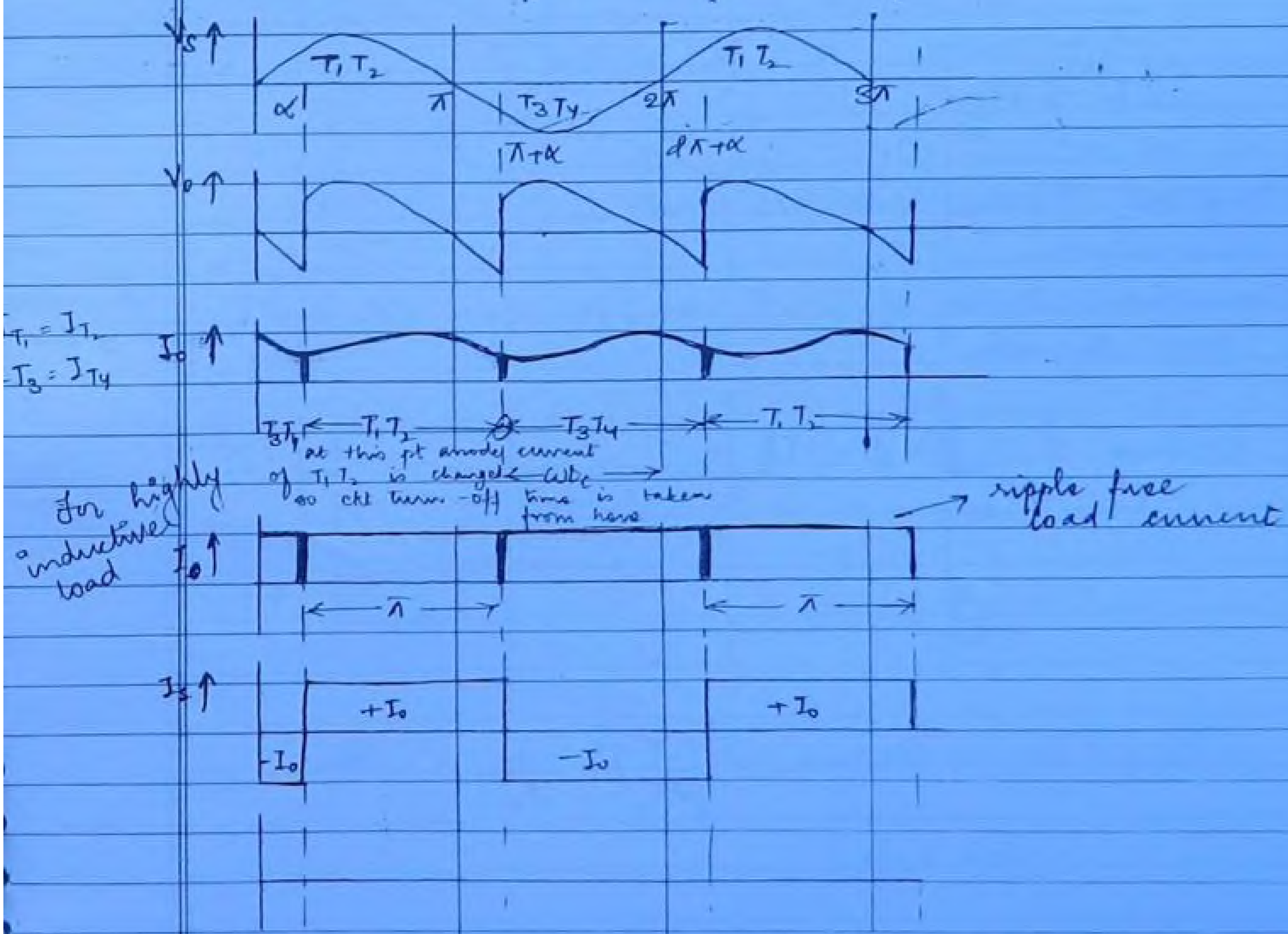
2. $\alpha \uparrow$ (High value of firing angle) — less energy is stored in inductor.

3. $I_0 \downarrow$ $\downarrow E = \frac{1}{2} L i^2 \downarrow$ not sufficient energy.

RL Load Continuous Conduction

1. $L \uparrow$ $\left[\uparrow T = \uparrow \frac{L}{R}\right]$ $\uparrow T = \uparrow \beta$ $[\beta \geq (\pi + \alpha)]$

2. $\alpha \downarrow$ 3. $I_0 \uparrow$



$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi+\alpha} V_m \sin\omega t \, d(\omega t)$$

$$\boxed{V_0 = \frac{2V_m}{\pi}\cos\alpha}$$

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$$V_{or} = \left\{ \frac{1}{\pi}\int_{\alpha}^{\pi+\alpha} V_m^2 \sin^2\omega t \, d(\omega t) \right\}^{1/2}$$

$$\boxed{V_{or} = \frac{V_m}{\sqrt{2}}}$$

$\longrightarrow$ squaring V_S or V_0 waveform gives same waveform so their rms values are equal.

~~Hence~~ Rms value of $V_s = \frac{V_m}{\sqrt{2}}$. so is that of V_0

$$\omega t_c = 2\pi - (\pi + \alpha)$$

$$\boxed{t_c = \frac{\pi - \alpha}{\omega}}$$

$T_1 T_2 \to ON$ $\quad I_s = I_0$

$T_3 T_4 \to ON$ $\quad I_s = -I_0$

Assume highly inductive load -

Conduction Angle of each thyristor $= \pi$ rad [for every 2π rad]

$$(I_0)_{avg} = \frac{1}{2\pi}\int_{\alpha}^{\pi+\alpha} I_0 \, d(\omega t) = \frac{I_0}{2}$$

$$= I_0\left(\frac{\pi}{2\pi}\right)$$

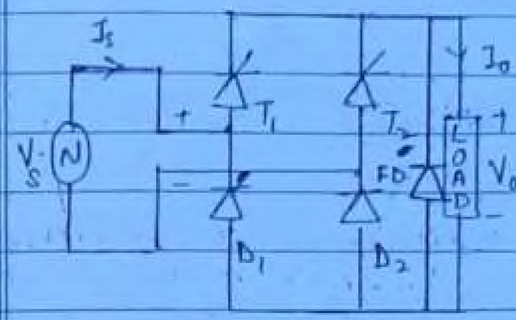
$$(I_T)_{avg} = \frac{I_0}{2} \qquad (I_T)_{rms} = I_0\left(\frac{\pi}{2\pi}\right)^{1/2} = \frac{I_0}{\sqrt{2}}$$

So

$I_{...} = I_0$

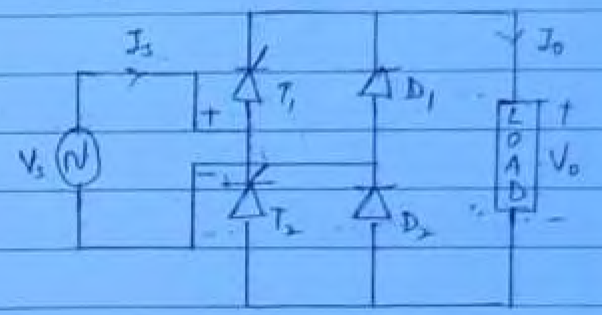
1-ϕ Half Controlled Rectifier - (Semi converter)

Symmetrical Connection -

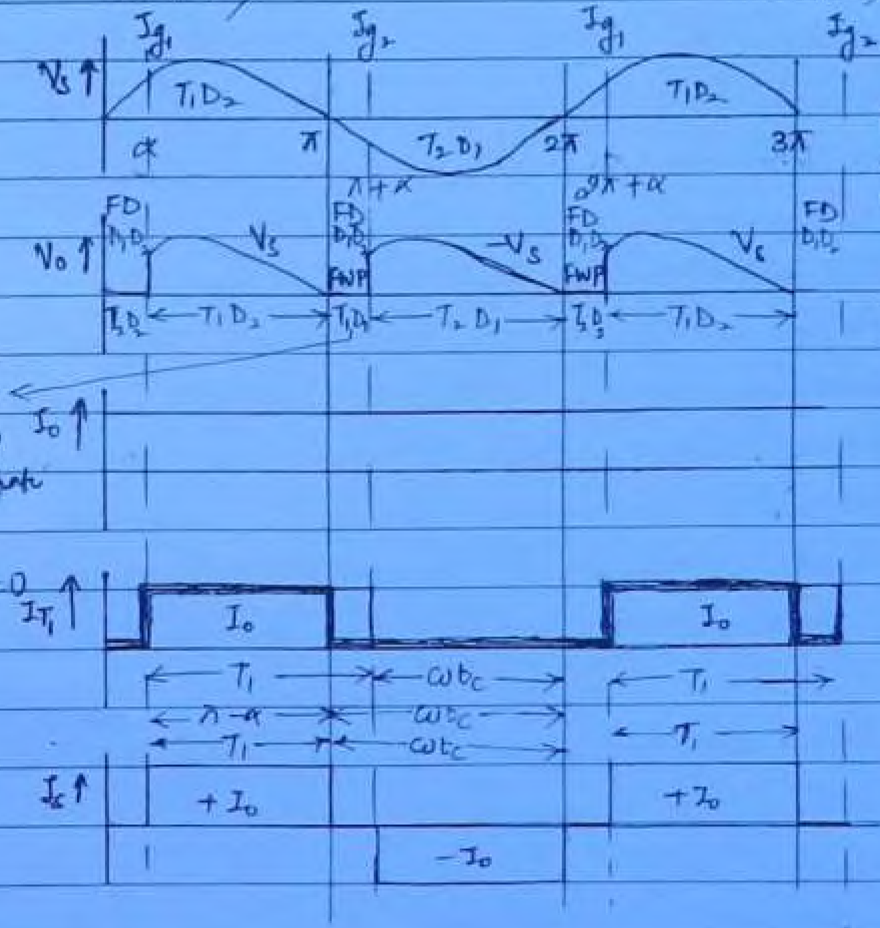


$+$ T_1 D_2 (F)
$-$ T_2 D_1 (F)

Asymmetrical Connection -



$+$ T_1 D_2 (F)
$-$ T_2 D_1 (F)



Due to Freewheeling diode -ve spikes are removed. thus the V_0 becomes same as that of R load in full converter

FWP
Free wheeling period
(that's why no need of separate FD)
↓
load vtg = 0

For resistive load waveform remains same for full converter & semi converter

Assume highly inductive load -

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π to $\pi+\alpha$ $\quad T_1 D_1 \to ON$

FWP $\quad V_0 = 0$

FWP $\quad I_S = 0$

$T_1 D_2 \to ON \quad I_S = I_0$

$T_2 D_1 \to ON \quad I_S = -I_0$

$\omega t_c = 2\pi - (\pi+\alpha)$ $\qquad\qquad$ $\omega t_c = 2\pi - \pi$

$$\boxed{t_c = \frac{\pi-\alpha}{\omega}} \qquad\qquad \boxed{t_c = \frac{\pi}{\omega}}$$

* In symmetrical connection theres a possibility of S.C across the supply when the incoming thyristor starts conducting before the outgoing thyristor stops conducting. Here the problem is severe, because before the incoming thyristor starts conducting, the freewheeling action is through outgoing thyristor because of which SC period is more.

* To rectify this problem we must use a separate FD.

Symmetrical connection with FD.

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi} V_m \sin\omega t \, d(\omega t)$$

$$V_0 = \frac{V_m}{\pi}[1+\cos\alpha]$$

$$V_{or} = \left\{ \frac{1}{\pi} \int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

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$$V_{or} = \frac{V_m}{\sqrt{2}\pi} \left[(\pi - \alpha) + \frac{1}{2} \sin 2\alpha \right]^{1/2}$$

Assume highly inductive load -

Conduction angle each thyristor $= (\pi - \alpha)$ rad [for every 2π rad]

$$= \frac{\pi - \alpha}{2\pi}$$

Conduction angle of FD $= \alpha$ rad [for every π rad]

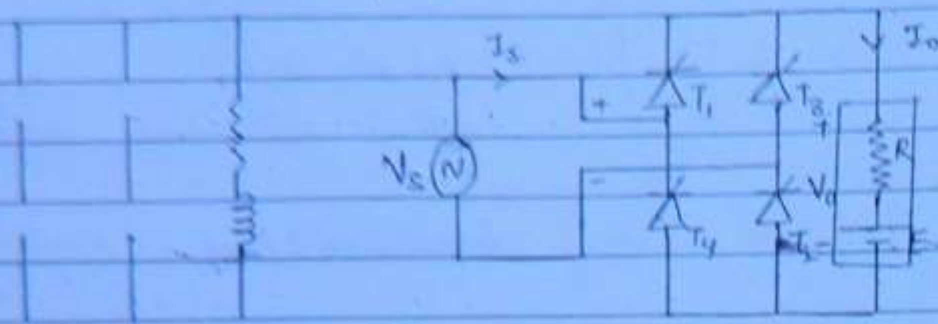
$$= \frac{\alpha}{\pi}$$

$$(I_T)_{avg} = I_0 \left(\frac{\pi - \alpha}{2\pi} \right) \qquad (I_{FD})_{avg} = I_0 \left(\frac{\alpha}{\pi} \right)$$

$$(I_T)_{rms} = I_0 \left(\frac{\pi - \alpha}{2\pi} \right)^{1/2} \qquad (I_{FD})_{rms} = I_0 \left(\frac{\alpha}{\pi} \right)^{1/2}$$

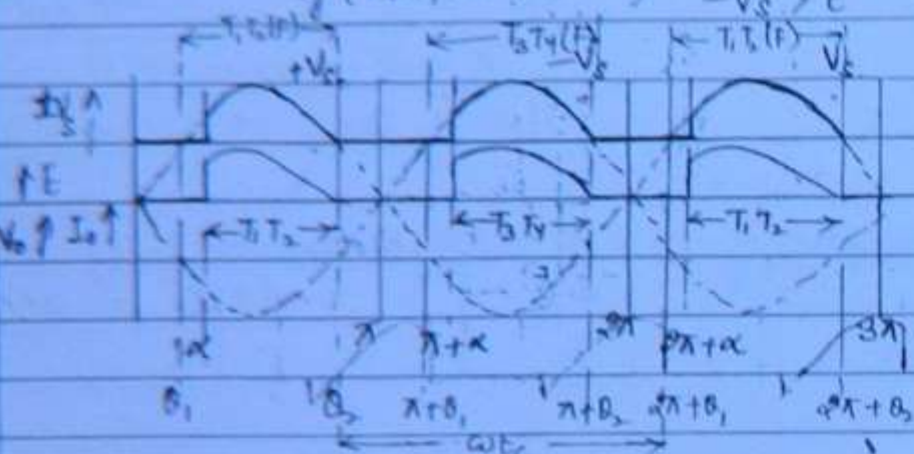
$$\boxed{I_{or} = I_0 \left(\frac{\pi - \alpha}{\pi} \right)^{1/2}}$$

1ϕ Full Converter - Charging a Battery



$I_g\ (\alpha,\ 2\pi+\alpha\ \ldots)$ $\quad + V_s > E \quad T_1 T_2$ (F.B)

$I_g\ (\pi+\alpha,\ 3\pi+\alpha\ \ldots)$ $\quad - V_s > E \quad T_3 T_4$ (F.B)



$T_1 T_2 \to ON$ $\qquad\qquad \theta_1 = \sin^{-1}\dfrac{E}{V_m}$

$V_0 = V_s$

$I_0 = \dfrac{V_s - E}{R}$ $\qquad\qquad \theta_2 = \pi - \theta_1$

$$\boxed{I_0 = \frac{V_m \sin\omega t - E}{R}}$$

$$\omega t_c = (2\pi + \theta_1) - \theta_2$$
$$= (2\pi + \theta_1) - (\pi - \theta_1)$$

$$\boxed{t_c = \frac{\pi + 2\theta_1}{\omega}}$$

$$V_0 = \frac{1}{\pi}\left[\int_{\alpha}^{\theta_2} V_m \sin\omega t\, d(\omega t) + \int_{\theta_2}^{\pi+\alpha} E\, d(\omega t)\right]$$

$$V_0 = \frac{1}{\pi}\left[V_m(\cos\alpha - \cos\theta_2) + E(\pi + \alpha - \theta_2)\right]$$

$$I_0 = \frac{1}{\pi} \int_{\alpha}^{\theta_2} \left[\left(\frac{V_m \sin \omega t - E}{R} \right) d(\omega t) \right]$$

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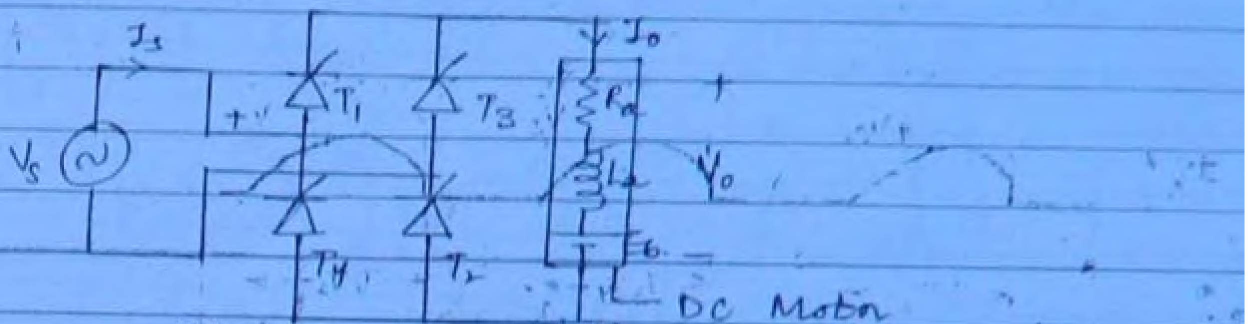
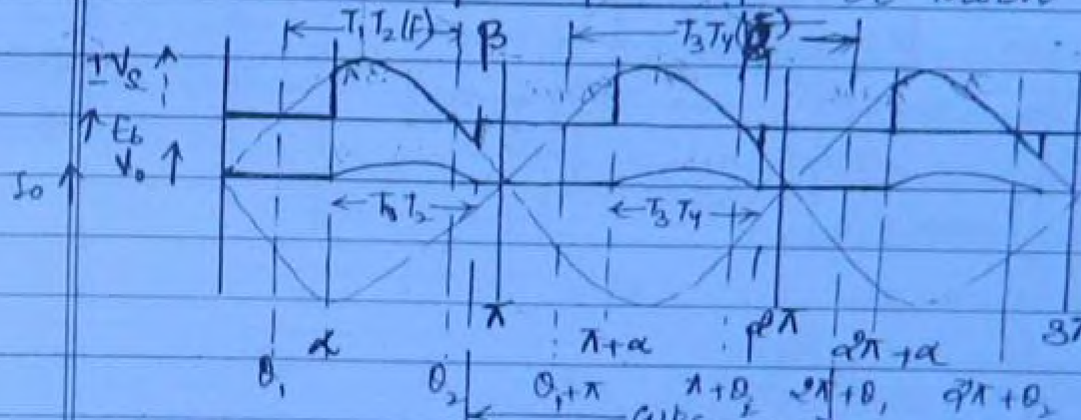
$$I_0 = \frac{1}{\pi R} \left[V_m (\cos\alpha - \cos\theta_2) - E(\theta_2 - \alpha) \right]$$

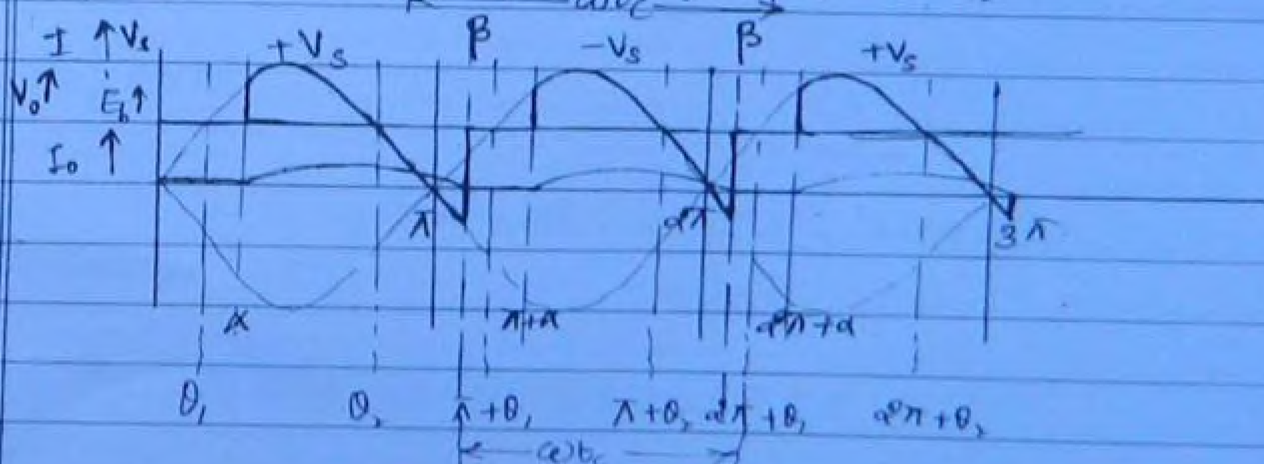
1 φ Full Converter – DC Motor (RLE Load)



$\beta < \pi$



$\beta > \pi$



$\omega t_2 = (2\pi + \theta_1) - \beta$.

$\downarrow t_c = \dfrac{2\pi + \theta_1 - \beta}{\omega} \uparrow$

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For continuous conduction waveform remains same for RL & RLE load.

$$V_0 = \frac{1}{\pi}\left[\int_{\alpha}^{\beta} V_m \sin\omega t \, d(\omega t) + \int_{\beta}^{\pi+\alpha} E_b \, d(\omega t)\right]$$

$$V_0 = \frac{1}{\pi}\left[V_m(\cos\alpha - \cos\beta) + E_b(\pi + \alpha - \beta)\right]$$

For continuous conduction:

$V_0 = \dfrac{2V_m}{\pi}\cos\alpha \rightarrow$ RL, RLE.

$I_0 = \dfrac{V_0}{R} \longrightarrow$ RL

$I_0 = \dfrac{V_0 - E_b}{R_a} \longrightarrow$ RLE

1 ϕ Semiconverter – DC Motor (RLE load)

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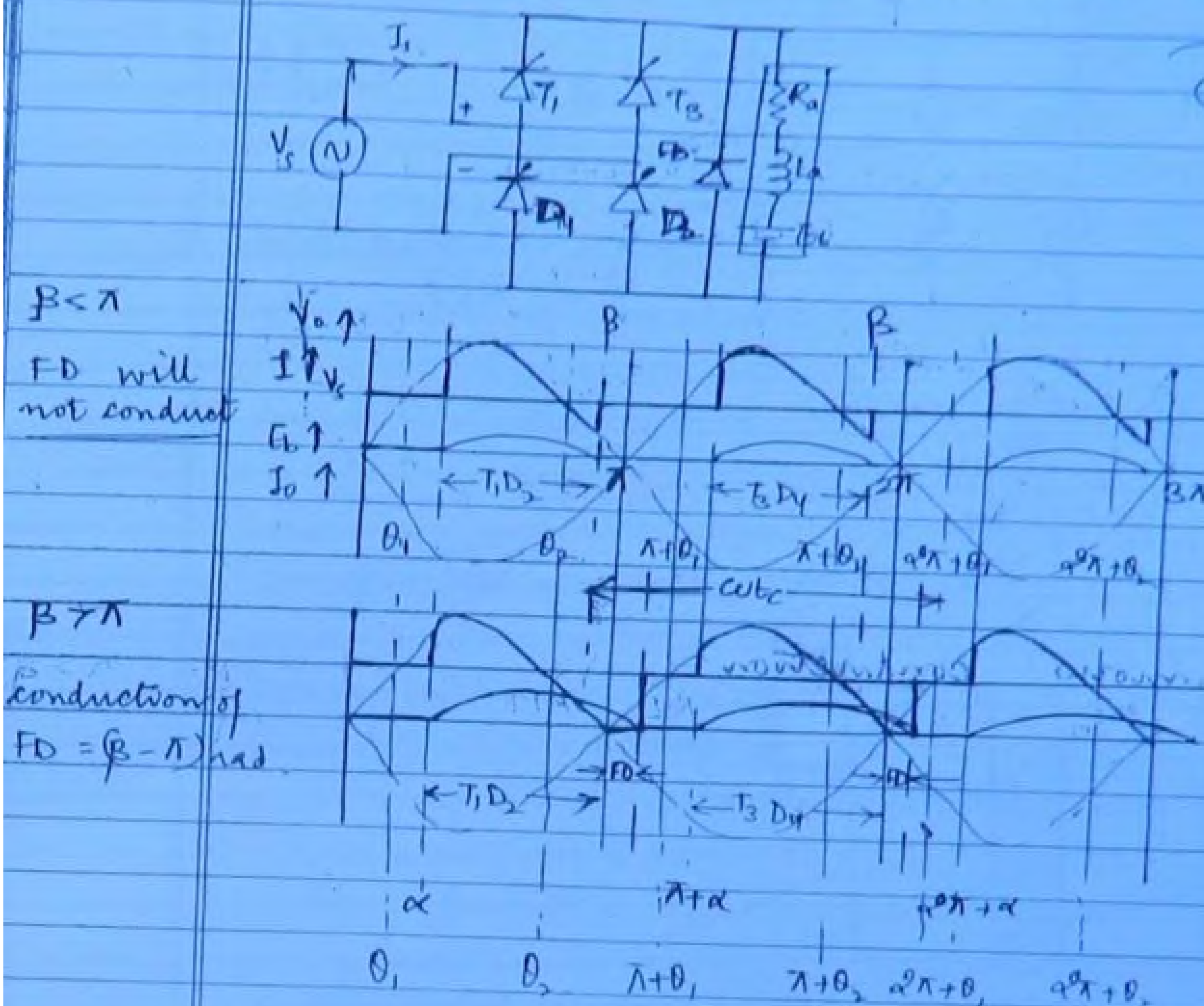


For $\boxed{\beta > \pi}$

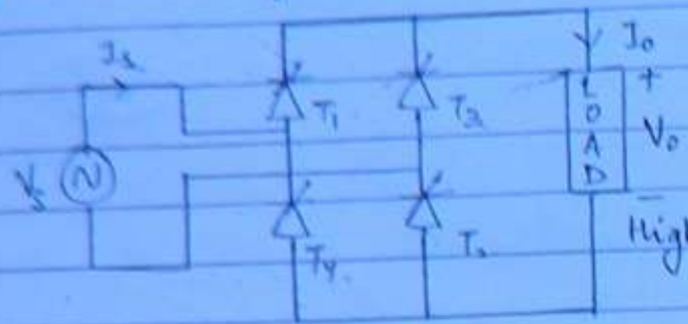
$$V_0 = \frac{1}{\pi}\left[\int_{\alpha}^{\pi} V_m \sin \omega t \, d(\omega t) + \int_{\beta}^{\pi+\alpha} E_b \, d(\omega t)\right]$$

$$V_0 = \frac{1}{\pi}\left[V_m(1+\cos\alpha) + E_b(\pi+\alpha-\beta)\right]$$

For continuous conduction waveform remains same for RL & RLE load.
Back emf will not affect the avg value

$$\boxed{V_0 = \frac{V_m}{\pi}(1+\cos\alpha)}$$

Performance of 1 ϕ Full Converter –

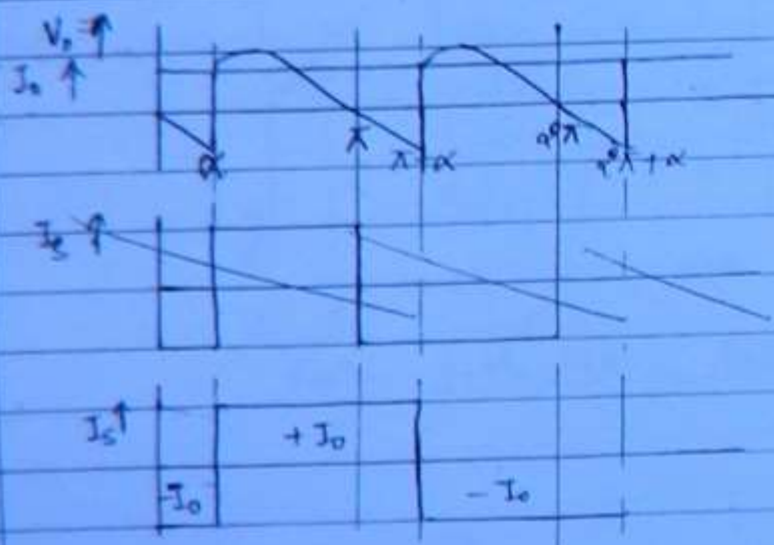


Highly inductive load (RL, RLE)

$$V_o = \frac{2V_m}{\pi}\cos\alpha$$

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

$$I_{sr} = I_o$$



Harmonic Analysis on AC side of converter for source current (I_s)

$$I_s = \sum_{n=1,3,5\ldots}^{\infty} \frac{4I_o}{n\pi}\sin(n\omega t + \phi_n) = \underbrace{\frac{4I_o}{\pi}\sin(\omega t - \alpha)}_{I_{s1}\,(F)} + \underbrace{\frac{4I_o}{3\pi}\sin(3\omega t - 3\alpha)}_{I_{s3}} + \underbrace{\frac{4I_o}{5\pi}\sin(5\omega t - 5\alpha)}_{I_{s5}} \ldots$$

$\phi_n = -n\alpha$

↳ n^{th} harmonic displacement angle

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$$I_{sn} = \frac{4I_0}{n\pi}\sin(n\omega t + \phi_n)$$

(97)

$$(I_{sn})_{rms} = \frac{2\sqrt{2}}{n\pi} I_0$$

$$\boxed{(I_{s1})_{rms} = \frac{2\sqrt{2}}{\pi} I_0} \quad \text{— (1)}$$

$$FDF = \cos\phi_1$$

$$\boxed{FDF = \cos(-\alpha) = \cos\alpha} \quad \text{— (2)}$$

$$g = \frac{(I_{s1})_{rms}}{I_{s}} \Rightarrow \frac{\frac{2\sqrt{2}}{\pi} I_0}{I_0}$$

$$\boxed{g = \frac{2\sqrt{2}}{\pi}} \quad \text{— (3)}$$

$$PF = g\,(FDF)$$

$$\boxed{PF = \frac{2\sqrt{2}}{\pi}\cos\alpha} \quad \text{— (4)}$$

$$THD = \left(\frac{1}{g^2} - 1\right)^{1/2}$$

$$THD = \left(\frac{\pi^2}{8} - 1\right)^{1/2} = 0.4834$$

$$\boxed{THD = 48.34\,\%} \quad \text{— (5)}$$

rms delivers useful power

avg delivers useful power.

Date / Page

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Active power —

$$P = V_{sr} I_{s1} \cos\alpha = V_0 I_0 \quad \text{— (6)}$$

$$= \left(\frac{V_m}{\sqrt{2}}\right)\left(\frac{2\sqrt{2}}{\pi} I_0\right)\cos\alpha$$

$$P = \frac{2V_m}{\pi}\cos\alpha \cdot I_0 = V_0 I_0$$

Reactive Power —

$$Q = V_{sr} I_{s1} \sin\alpha = V_0 I_0 \tan\alpha \quad \text{— (7)}$$

$$= V_{sr} I_{s1} \cos\alpha \cdot \frac{\sin\alpha}{\cos\alpha}$$

$$\boxed{Q = P\tan\alpha}$$

Harmonics on DC side of converter — (V_0)

$$VRF = \sqrt{FF^2 - 1}$$

$$FF = \frac{V_{0r}}{V_0} = \frac{V_m/\sqrt{2}}{\frac{2V_m\cos\alpha}{\pi}} = \frac{\pi}{2\sqrt{2}\cos\alpha}$$

$$\boxed{VRF = \sqrt{\frac{\pi^2}{8\cos^2\alpha} - 1}}$$

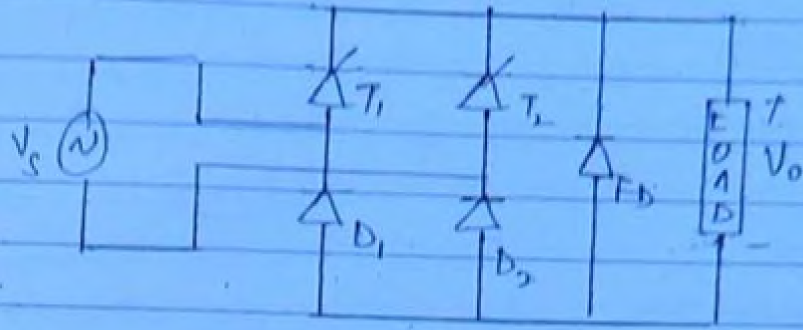
when $0 \le \alpha \le 90$

$\alpha \uparrow$ ripple $\uparrow$ harmonics $\uparrow$

when $90 \le \alpha \le 180°$

$\alpha \uparrow$ ripple $\downarrow$ harmonics $\downarrow$

Performance of 1 φ Semi Converter -

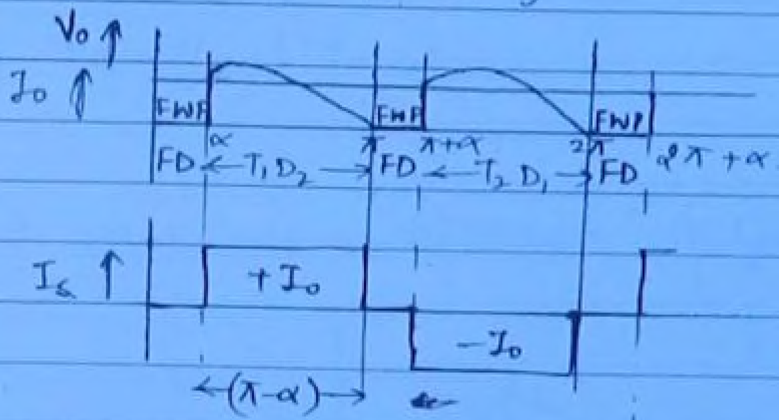


Highly Inductive Load (RL, RLE)

$V_s = V_m \sin \omega t \qquad V_{sr} = \dfrac{V_m}{\sqrt{2}}$

$V_0 = \dfrac{V_m}{\pi}(1 + \cos\alpha)$

$I_{sr} = I_0 \left(\dfrac{\pi - \alpha}{\pi}\right)^{1/2}$.



Harmonic Analysis ~~of~~ on AC side of Converter (I_s)

$$I_s = \sum_{n=1,3,5\ldots}^{\infty} \frac{4 I_0}{n\pi} \cos\frac{n\alpha}{2} \sin(n\omega t + \phi_n)$$

where $\phi_n = -\dfrac{n\alpha}{2}$

$$(I_{sn})_{rms} = \frac{2\sqrt{2} I_0}{n\pi} \cos\frac{n\alpha}{2}$$

$$(I_{s1})_{rms} = \frac{2\sqrt{2}}{\pi} I_0 \cos\frac{\alpha}{2} \quad \text{—(1)}$$

$$FDF = \cos\frac{\alpha}{2} \quad — (2)$$

$$g = \frac{I_{s1}}{I_{sr}} = \frac{\frac{2\sqrt{2}}{\pi} I_o \cos\alpha/2}{I_o\left(\frac{\pi-\alpha}{\pi}\right)^{1/2}}$$

(10M)

$$g = \frac{2\sqrt{2}\cos\frac{\alpha}{2}}{\sqrt{\pi(\pi-\alpha)}} \quad — (3)$$

$$PF = g\,(FDF)$$

$$PF = \frac{2\sqrt{2}\cos\frac{\alpha}{2}}{\sqrt{\pi(\pi-\alpha)}} = \frac{\sqrt{2}(1+\cos\alpha)}{\sqrt{\pi(\pi-\alpha)}} \quad — (4)$$

$$THD = \left(\frac{1}{g^2} - 1\right)^{1/2}$$

$$THD = \left[\frac{\pi(\pi-\alpha)}{8\cos^2\frac{\alpha}{2}} - 1\right]^{1/2} \quad — (5)$$

Active Power –

$$P = V_{sr} I_{s1} \cos\frac{\alpha}{2} = V_o I_o \quad — (6)$$

$$= \frac{V_m}{\sqrt{2}} \cdot \frac{2\sqrt{2}}{\pi} I_o \cos\frac{\alpha}{2} \cos\frac{\alpha}{2}$$

$$= \frac{V_m}{\pi}(1+\cos\alpha) I_o$$

$$P = V_o I_o$$

Reactive Power -

$$Q = V_{sr} I_{s1} \sin\frac{\alpha}{2} = V_0 I_0 \tan\frac{\alpha}{2} \quad \text{--- (7)}$$

$$Q = P \tan\frac{\alpha}{2}$$

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Harmonics on DC side of converter (V_0)

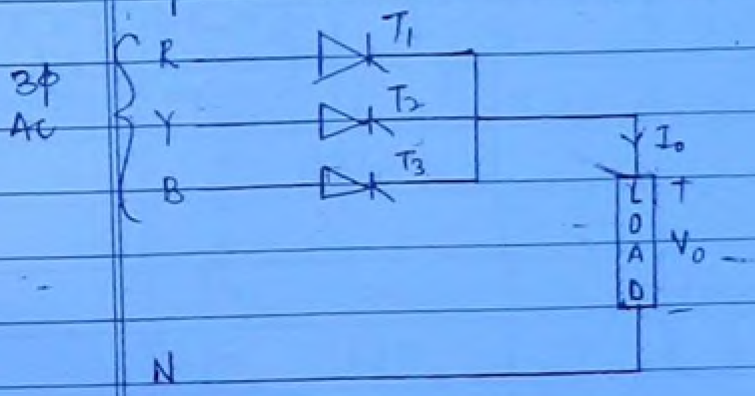
$$VRF = \sqrt{FF^2 - 1}$$

$$FF = \frac{V_{0r}}{V_0}$$

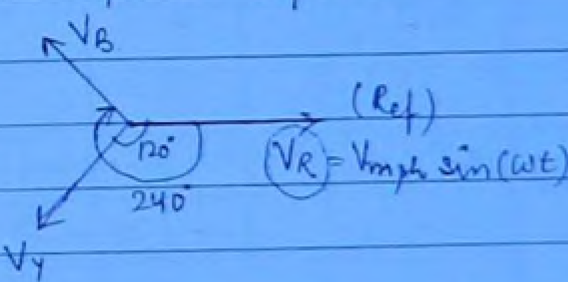
$$V_{0r} = \frac{V_m}{\sqrt{2\pi}}\left[(\pi - \alpha) + \frac{1}{2}\sin 2\alpha\right]^{1/2}$$

$$V_0 = \frac{V_m}{\pi}(1 + \cos\alpha)$$

3φ HALF WAVE RECTIFIER (3 pulse converter)

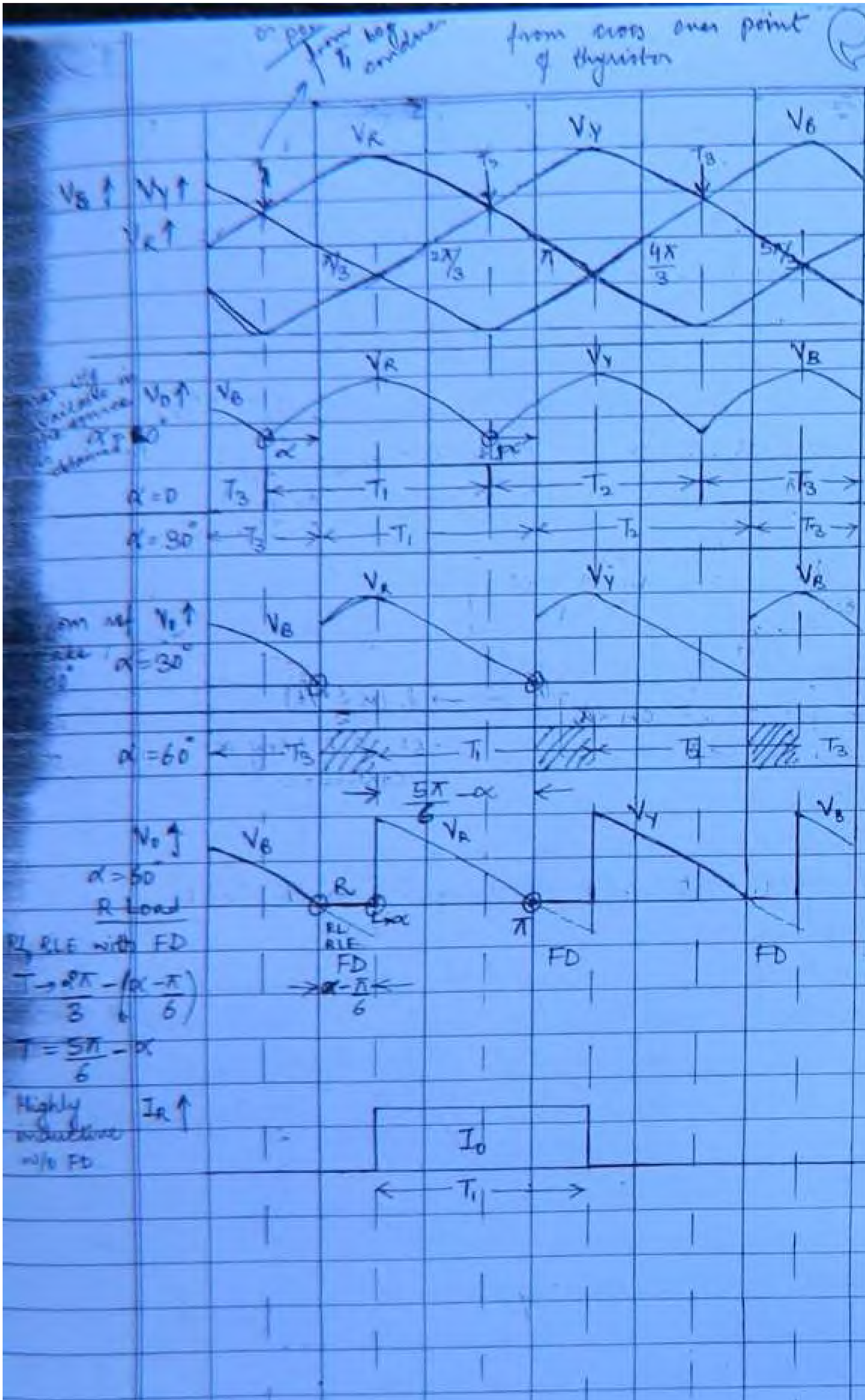


RYB phase sequence



$$I_L = I_{ph} = I_T$$

$$V_{ML} = \sqrt{3}\, V_{mph}$$



Always output voltage is taken the max v/g (in which ever phase it might be)

FD → $(\alpha - \pi/6)$

For R load, upper limit remains π only as its independent of α. Even if $\alpha \uparrow$ from 60° to 61° the makes are remove at π only.

for L load -ve spikes occur

I $\boxed{\alpha \leq \frac{\pi}{6}}$ ⇒ continuous conduction for R load

$$V_0 = \frac{1}{2\pi/3}\int_{\alpha+\pi/6}^{\alpha+5\pi/6} V_{Mph}\sin\omega t \, d(\omega t)$$

TD3

$$\boxed{V_0 = \frac{3\sqrt{3}\,V_{Mph}}{2\pi}\cos\alpha = \frac{3V_{ML}}{2\pi}\cos\alpha}$$

→ R ($\alpha \leq \pi/6$)

→ RL, RLE (any α) continuous

$$V_{or} = \left\{\frac{1}{2\pi/3}\int_{\alpha+\pi/6}^{\alpha+5\pi/6} V_{Mph}^2\sin^2\omega t \, d(\omega t)\right\}^{1/2}$$

$$\boxed{V_{or} = V_{mL}\left[\frac{1}{6} + \frac{\sqrt{3}}{8\pi}\cos 2\alpha\right]^{1/2}}$$

→ R ($\alpha \leq \pi/6$)

→ RL, RLE (Any α) continuous

II $\boxed{\alpha > \frac{\pi}{6}}$ ⇒ discontinuous conduction for R load.

$$V_0 = \frac{1}{2\pi/3}\int_{\alpha+\pi/6}^{\pi} V_{Mph}\sin\omega t \, d(\omega t)$$

$$\boxed{V_0 = \frac{3V_{Mph}}{2\pi}\left[1+\cos\left(\alpha+\frac{\pi}{6}\right)\right]}$$

→ R ($\alpha > \pi/6$)

→ RL, RLE with FD ($\alpha > \frac{\pi}{6}$)

$$V_{or} = \left\{\frac{1}{2\pi/3}\int_{\alpha+\pi/6}^{\pi} V_{Mph}^2\sin^2\omega t \, d(\omega t)\right\}^{1/2}$$

$$\boxed{V_{or} = \frac{V_{ML}}{2\sqrt{\pi}}\left[\left(\frac{5\pi}{6}-\alpha\right) + \frac{1}{2}\sin\left(2\alpha+\frac{\pi}{3}\right)\right]^{1/2}}$$

→ R ($\alpha > \pi/6$)

→ RL, RLE with FD ($\alpha > \pi/6$)

Assume highly Inductive Load with FD -

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I $\boxed{\alpha \leq \frac{\pi}{6}}$ * FD will not conduct

$\therefore$ Conduction angle of each thyristor $= \frac{2\pi}{3}$ ie 120°

[for every 2π rad.]

$$(I_T)_{avg} = I_0 \left(\frac{2\pi/3}{2\pi}\right) = \frac{I_0}{3}$$

$$(I_L = I_{ph} = I_T)_{avg} = \frac{I_0}{3}$$

$$(I_L = I_{ph} = I_T)_{rms} = \frac{I_0}{\sqrt{3}}$$

II $\boxed{\alpha > \frac{\pi}{6}}$ Conduction angle of FD $= \left(\alpha - \frac{\pi}{6}\right)$ for every $2\pi/3$ radians.

Conduction angle of each thyristor $= \left(\frac{5\pi}{6} - \alpha\right)$ for every 2π radians

$$(I_L = I_{ph} = I_T)_{avg} = I_0 \left[\frac{(5\pi/6 - \alpha)}{2\pi}\right]$$

$$(I_L = I_{ph} = I_T)_{rms} = I_0 \left[\frac{(5\pi/6 - \alpha)}{2\pi}\right]^{1/2}$$

$$(I_{FD})_{avg} = I_0 \left[\frac{(\alpha - \pi/6)}{2\pi/3}\right]$$

$$(I_{FD})_{rms} = I_0 \left[\frac{(\alpha - \pi/6)}{2\pi/3}\right]^{1/2}$$

Assume highly Inductive Load without FD -
Any α

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Conduction angle of each thyristor $= \frac{2\pi}{3}$ $\left[\begin{matrix}\text{for every} \\ 2\pi \text{ rad}\end{matrix}\right]$

$$\therefore (I_L = I_{ph} = I_T)_{avg} = I_0 \left(\frac{2\pi/3}{2\pi}\right) = \frac{I_0}{3}$$

$$(I_L = I_{ph} = I_T)_{rms} = \frac{I_0}{\sqrt{3}}$$

$$(I_s)_{avg} = \frac{I_0}{3} \text{ (DC comp)}$$

Drawback
The source current contains DC component & saturates the ~~tran~~ supply transformer core

RATINGS OF SCR -

1. $(I_T)_{RMS}$ Rating (RMS Rating of ON state current) provided by manufacturer

$(I_T)_{RMS}$ Rating $\geq (I_T)_{rms}$ value of in a converter

eg

1-ϕ full conv $(I_T)_{rms} = \frac{I_0}{\sqrt{2}}$

1-ϕ semi conv $(I_T)_{rms} = I_0 \left(\frac{\pi - \alpha}{2\pi}\right)^{1/2}$

3 pulse $(I_T)_{rms} = \frac{I_0}{\sqrt{3}}$

2. (I_{TAV}) Rating [Average ON state current rating]

$$(I_{TAV})\text{Rating} = \frac{(I_T)_{RMS}\text{ Rating}}{FF}$$

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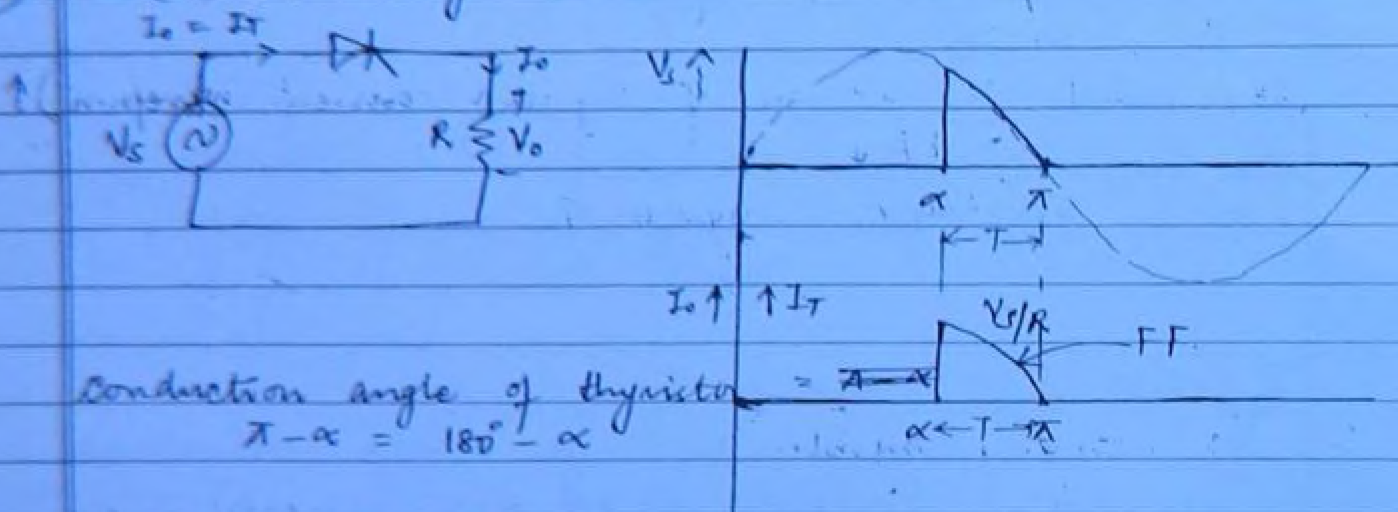
FF = form factor of thyristor current waveform in a converter
↓
depends on shape of waveform.

CWB chapter 1

20 $(I_T)_{rms}$ Rating = 35 A



Conduction angle of thyristor
$\pi - \alpha = 180° - \alpha$

Given $180° - \alpha = 30°$
$\alpha = 150°$

$$FF = \frac{(I_T)_{RMS}}{(I_T)_{avg}} = \frac{I_{or}}{I_o} = \frac{V_{or}/R}{V_o/R} = \frac{V_{or}}{V_o}$$

$$FF = \frac{\dfrac{V_m}{2\sqrt{\pi}}\left[\underbrace{(\pi-\alpha)}_{\pi/6\text{ given}} + \frac{1}{2}\sin 2\alpha\right]^{1/2}}{\dfrac{V_m}{2\pi}[1+\cos\alpha]}$$

$$= \frac{\sqrt{\pi}\left[\left(\frac{\pi}{6}\right) + \frac{1}{2}\sin\frac{5\pi}{3}\right]^{1/2}}{[1+\cos 150°]} = 3.98$$

$$I_{TAV} = \frac{(I_T)_{RMS}\ \text{Rating}}{FF} = \frac{35}{3.98}$$

$$= 8.79\ A \quad (d)$$

(107)

Avg rating depends on –

Conduction angle of thyristor

As conduction angle $\uparrow$ $\Rightarrow$ (smoothness of I_T waveform) $\uparrow$

$\Rightarrow FF \downarrow$

Thus avg rating of thyristor $\uparrow$

Load parameters

eg. $L \uparrow \Rightarrow$ (smoothness of thyristor current waveform) $\uparrow$

$\Rightarrow FF \downarrow$

(I_{TAV}) Rating of Thyristor $\uparrow$

(I^2t rating of thyristor)

provided by manufacturer – to select a proper fuse for the thyristor

I^2t rating of thyristor $>$ I^2t rating of fuse

Surge Current Rating of Thyristor

i) n cycle surge current rating – (I_{sn})

Its the surge current that the SCR can withstand for n cycles

$$\therefore (I_{sn})^2\, n \frac{T}{2} \overset{\text{at the most}}{=} I^2t \text{ rating}$$

ii) one cycle surge current rating (I_{sr})

It is the surge current that the SCR can withstand for one cycle.

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$$(I_{s1})^2 \cdot \frac{T}{2} = (I_{sn})^2 \cdot n\frac{T}{2}$$

$$I_{s1} = \sqrt{n}\, I_{sn}$$

iii) Sub-cycle surge current rating ($I_{s/n}$)

It is the surge current that the SCR can withstand for 1/n th period of a cycle

$$(I_{s/n})^2 \frac{T}{2} \cdot \frac{1}{n} = (I_{s1})^2 \cdot \frac{T}{2}$$

$$I_{s/n} = \sqrt{n}\, I_{s1}$$

iv) Half cycle surge current.

$$I_{s/2} = \sqrt{2}\, I_{s1}$$

CWB chapter 1

(19) $I_{s/n} = 3000$

$$I_{s1} = \frac{3000}{\sqrt{2}} = 2121.32\ A \quad (b)$$

(21) (c)

CWB chapter 2

(2) (b) −ve spikes are removed.

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(3) Half wave rectifier

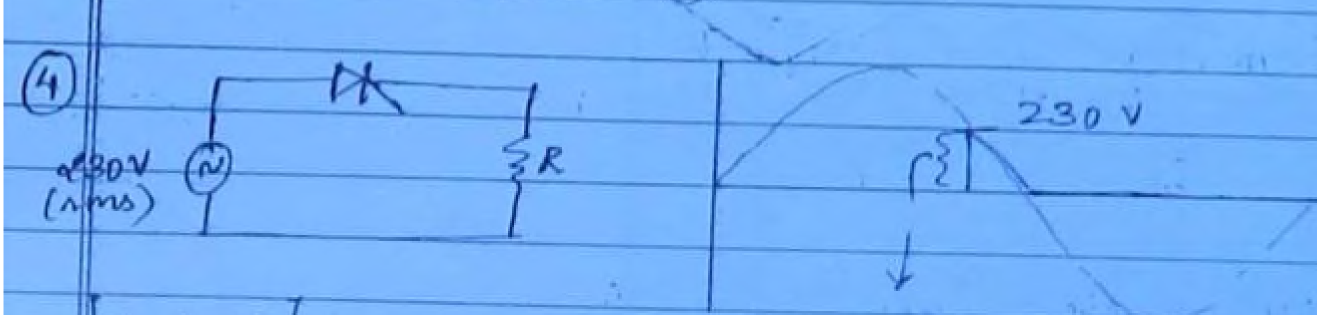
PIV depends on secondary not on primary.

$V_s = 50\,V$ (rms) $\therefore V_m = 50\sqrt{2}$

$PIV = 2V_M = 50\sqrt{2}\,(2)$

$= 100\sqrt{2}$ (a)

(4)



$\left[V_0(\omega t)\right]_{peak} = 230\,V$

V_0 for $\alpha \leq 90^\circ$ ×
$V_m = 230\sqrt{2}$

So $\alpha > 90^\circ$

$V_m \sin\alpha = 230^\circ$

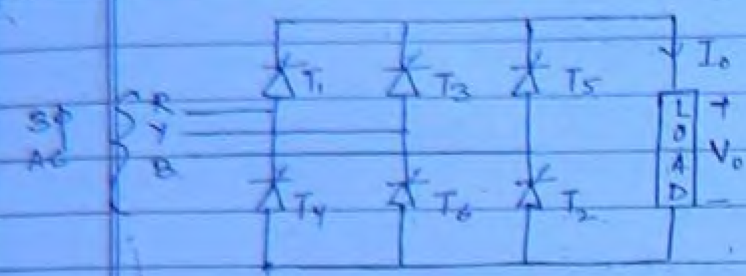
$230\sqrt{2}\sin\alpha = 230^\circ$

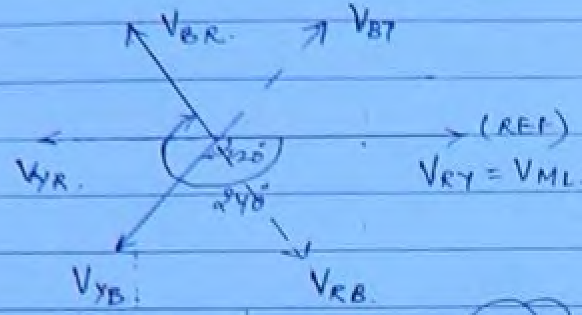
$\sin\alpha = \frac{1}{\sqrt{2}}$

$\alpha = \cancel{45^\circ}, 135^\circ$

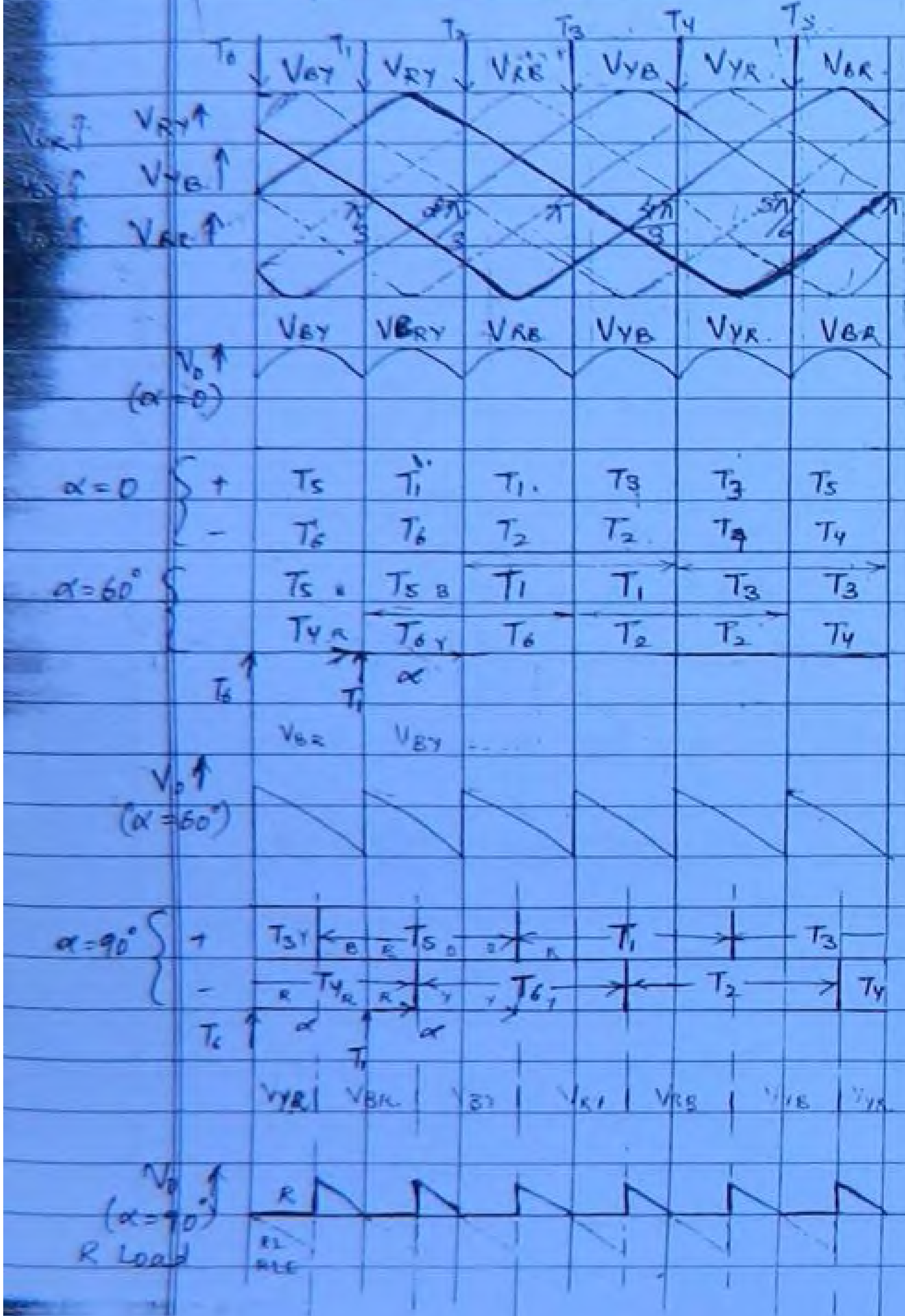
Ans (b)

3ϕ Fully Controlled Rectifier (6 pulse converter)





$V_{RY} = V_{ML} \sin \omega t$

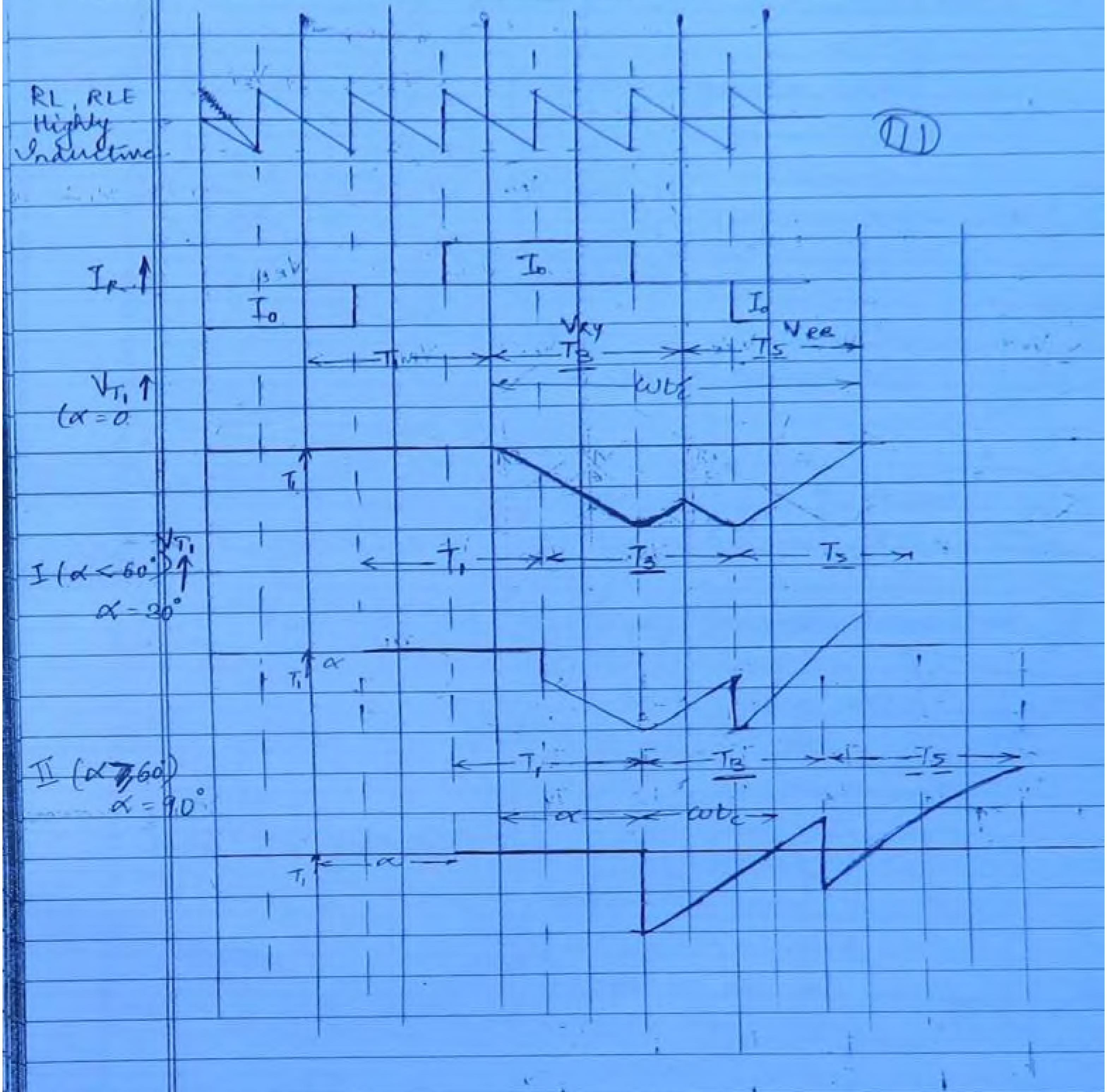


count α from crossover pt.

$+$ $T_1\ T_3\ T_5$ } seq remains same
$-$ $T_4\ T_6\ T_2$ }

$T_3 \to ON \quad V_{T1} = \underline{V_{RY}}$

$T_5 \to ON \quad V_{T1} \to \underline{V_{RB}}$



$\alpha < \pi/3 \rightarrow$ continuous conduction for R load

$$V_0 = \frac{1}{\pi/3}\int_{\alpha+\pi/3}^{\alpha+2\pi/3} V_{ML}\sin\omega t\, d(\omega t)$$

$$\boxed{V_0 = \frac{3V_{ML}}{\pi}\cos\alpha}$$

$\rightarrow$ R ($\alpha \leq \pi/3$)

$\rightarrow$ RL, RLE (Any α) continuous

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$$V_{0R} = \left\{\frac{1}{\pi/3}\int_{\alpha+\pi/3}^{\alpha+2\pi/3} V_{ML}^2\sin^2\omega t\, d(\omega t)\right\}^{1/2}$$

$$= \sqrt{\frac{3}{2\pi}}\, V_{ML}\left\{\frac{\pi}{3} + \frac{1}{2}\left[\sin\left(2\alpha + \frac{2\pi}{3}\right) - \sin\left(2\alpha + \frac{4\pi}{3}\right)\right]\right\}^{1/2}$$

$\alpha > \pi/3 \rightarrow$ discontinuous conduction for R load -

$$V_0 = \frac{1}{\pi/3}\int_{\alpha+\pi/3}^{\pi} V_{ML}\sin\omega t\, d\omega t$$

$$\boxed{V_0 = \frac{3V_{ML}}{\pi}\left[1+\cos\left(\alpha+\frac{\pi}{3}\right)\right]} \rightarrow \text{R } (\alpha > \pi/3)$$

$$V_{0R} = \left\{\frac{1}{\pi/3}\int_{\alpha+\pi/3}^{\pi} V_{ML}^2\sin^2\omega t\, d\omega t\right\}^{1/2}$$

$$= \sqrt{\frac{3}{2\pi}}\, V_{ML}\left\{\left(\frac{2\pi}{3}-\alpha\right) + \frac{1}{2}\sin\left(2\alpha+\frac{2\pi}{3}\right)\right\}^{1/2} \rightarrow \text{R } (\alpha > \pi/3)$$

Assume highly inductive load - (RL, RLE)

Conduction angle of each thyristor $= \frac{2\pi}{3}$ (for every 2π rad)

$$(I_T)_{avg} = I_0\left(\frac{2\pi/3}{2\pi}\right) = \frac{I_0}{3}$$

(TB)

$$(I_T)_{rms} = \frac{I_0}{\sqrt{3}}$$

$$(I_R)_{rms} = I_0\left(\frac{2\pi/3}{\pi}\right)^{1/2}$$

$$I_{SR} = (I_R)_{rms} = I_0\sqrt{\frac{2}{3}}$$

Harmonic Analysis on AC side of converter for source current (I_S) waveform

$$I_S = \sum_{n=1,5,5}^{\infty} \frac{4I_0}{n\pi}\sin\frac{n\pi}{3}\sin(n\omega t + \phi_n)$$

$n = 6K \pm 1$

Since $\sin\frac{3\pi}{3} = 0$ so 3^{rd} harmonic & multiples of 3 harmonics (triple harmonics) are absent. So are even harmonics

NOTE: Even & triple harmonics are absent

$\phi_n = -n\alpha \qquad \phi_1 = -\alpha$

$$(I_{Sn})_{rms} = \frac{2\sqrt{2}}{n\pi} I_0 \sin\frac{n\pi}{3}$$

$$(I_{S1})_{rms} = \frac{2\sqrt{2}}{\pi} I_0 \sin\frac{\pi}{3}$$

$$(I_{s1})_{rms} = \frac{\sqrt{6}}{\pi} I_0 \quad \text{—(1)}$$

$$FDF = \cos\phi_1$$

$$FDF = \cos\alpha \quad \text{—(2)}$$

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$$g = \frac{(I_{s1})_{rms}}{I_{sr}} = \frac{\frac{\sqrt{6}}{\pi} \cdot I_0}{I_0\sqrt{\frac{2}{3}}}$$

$$g = \frac{3}{\pi} \quad \text{—(3)}$$

$$THD = \left(\frac{1}{g^2} - 1\right)^{1/2} = \left(\frac{\pi^2}{9} - 1\right)^{1/2}$$

$$THD = 31\% \quad \text{—(4)}$$

m ↑ THD ↓ ∴ harmonics ↓

$$PF = g(FDF)$$

$$PF = \frac{3}{\pi} \cos\alpha \quad \text{—(5)}$$

Active power $P = \sqrt{3}\, V_{sr} I_{s1} \cos\alpha$ —(6)

$$= \sqrt{3} \frac{V_{ML}}{\sqrt{2}} \frac{\sqrt{6}}{\pi} I_0 \cos\alpha$$

$$= \frac{3V_{ML}}{\pi} \cos\alpha \cdot I_0 = V_0 I_0$$

Reactive power ⇒ $Q = \sqrt{3}\, V_{sr} I_{s1} \sin\alpha = V_0 I_0 \tan\alpha$ —(7)

$$= P\tan\alpha$$

To find t_c & PIV across thyristor we need to plot V_{T_1}.

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$T_1 \rightarrow ON \quad V_{T1} = 0$

$T_3 \rightarrow ON \quad V_{T1} = V_{RY}$

$T_5 \rightarrow ON \quad V_{T1} = V_{RB}$

$\omega t_c = \frac{4\pi}{3} \qquad t_c = \frac{4\pi}{3\omega}$ secs.

I) $\alpha < 60^\circ$

$\alpha + \omega t_c = \frac{4\pi}{3}$

$\omega t_c \Rightarrow \frac{4\pi}{3} - \alpha$

$t_c = \frac{\frac{4\pi}{3} - \alpha}{\omega}$

II $\alpha \geq 60^\circ$

$\alpha + \omega t_c = \pi$

$\omega t_c = \pi - \alpha$

$t_c = \frac{\pi - \alpha}{\omega}$ secs.

3 pulse converter –

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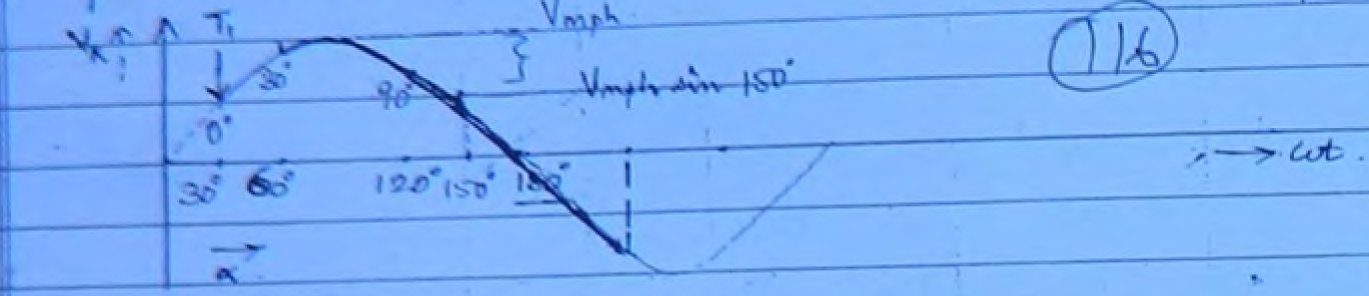
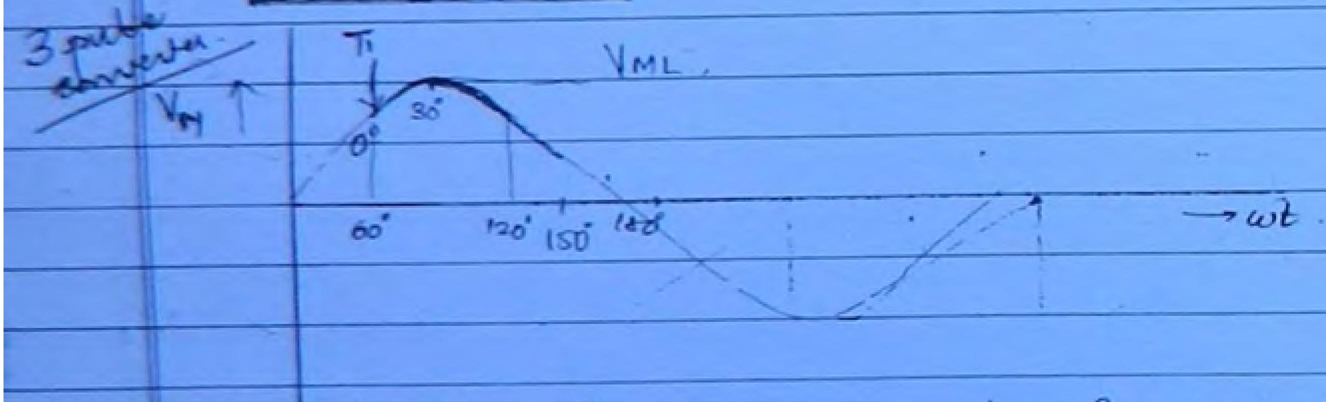


$\alpha = \omega t - 30°$

Length of pulse $= \frac{2\pi}{3} = 120°$

$\boxed{\begin{matrix} 0 \le \alpha \le 150° \\ 0 \le \alpha \le 180° \end{matrix}}$ $\begin{matrix} \rightarrow \text{R load} \\ \rightarrow \text{Inductive Load} \end{matrix}$



$\alpha = \omega t - 60°$

Length of pulse $= \pi/3 = 60°$, for R load → $0 \le \alpha \le$

RL, RLE load → $0 \le \alpha \le 1$

CWB chapter 2

① $\frac{\text{Peak to peak vlg ripple}}{\text{peak o/p dc voltage}} = \frac{V_{ML} - V_{ML} \sin 150°}{V_{ML}}$

$= 1 - \sin 150°$

$= 0.5$ (a)

Date
Page

3φ Half Controlled Rectifier

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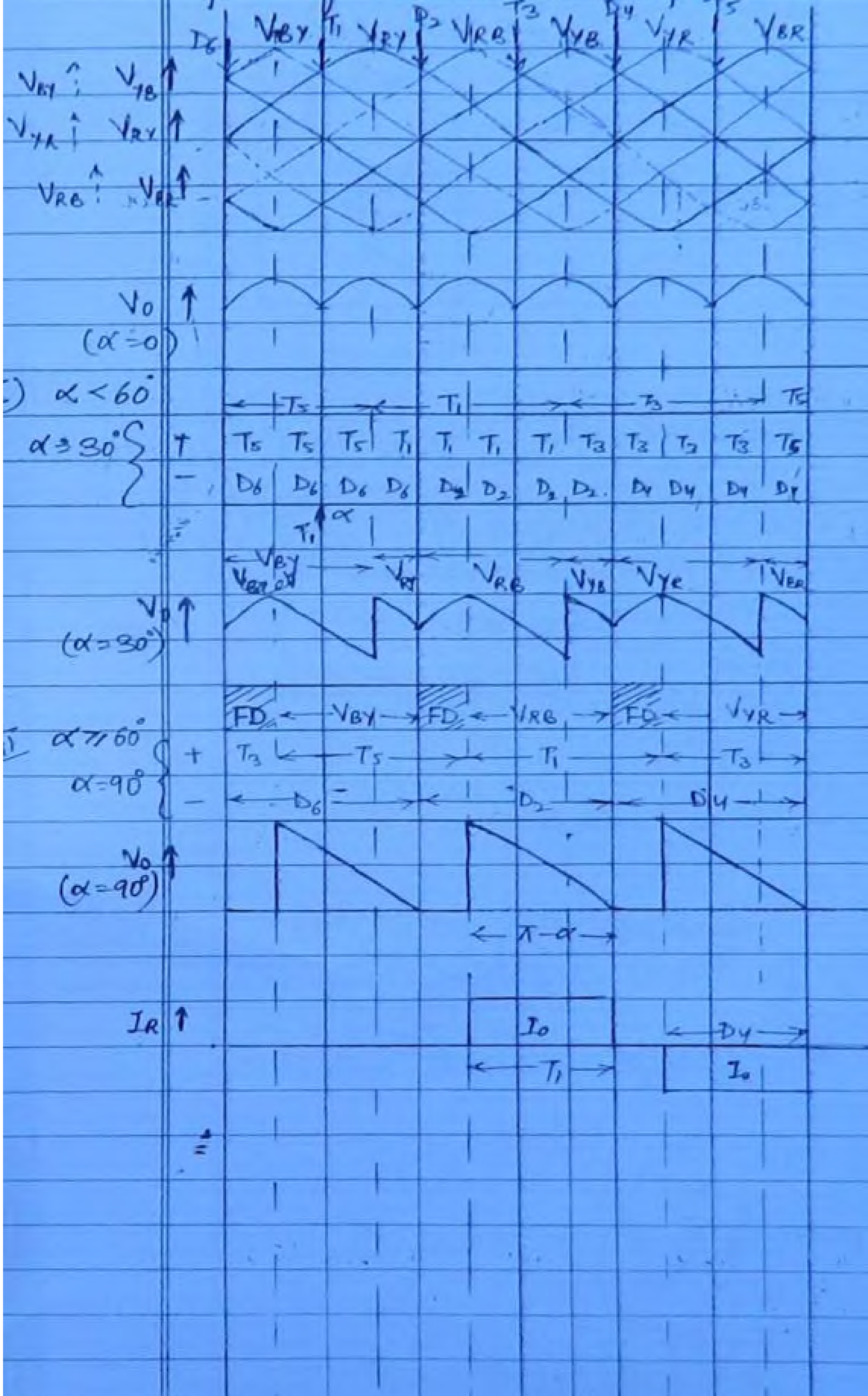

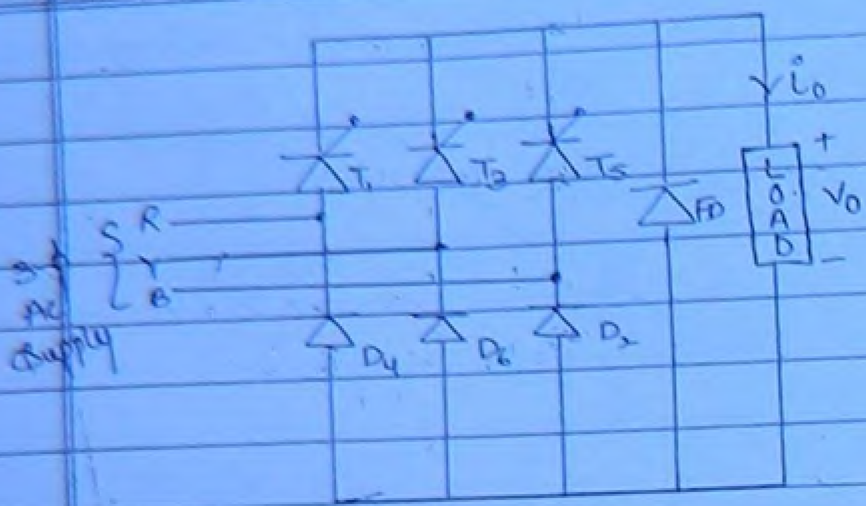

$+\ T_1, T_3, T_5$

$-\ D_4, D_6, D_2$

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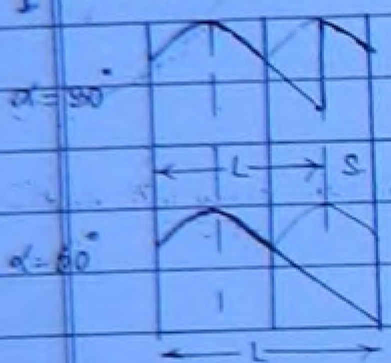
Short method − 1 mark

I $\alpha < 60^\circ$ FD will not conduct.



$\rightarrow S = 60 - \alpha$
$= 60 - 60$
$= 0$

Take $V_0 (\alpha = 0)$ as ref

Long pulse $= 60 + \alpha$

Short pulse length $= 60 - \alpha$

$$\boxed{\begin{array}{l} \alpha < 60^\circ \Rightarrow 6 \text{ pulse} \\ \alpha \geq 60^\circ \Rightarrow 3 \text{ pulse.} \end{array}}$$ IES

For $\alpha < 60^\circ$ FD will not conduct.

for $\alpha \geq 60^\circ$

conduction period of FD $= \left(\alpha - \frac{\pi}{3}\right)$ radians

conduction period of thyristor $= \frac{2\pi}{3} - \left(\alpha - \frac{\pi}{3}\right)$

$= \pi - \alpha$

$\therefore$ Length of pulse $= \pi - \alpha$

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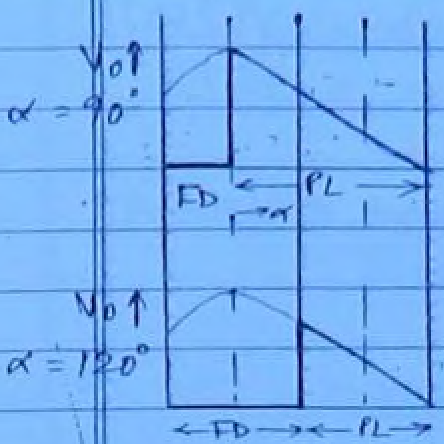


Take $V_0(\alpha = 60°)$ as ref

FD $\rightarrow \alpha - \pi/3$

Pulse length $= \pi - \alpha$

As $\alpha \uparrow$ Free wheeling $\uparrow$

Pulse length $\downarrow$

Assume Highly Inductive Load

$\alpha \leq 60°$ FD will not conduct

~~Conduction angle of FD = $(\alpha - \pi/3)$~~

Conduction angle of thyristor $= \frac{2\pi}{3}$ rad (for every 2π radian

$$(I_T)_{avg} = I_0 \left(\frac{2\pi/3}{2\pi}\right) = \frac{I_0}{3}$$

$$(I_T)_{rms} = \frac{I_0}{\sqrt{3}} \qquad I_{sr} = I_0 \sqrt{\frac{2}{3}}$$

$\alpha > 60°$ FD will conduct

Conduction angle of FD $= \alpha - \frac{\pi}{3}$ (for every $\frac{2\pi}{3}$ rad)

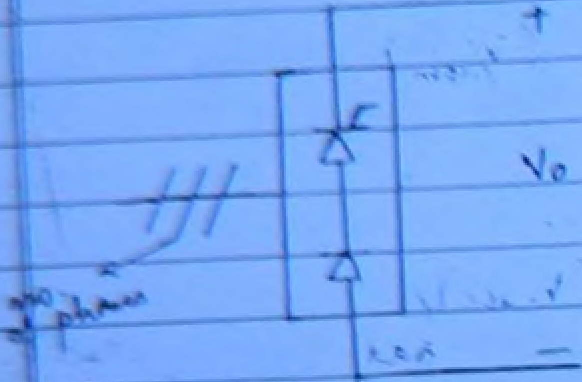
Conduction angle of thyristor $= \pi - \alpha$ (for every 2π rad)

$$(I_T)_{avg} = I_0 \left(\frac{\pi - \alpha}{2\pi}\right)$$

$$(I_T)_{rms} = I_0 \left(\frac{\pi - \alpha}{2\pi}\right)^{1/2}$$

$$I_{sr} = I_0 \left(\frac{\pi - \alpha}{\quad}\right)^{1/2}$$

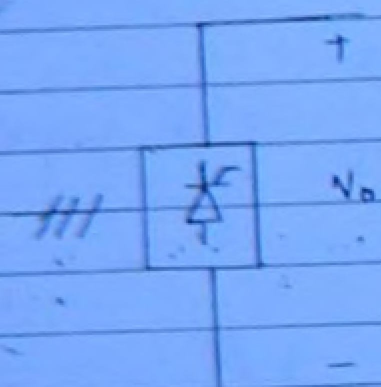
$$\boxed{V_o = \frac{3V_{ML}}{2\pi}(1+\cos\alpha)}$$



Representation of semi-converter

$$V_o = \frac{V_m}{\pi}(1+\cos\alpha) \rightarrow 1\phi$$

$$V_o = \frac{3V_{ML}}{2\pi}(1+\cos\alpha) \rightarrow 3\phi$$



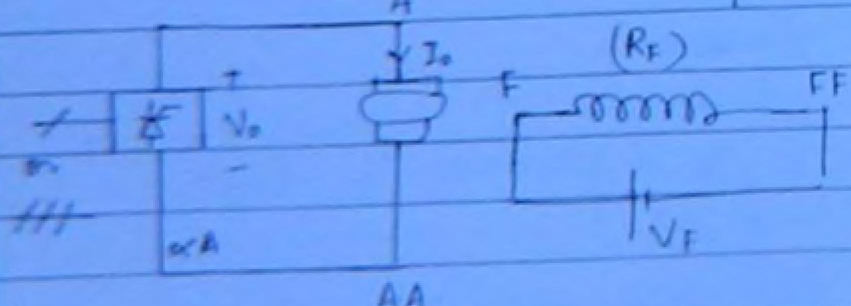
Representation of full converter.

APPLICATIONS -

DC Drives -

I Armature Voltage Control $(\omega < \omega_R)$

$\phi =$ const



$1\phi \quad V_o = \frac{2V_m}{\pi}\cos\alpha_A$

$E_b \propto \phi N$

$E \propto N \quad (\because \phi = \text{const})$

$\boxed{E = \underbrace{K}N}$ → EMF const (V/rpm)
or Motor const

$\boxed{E = \underbrace{K}\omega}$ → EMF const (V.sec)/rad
or Motor const

$T_a \propto \phi I_0$

$T_a \propto I_0 \quad (\phi = \text{const})$

$\boxed{T_a = \underbrace{K} I_0}$ → $I_0 = T_a/K$

→ Motor const $NM/A = (V.sec)/rad$
or Torque const

$\omega = \dfrac{2\pi N}{60}$ $\quad N \to rpm$
$\quad \omega \to rad/sec$

For Motoring Mode

$V_0 = E_b + I_0 R_a$

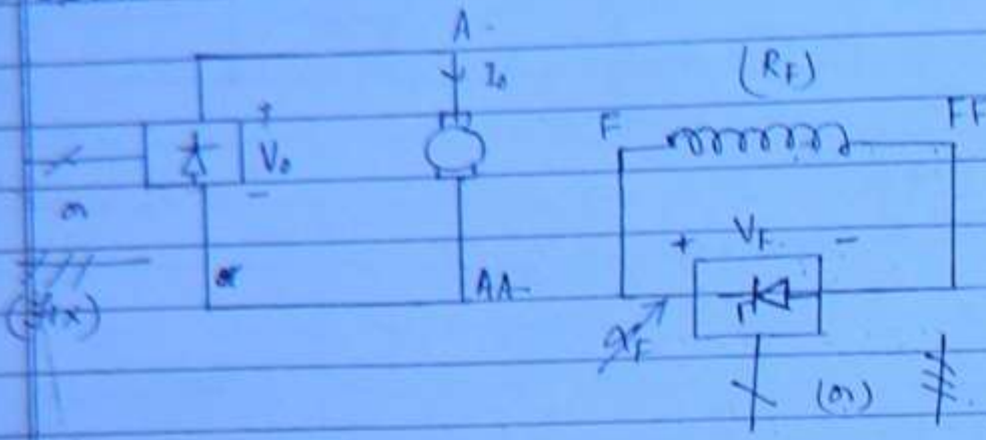
$V_0 = K\omega + I_0 R_a$

$\omega = \dfrac{V_0}{K} - \dfrac{I_0 R_a}{K}$

$$\boxed{\omega = \frac{V_0}{K} - \frac{R_a T_a}{K^2}}$$

$\alpha_A \uparrow \quad V_0 \downarrow \quad \therefore \omega \downarrow \quad (\omega \propto \omega_1)$

(II) ~~Field~~ Field Control Method ($\omega > \omega_N$)



$1\phi \quad V_F = \frac{2V_m}{\pi} \cos \alpha_F$

$I_F = \frac{V_F}{R_F}$

$3\phi \quad V_F = \frac{3V_{ML}}{\pi} \cos \alpha_F$

$\phi \propto I_F$

$\phi = K_F I_F$

$E_b \propto \phi N \propto (K_F I_F) N$

$E_b = K_1 (K_F I_F) N$

$E = \underbrace{K}_{} I_F N$

$\rightarrow$ EMF const V/rpm. A

or Motor const

$E = \underbrace{k}_{} I_F \omega$

$\rightarrow$ EMF const ~~V/sec~~ V.sec/rad. A

or Motor const

$T_a \propto \phi I_a$

$T_a = k I_F I_a$

$T_a = \underbrace{k}_{} I_F I_a \rightarrow I_a = \frac{T_a}{k I_F}$

$\rightarrow$ Motor const V.sec/rad A

or Torque const

Date
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$\omega = \frac{2\pi N}{60}$

(123)

For Motoring Mode

$V_0 = E_b + I_0 R_a$
$V_0 = K I_F \omega + I_0 R_a$

$$\omega = \frac{V_0}{K I_F} - \frac{I_0 R_a}{K I_F}$$

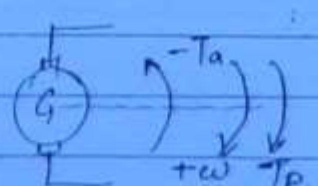
$$\omega \uparrow = \frac{V_0}{K I_F \uparrow} - \frac{R_a}{(K I_F)^2} T_a$$

$\alpha_f \uparrow \quad V_f \downarrow \Rightarrow I_f \downarrow \Rightarrow \omega_0 > \omega_r$

for Motoring Mode ⇒ Torque developed is in same dirⁿ as speed

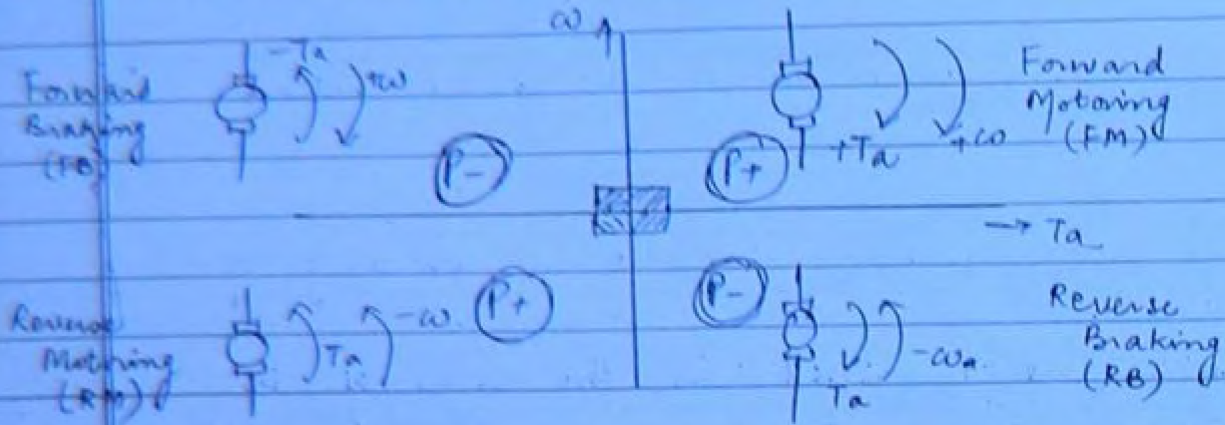
Current enters at +ve terminal of back emf so that electrical effect absorbed is transformed in mech. energy.

for Generating Mode ⇒ Torque developed & speed in opp dirⁿ



Four Modes of DC M/c -

We can utilise DC m/c in 4 modes

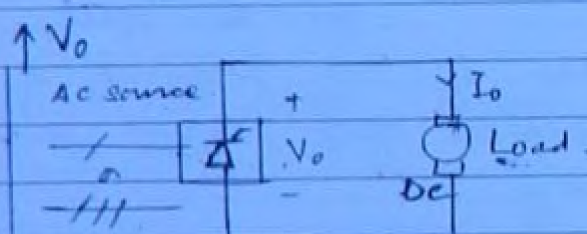


Quadrant Operation of Full converter - (Two Quadrant Operation)

1ϕ. $V_0 = \frac{2V_m}{\pi} \cos\alpha$ $\qquad \alpha < 90 \quad V_0 +$.

$\qquad \alpha > 90 \quad V_0 -$

3ϕ. $V_0 = \frac{3V_{ML}}{\alpha} \cos\alpha$



$V_0 + \ (\alpha < 90°)$

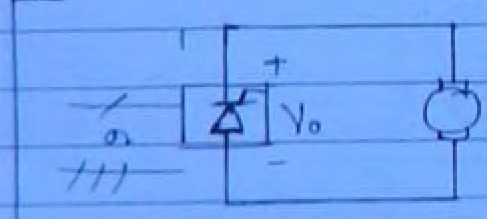
$I_0 +$ P+

$P_{source} \rightarrow P_{load}$

$P_{AC} \rightarrow P_{DC}$ (Rectification mode)

$(\alpha < 90°)$

I_0 (always) +



$V_0 - \ (\alpha > 90°)$

$I_0 +$ P−

$P_{source} \leftarrow P_{load}$

Rectification Mode –

→ can be used for motoring mode of a DC m/c.

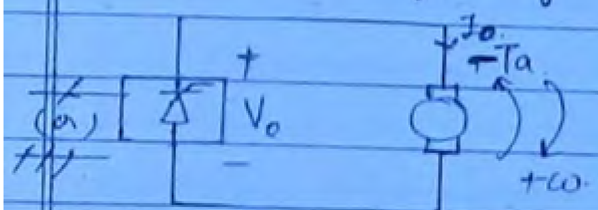
→ can also be used to charge a battery.

Inversion Mode

→ can be used for regenerative braking of DC m/c.

→ solar energy stored in the form of DC can be given to the AC side of utility system where the converter is operating in (inversion) mode.

Full converter feeding DC m/c –

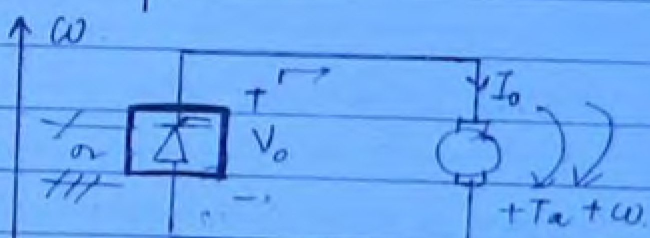


$P_{AC} \xleftarrow[(\alpha > 90^\circ)]{\text{Inv. Mode}} P_{DC}$

Source ← Braking Energy (Regenerative Braking)

Converter will support the invertion if $\alpha > 90^\circ$

Load supports invertion if emf is having ability to deliver power.



$P_{AC} \xrightarrow[(\alpha < 90^\circ)]{\text{Rec. Mode}} P_{DC}$

$\alpha < 90^\circ$ $V_o +$ $E_b +$ $\therefore \omega +$

$I_a +$ $\therefore T_a +$

$T_a \to \phi I_a$

$++$

Converter will support the rectification if $\alpha < 90^\circ$

Load supports rectification if emf is having ability to absorb the power

$V_0 = -E_b + I_0 R_a$ 126	$V_0 = E_b + I_0 R_a$ $\omega = \frac{2V_m}{\pi K}\cos\alpha - \frac{R_a}{K^2}T_a \rightarrow 1\phi$ $\omega = \frac{3V_{mL}}{\pi K}\cos\alpha - \frac{R_a}{K^2}T_a \rightarrow 3\phi$ $T_a +, \omega + \therefore$ P+ $(\phi +)$

Quadrant Operation of Semi converter
Supports

16-12-12

DUAL CONVERTER (Four Quadrant Operation)

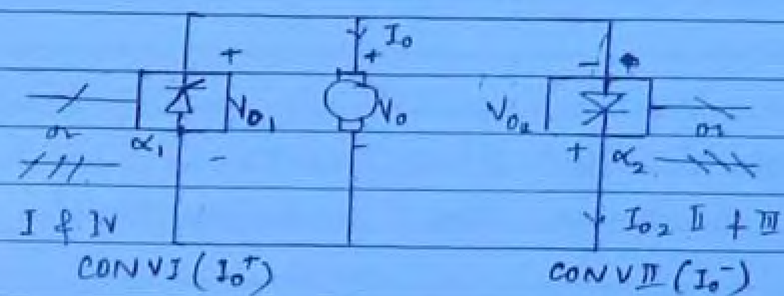
T27

i) Non-Circulating current Type -

In non-circulating current type dual converter, if one converter is in the ON state then other converter is in the OFF state.

Advantage -

There is no circulating current b/w the converters.



1ϕ, $V_{o1} = \frac{2V_m}{\pi}\cos\alpha_1$ $\qquad$ 1ϕ $V_{o2} = \frac{2V_m}{\pi}\cos\alpha_2$

3ϕ, $V_{o1} = \frac{3V_{ML}}{\pi}\cos\alpha_1$ $\qquad$ 3ϕ $V_{o2} = \frac{3V_{ML}}{\pi}\cos\alpha_2$

V_o

CONV I → OFF $\quad$ CONV II → ON
$\alpha_2 > 90°$
$V_{o2}-$, V_o+

R.B. $\quad$ $T_a \to \phi I_o$ ∴ T_a+
$E_b \to \phi\omega$ ∴ E_b+
$-\omega$, $+T_a$

CONV I → ON $\quad$ CONV II → OFF
$\alpha_1 < 90°$
$V_{o1}+$, V_o+, E_b+, $\omega+$
I_a+, T_a+
I_o, $+T_a$, ω

V_o+ $\quad$ ($\phi-$) P− RB $\quad$ FM $\quad$ V_o+, I_o+ P+ ($\phi+$)

CONV I → OFF $\quad$ CONV II → ON $\quad$ CONV I → ON $\quad$ CONV II → OFF → I_o
$\alpha_2 < 90$ $\quad$ P+ V_o- RM $\quad$ FB V_o-, I_o+ P− ($\alpha_1 > 90$)
($\phi+$) I_o- $\quad$ ($\phi-$)

RM $\quad$ $V_{o2}+$, V_o- E_b- $\omega-$
I_o-, T_a- $\quad$ $-T_a$ $-\omega$

$V_{o1}-$, V_o-,
$T_a \to \phi I_a$ → ∴ $T_a = -ve$ $\quad$ $-T_a$ ω

Disadvantage -

→ It gives slow speed response & the reversal of armature current is not smooth during switching transition of the converters.

↓

We must provide commutation delay time (Δt_d) to the outgoing converter before the incoming converter is switched ON to avoid high circulating current during switching transitions of converter. This commutation delay time is responsible for slow speed response.

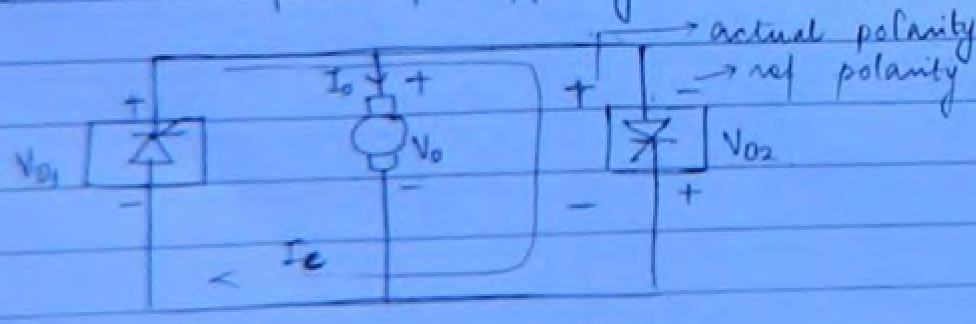
ii) Circulating Current Type -

Here, both the converters are simultaneously in the ON state.

Disadvantage -

There will be circulating current b/w the converters & hence responsible for additional power loss.

Circulating current is due to the v/g difference b/w the two converters. We can reduce the circulating current if the output v/gs of the converter are equal & opposing each other.



$I_c \downarrow \rightarrow \quad V_{01} = -V_{02} \qquad \alpha_1 + \alpha_2 = 180^\circ$

$$\frac{2V_m}{\pi} \cos\alpha_1 = -\frac{2V_m}{\pi} \cos\alpha_2$$

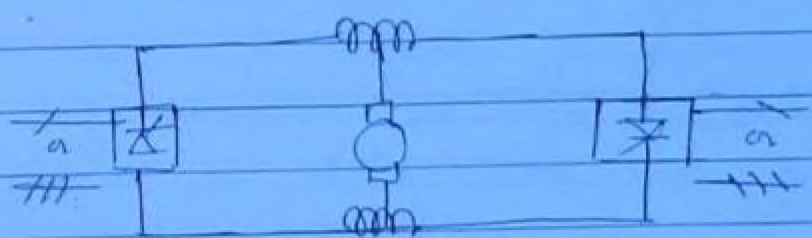
$$\cos\alpha_1 + \cos\alpha_2 = 0$$

$$\alpha_2 = 180 - \alpha_1$$

$$\alpha_1 + \alpha_2 = 180^\circ$$

Even after maintaining $\alpha_1 + \alpha_2 = 180^\circ$, still theres some circulating current due to the instantaneous v/g difference b/w the converters.

* To reduce this circulating current, we must connect a ~~reactive power~~ reactor core b/w the converters as shown.

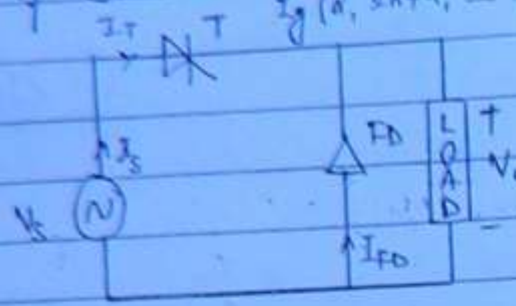


* In order to satisfy the relation $\alpha_1 + \alpha_2 = 180^\circ$ if one converter is operating in rectification mode then other converter must work in inversion mode.

Advantage -
It gives high speed response of the ~~as~~ reversal of armature current is smooth during switching transition of converters.

Effect of Source Inductance (L_s) on one Pulse Converter

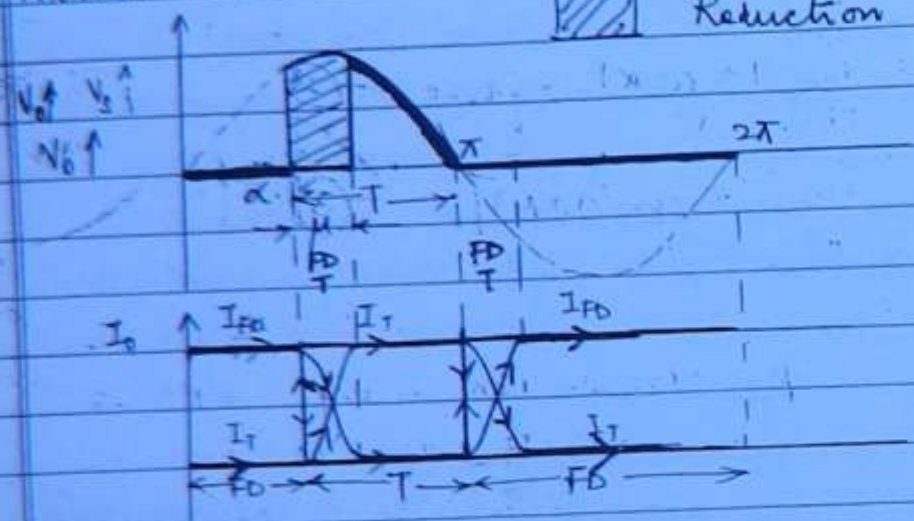
(130)



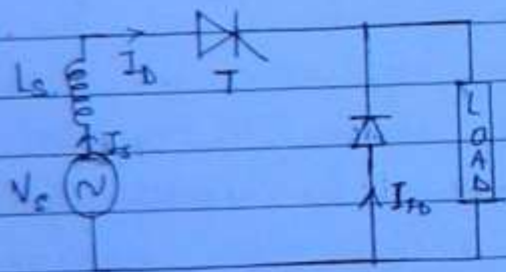
Assume highly inductive load

- Without L_s → $V_0 = \frac{V_m}{2\pi}(1+\cos\alpha)$

▨ Reduction in v/g due to L_s.



- With L_s



μ = overlap period

During overlap period both T & FD are ~~are~~ exchanging load current thus v/g = 0

During Overlap ⇒ FD & T → ON $V_0 = 0$

$$V_s = L_s \frac{d I_s}{dt}$$

$\alpha+\mu$ I_0

$$\frac{V_M}{2}[\cos\alpha - \cos(\alpha+\mu)] = \frac{\omega L_s I_o}{2}$$

divide to give avg reduction in v/g.

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$$\boxed{\Delta V_{do} = \frac{V_m}{2\pi}[\cos\alpha - \cos(\alpha+\mu)] = \frac{\omega L_s I_o}{2\pi} = f L_s I_o} \quad \text{—(1)}$$

ΔV_{do} = Avg reduction in V_o due to L_s.

$$V_o = \frac{V_m}{2\pi}(1+\cos\alpha) - f L_s I_o \quad \text{—(2)}$$

$$V_o = \frac{V_m}{2\pi}[1+\cos(\alpha+\mu)] \quad \text{—(3)}$$

** Overlap angle μ depends on firing angle α but avg reduction in v/g due to source inductance ie ΔV_{do} does not depend on firing angle α.

* Avg reduction in v/g due to source inductance depends upon frequency, L_s & I_o

* If $f\uparrow$ or $L_s\uparrow$ or $I_o\uparrow$ w/o changing V_s & α then μ also $\uparrow$.

If $V_s\uparrow$ w/o changing f, L_s I_o & α the $\mu\downarrow$ bcz due to $\uparrow$ in V_s height of pulse $\uparrow$. To maintain same area ΔV_{do} the width $\downarrow$. Thus $\mu\downarrow$

Effect of Source Inductance for Two Pulse Converter.

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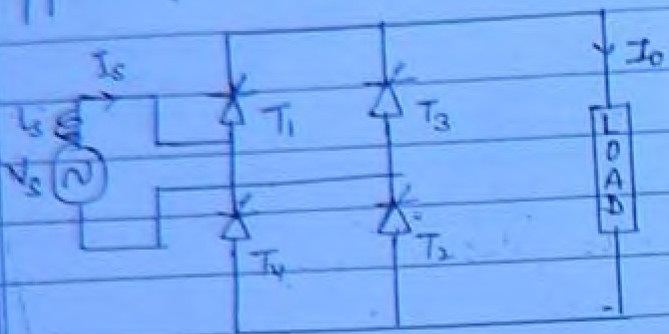
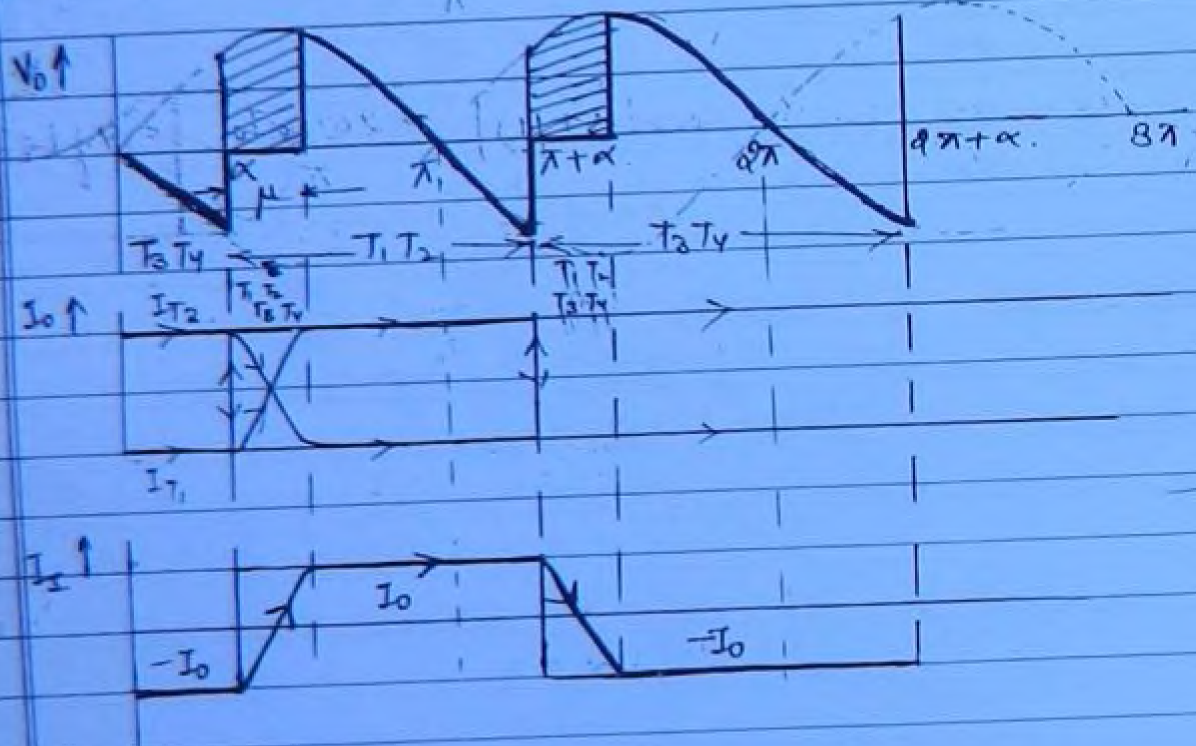


$V_{do} = \frac{2V_m}{\pi}$ (max dc o/p v/g)

$V_{o} = V_{do} \cos\alpha$

- Without $L_s \longrightarrow$



$T_1T_2 \rightarrow ON \quad I_{T_1} = I_{T_2} = I_o$

$T_3T_4 \rightarrow ON \quad I_{T_3} = I_{T_4} = I_o$

- with Ls.

During ~~ott~~ overlap period ~~v/g in 0~~ as all thyristors are conducting ∴ v/g is 0.

During μ T_1 T_2 & $T_3 T_4 \rightarrow ON$ $\therefore V_0 = 0$

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$$V_S = L_S \frac{dI_S}{dt}$$

$$\int_{\alpha}^{\alpha+\mu} V_m \sin \omega t \, d(\omega t) = \omega L_s \int_{-I_0}^{+I_0} dI_s$$

Wrong in P.S.B.

$$\frac{V_m}{\pi} [\cos \alpha - \cos(\alpha + \mu)] = \frac{2\omega L_s I_0}{\pi}$$

$$\boxed{\Delta V_{do} = \frac{V_m}{\pi} [\cos \alpha - \cos(\alpha + \mu)] = \frac{2\omega L_s I_0}{\pi}} \boxed{= 4 f L_s I_0} \quad - (1)$$

$$V_{do} = V_m \cdot \frac{2}{\pi} \Rightarrow \frac{V_{do}}{2} = \frac{V_m}{\pi}$$

$$\Delta V_{do} = V_{do} [\cos \alpha - \cos(\alpha + \mu)] = \frac{2\omega L_s I_0}{\pi}$$

$$V_0 = V_{do} \cos \alpha - 4 f L_s I_0 \quad - (2)$$

$$V_0 = \frac{V_{do}}{2} [\cos \alpha + \cos(\alpha + \mu)] \quad - (3)$$

From (1)

$$V_M [\cos \alpha - \cos(\alpha + \mu)] - 2\omega L_s I_0$$

$$I_0 = \frac{V_M}{2 \omega L_s} [\cos \alpha - \cos(\alpha + \mu)]$$

$$\underline{I_0} = \cos \alpha - \cos(\alpha + \mu) \quad - (4)$$

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Inductive Voltage Regulation –
Measure of reduction in v/g due to the source inductance.

$$\Rightarrow \frac{\Delta V_{do}}{V_{do}}$$

$$= \frac{\frac{V_{do}}{2}\left[\cos\alpha - \cos(\alpha+\mu)\right]}{V_{do}}$$

$$= \frac{\cos\alpha - \cos(\alpha+\mu)}{2}$$

ΔV_{do} does not depend on α.

At $\alpha = 0$ let $\mu = \mu_0$

Inductive V/g Reg $= \frac{\cos 0 - \cos(0+\mu_0)}{2}$

$$= \boxed{\frac{1-\cos\mu_0}{2}}$$

Effect of L_s on the performance of converter –

- Reduces avg output v/g of the converter
- It limits the range of α

$$\alpha_{max} = 180 - (\omega t_q + \mu_0)$$

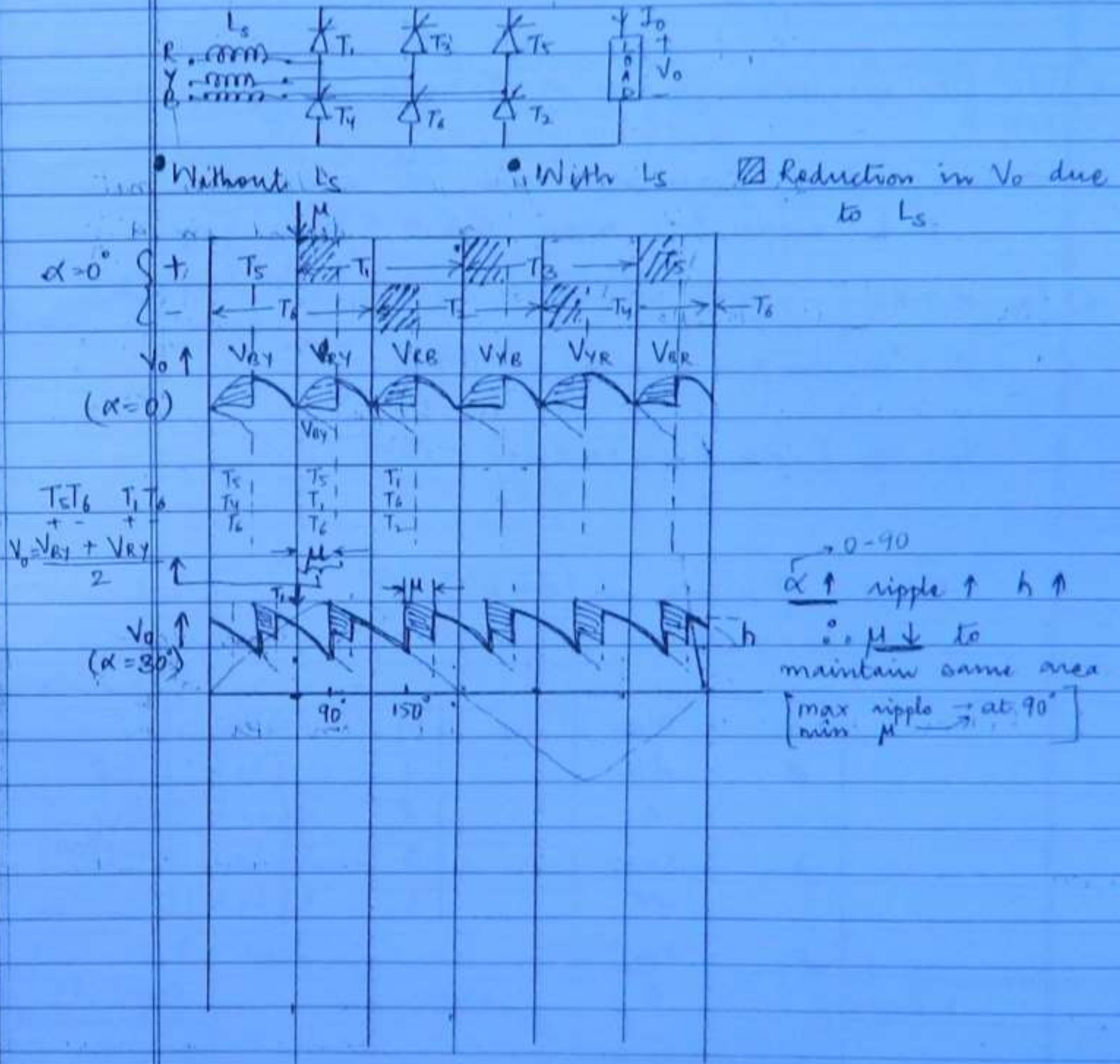
t_q = device turn off time

Disadv
- $DF = \cos\left(\alpha + \frac{\mu}{2}\right)$ due to L_s

Adv
- $\uparrow g$ = gives smoothness of waveform towards sine wave, as $g \uparrow$ waveform approaches towards sine wave
  $\therefore$ with L_s, $g \uparrow$ THD $\downarrow$
  Harmonics $\downarrow$ on AC side of converter.

Here the ↑ in g value is dominating the ↓ in FDF ∴ the PF is slightly ↑ 135

Effect of Source Inductance → 6 pulse converter



α ↑ (0-90) ripple ↑ h ↑

∴ μ ↓ to maintain same area

[max ripple → at 90°
min μ →]

$$\Delta V_{do} = \frac{V_{do}}{2}\left(\cos\alpha - \cos(\alpha+\mu)\right) = 6 f_s L_s I_o \quad \text{— (1)}$$

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* We get minimum μ at $\alpha = 90^\circ$ cuz we get maximum ripple.

* We get maximum μ at $\alpha = 0^\circ$ cuz ripple is minimum

⇒ When other parameters are ~~at~~ held constant
$\mu \downarrow$ ripple $\uparrow$ when $0 \leq \alpha \leq 90^\circ$

* for $\alpha > 90^\circ$ $\alpha \uparrow$ $\mu \uparrow$

$$V_o = V_{do}\cos\alpha - 6 f L_s I_o \quad \text{— (2)}$$

$$V_o = \frac{V_{do}}{2}\left[\cos\alpha + \cos(\alpha+\mu)\right] \quad \text{— (3)}$$

$$I_o = \frac{V_{ML}}{2\omega L_s}\left[\cos\alpha - \cos(\alpha+\mu)\right] \quad \text{— (4)}$$

m	V_{do}
2	$\frac{2V_m}{\pi}$
3	$\frac{3V_{ML}}{2\pi}$
6	$\frac{3V_{ML}}{\pi}$

m	ΔV_{do}
1	$f L_s I_o$
2	$4 f L_s I_o$
3	$3 f L_s I_o$
6	$6 f L_s I_o$

$\int_{-I_o}^{I_o}$

$\int_{0}^{I_o}$

INVERTERS

Fixed DC → Variable AC
(V_0 f_0)

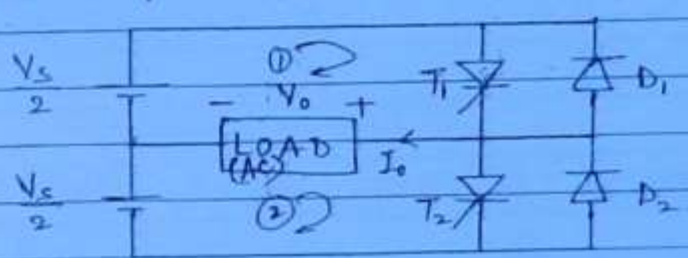
Classification of Inverters –

→ Voltage Source Inverters (VSI) → v/g waveform is independent of load; i waveform depends on load

→ Current Source Inverters (CSI) → opp of VSI

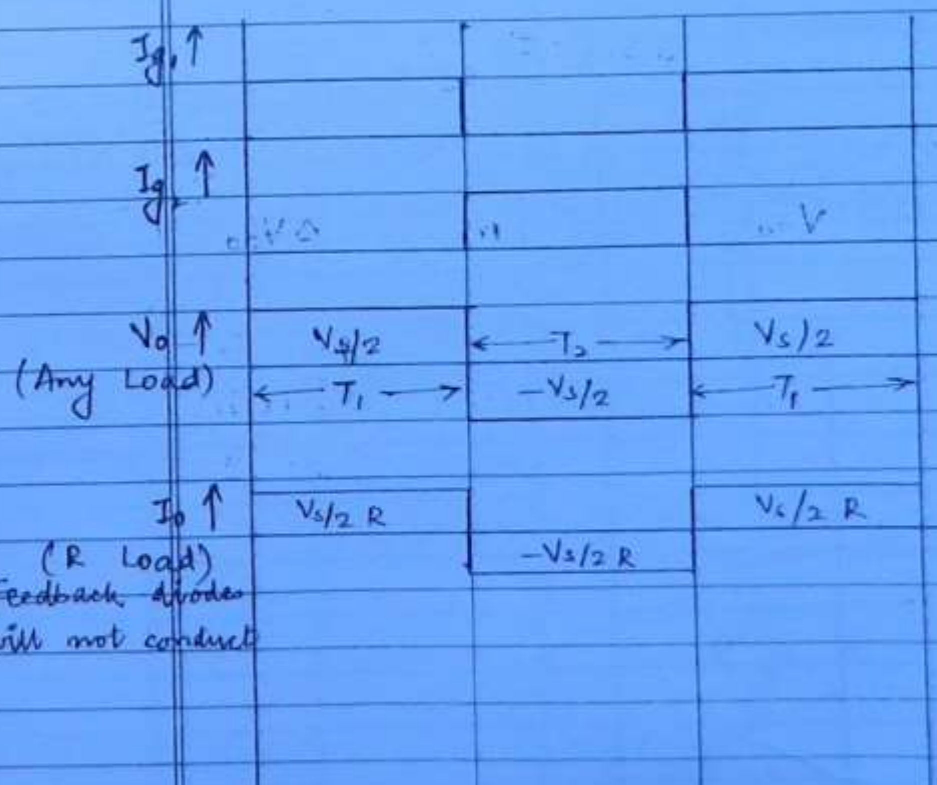
VSI

a) 1φ Half Bridge Inverter –



$\underline{T_1 \to ON}$ $V_0 = \frac{V_s}{2}$

$T_2 \to ON$ $V_0 = \frac{-V_s}{2}$



Feedback diodes will not conduct

→ Forced commutation is required

$$V_0 = \sum_{n=1,3,5..}^{\infty} \frac{4}{n\pi}\left(\frac{V_s}{2}\right) \sin n\omega t$$

$$\Rightarrow V_0 = \sum_{n=1,3,5..}^{\infty} \frac{2V_s}{n\pi} \sin n\omega t$$

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$$\Rightarrow V_{on} = \frac{2V_s}{n\pi} \sin n\omega t$$

$$(V_{on})_{rms} = \frac{\sqrt{2}\,V_s}{n\pi}$$

$$(V_{o1})_{rms} = \frac{\sqrt{2}\,V_s}{\pi}$$

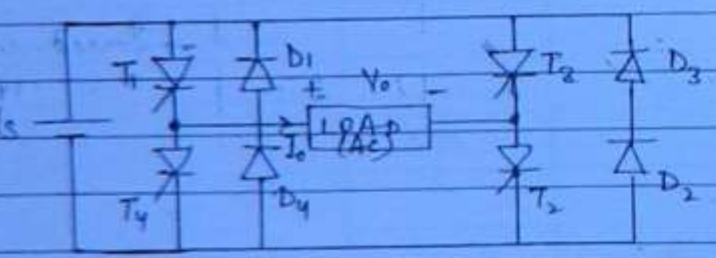
$$g = \frac{(V_{o1})_{rms}}{V_{on}} = \frac{\frac{\sqrt{2}\,V_s}{\pi}}{\frac{V_s}{2}} = \frac{2\sqrt{2}}{\pi}$$

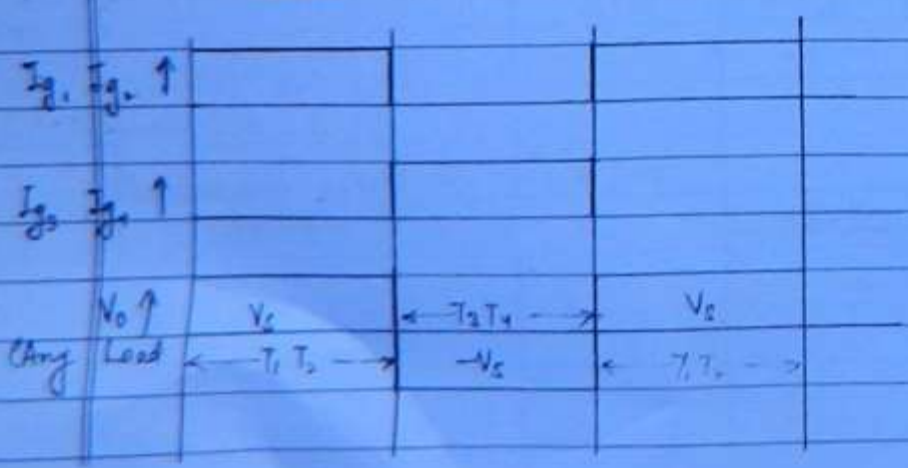
THD = 48.34 %

b) 1ϕ Full Bridge Inverter

$T_1 T_2 \to ON$ $\quad$ $T_3 T_4 \to ON$

$V_0 = V_s$ $\quad$ $V_0 = -V_s$





$$V_0 = \sum_{n=1,3,5\ldots}^{\infty} \frac{4V_s}{n\pi} \sin n\omega t$$

$$V_{0n} = \frac{4V_s}{n\pi} \sin n\omega t$$

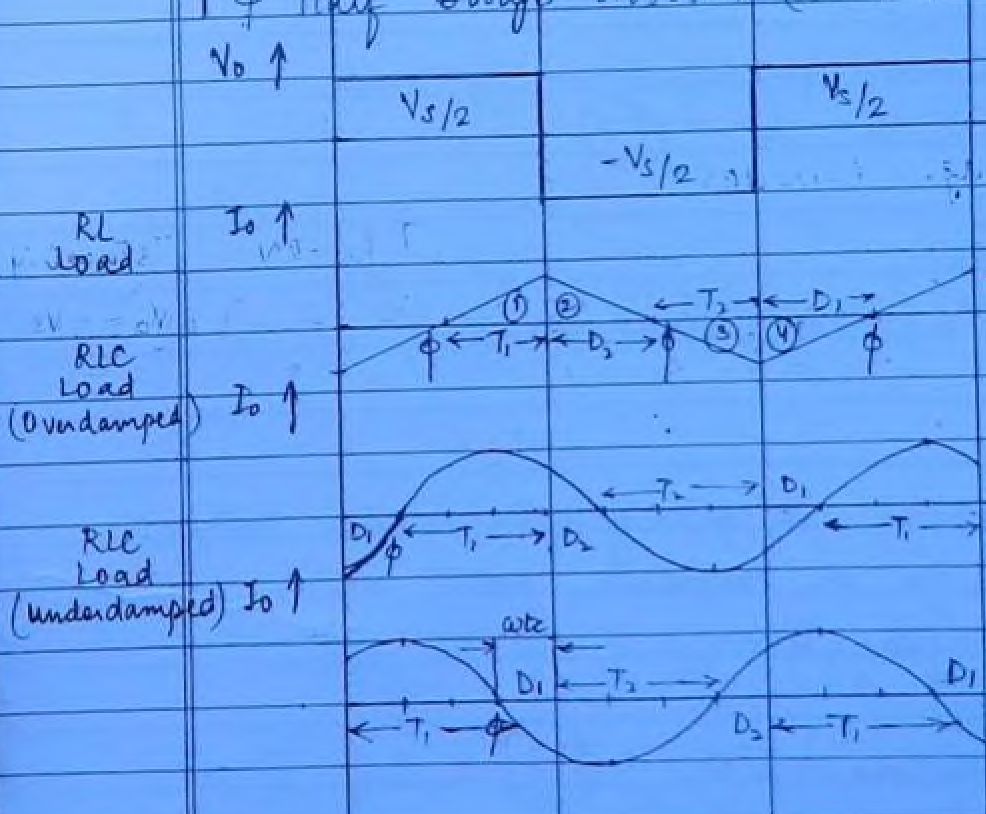
$$(V_{0n})_{rms} = \frac{2\sqrt{2}}{n\pi} V_s$$

$$(V_{01})_{rms} = \frac{2\sqrt{2}}{\pi} V_s$$

$$g = \frac{(V_{01})_{rms}}{V_{0R}} = \frac{\frac{2\sqrt{2}\,V_s}{\pi}}{V_s} = \frac{2\sqrt{2}}{\pi}$$

$$THD = 48.34\,\%$$

1 φ Half Bridge Inverter (Various Loads)



→ Forced commutation is required

→ Forced commutation is required

→ Forced commutation is not required (Load comm occurs)

for load comm $(X_C > X_L)$

Logics for 1 mark ques

① $T \rightarrow ON$ $P(+)$

② $D \rightarrow ON$ $P(-)$

③ $(T_1, D_1) \rightarrow ON$ $V_0(+)$

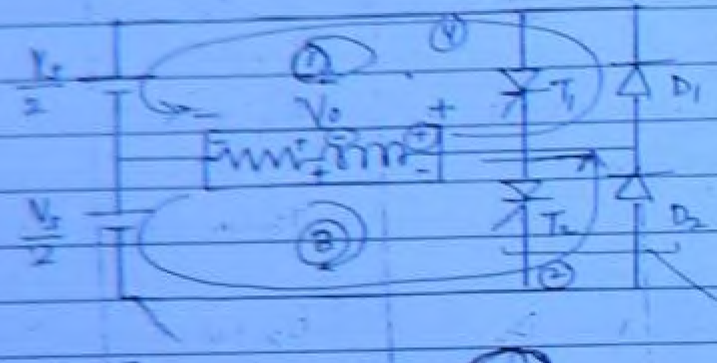
④ $(T_2, D_2) \rightarrow ON$ $V_0(-)$

I RL Load -

After reaching steady state

I_0 lags V_0 by $\phi = \tan^{-1}\frac{\omega L}{R}$

$I\ V_0$



① $T_1 \rightarrow ON$

$V_0 = V_s$ ie V_0+ I_0+ (P+)

$P_{source} \rightarrow P_{load}$

$(I^2R + \frac{1}{2}LI^2)$

Stores energy.

② $D_2 \rightarrow ON$ (P−)

$\frac{1}{2}LI^2 \rightarrow$ source $+ I^2R$

(Releasing Energy)

2 antiparallel devices cannot be ON at the same time.

eg: D_2 when ON provides RB to T_2 so T_2 is OFF

③ $T_2 \rightarrow ON$

(D_2 releases reverse polarity thus T_2 starts conducting)

$V_0 = -V_s$ ie V_0- I_0- (P+)

$P_{source} \rightarrow P_{load}$

$(I^2R + \frac{1}{2}LI^2)$ Store Energy

④ $D_1 \rightarrow ON$

Inductor reduces its polarity to release energy searching for a favourable path which is achieved from D_1

$\frac{1}{2}LI^2 \rightarrow$ source $+ I^2R$ (P−)

(Releasing Energy)

* Forced commutation is required for RL Load.

Switching Logic Table

V_0	I_0	P	½ Bridge	Full Bridge
			Device (ON)	
$\overset{+}{T_1\ D_1}$	+	$\overset{+}{T_1\ T_2}$	T_1	$T_1\ T_2$
$\overset{+}{T_1\ D_1}$	−	$\overset{-}{D_1\ D_2}$	D_1	$D_1 D_2$
$\overset{-}{T_2\ D_2}$	−	$\overset{+}{T_1\ T_2}$	T_2	$T_3 T_4$
$\overset{-}{T_2\ D_2}$	+	$\overset{-}{D_1\ D_2}$	D_2	$D_3 D_4$

for all leading loads it starts with **T**

for all lagging loads it starts with **D**

II RLC (Overdamped)

$$X_L > X_C$$

I_0 lags V_0 by $\phi = \tan^{-1}\left(\frac{X_L - X_C}{R}\right)$

III RLC (Underdamped)

$$X_C > X_L$$

I_0 lags V_0 by $\phi = \tan^{-1}\left(\frac{X_L - X_C}{R}\right)$

or I_0 leads V_0 by $\phi = \tan^{-1}\left(\frac{X_C - X_L}{R}\right)$

$$\omega t_c = \phi$$

$$t_c = \frac{1}{\omega} \tan^{-1}\left(\frac{X_C - X_L}{R}\right)$$

* If $t_c < t_d$ then load commutation fails
  ($\therefore$ forced commutation is required)

$$\boxed{V_{on} = \frac{\overset{\to 4 \text{ for full bridge}}{2} V_s}{n\pi} \sin n\omega t \quad (\text{Any Load})}$$

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Consider an RLC load

$$I_{on} = \frac{V_{on}}{Z_n} \qquad Z_n = R + j(X_{Ln} - X_{cn})$$

$$X_{Ln} = n\omega L \qquad X_{cn} = \frac{1}{n\omega C}$$

n^{th} harmonic inductive impedance — n^{th} harmonic capacitive impedance

$$Z_n = |Z_n| \angle \phi_n$$

$$|Z_n| = \sqrt{(R)^2 + (X_{Ln} - X_{cn})^2}$$

$$\phi_n = \tan^{-1}\left(\frac{X_{Ln} - X_{cn}}{R}\right)$$

n^{th} harmonic impedance angle
(or) displacement angle.

$$I_{on} = \frac{V_{on}}{Z_n} = \frac{V_{on}}{|Z_n| \angle \phi_n} = \frac{V_{on}}{|Z_n|} \angle -\phi_n$$

$$\boxed{I_{on} = \frac{\overset{\to 4 \text{ for full bridge}}{2} V_s}{n\pi |Z_n|} \sin(n\omega t - \phi_n)} \rightarrow \text{for RLC load.}$$

eg for pure inductive load

$|Z_n| = n\omega L$

$\phi_n = 90°$

$I_{on} = \frac{2V_s}{n\pi\, n\omega L} \sin(n\omega t - 90°) \Rightarrow I_{on} \propto \frac{1}{n^2}$

If n^{th} harmonic is inv. prop. to n^2 shape

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3ϕ VSI

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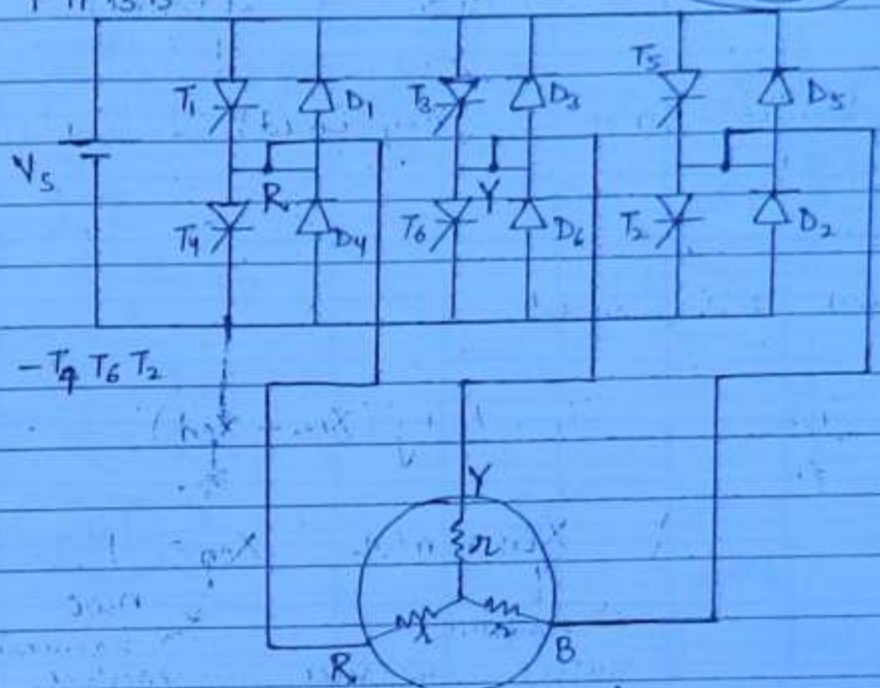
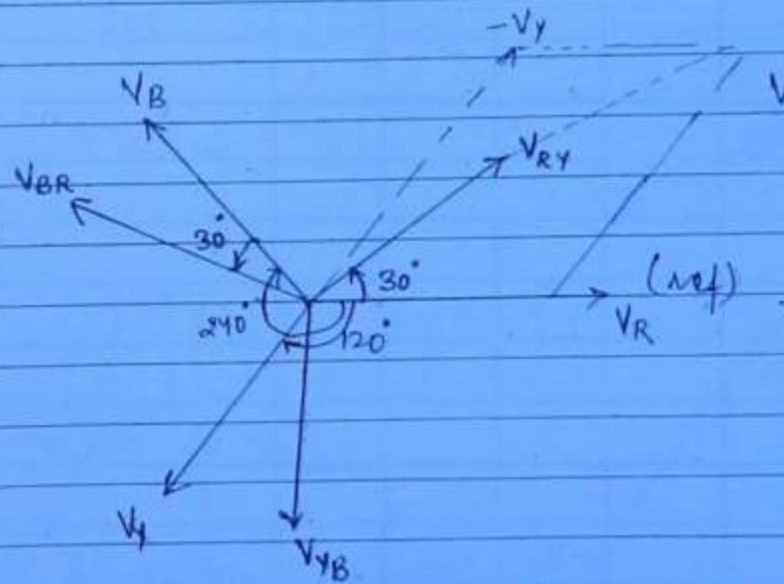
a) 180° MODE

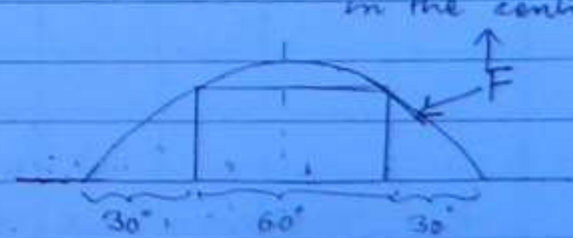


3ϕ Y connected R Load

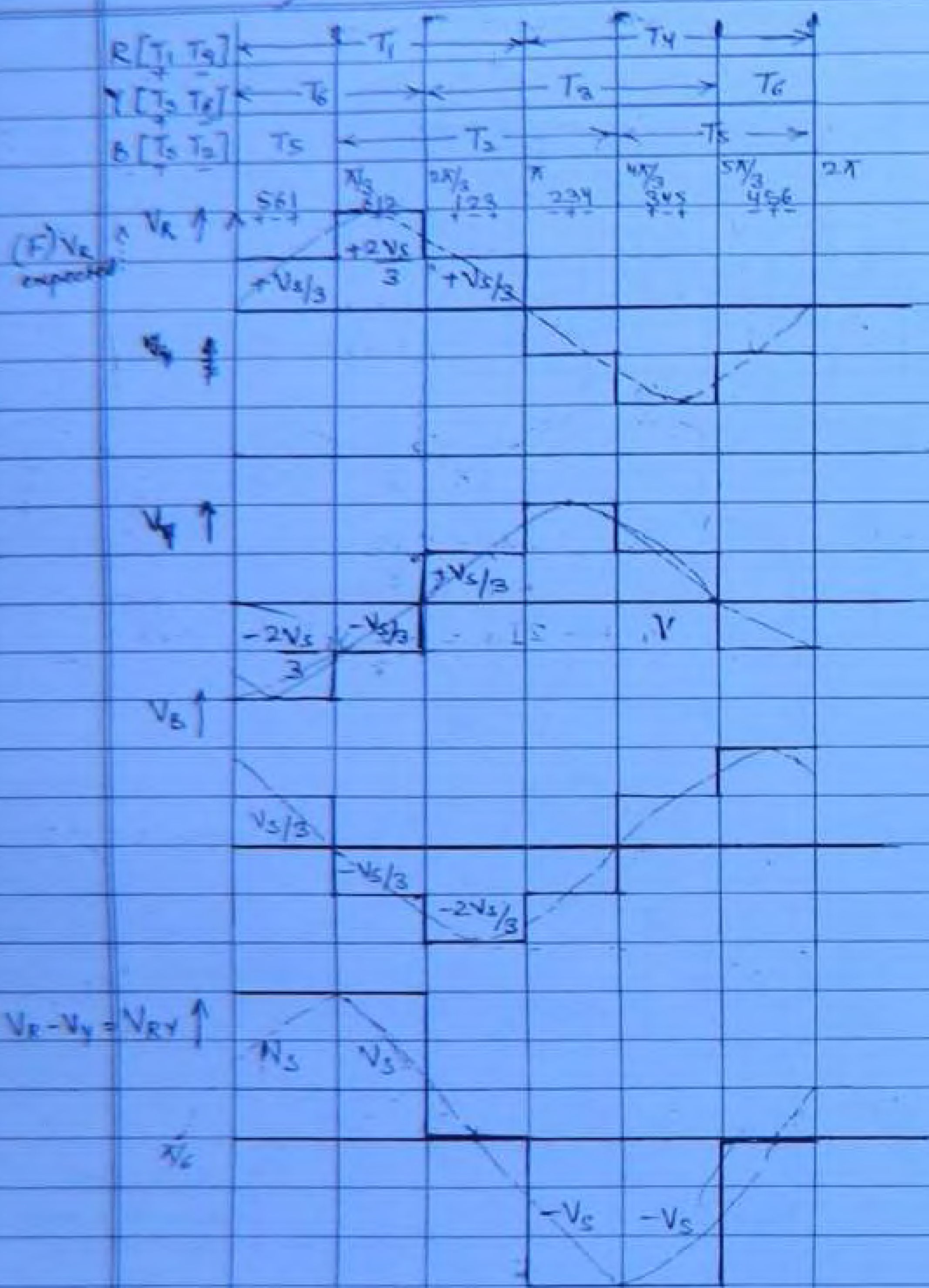


$V_{RY} = V_R - V_Y$

(fundamental inserts the pulse in the centre)



Switching pattern in timing diagram.



S 6 1
+ − +

this −ve share current of 2 +ve phase
so its double
ie Y φ V & I are double of R & B.
since its −ve its polarity is −ve
so 5 → B → $+\frac{V_s}{3}$

6 → Y → $-\frac{2V_s}{3}$

1 → R → $+\frac{V_s}{3}$

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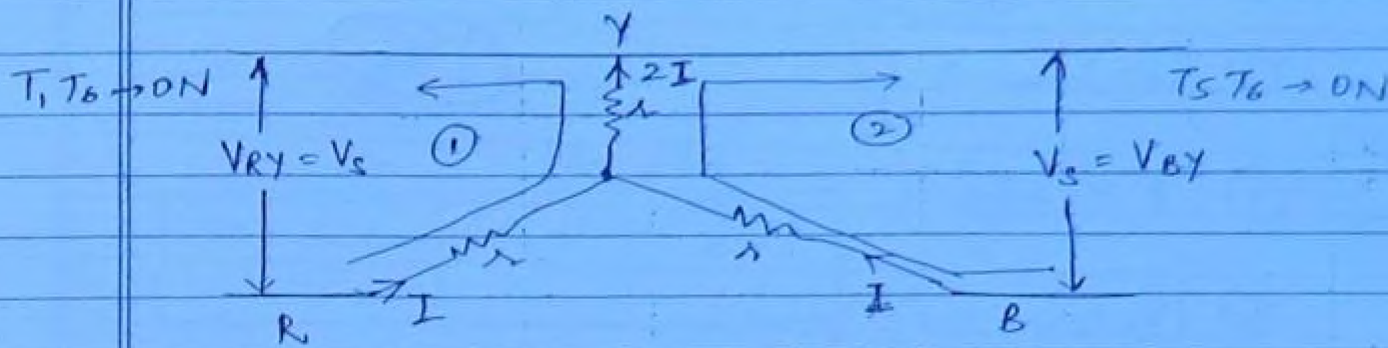
The waveforms are same for any load Δ connected or λ connected, DC motor, induction motor etc

∴ VSI v/s waveform is independent of load

$\boxed{0 \text{ to } \frac{\pi}{3}}$: $\underbrace{T_5\, T_6\, T_1}$

$T_1 T_6$ (+ −) $\qquad$ $T_5 T_6$ (+ −)

J45



$V_S = Ir + 2Ir$

$V_S = 3Ir$

$Ir = \frac{V_S}{3}$

$V_R = +Ir = +\frac{V_S}{3}$

$V_Y = -2Ir = -\frac{2}{3} V_S$

$V_B = +Ir = +\frac{V_S}{3}$

$$V_{RY} = V_S \left(\frac{2\pi/3}{\pi} \right)^{1/2}$$

$$(V_L)_{rms} = V_S \sqrt{\frac{2}{3}}$$

$$V_R = \left\{ \frac{1}{\pi} \left[\left(\frac{V_S}{3}\right)^2 \frac{\pi}{3} + \left(\frac{2V_S}{3}\right)^2 \frac{\pi}{3} + \left(\frac{V_S}{3}\right)^2 \frac{\pi}{3} \right\} \right.$$

$$V_{ph} = \frac{\sqrt{2}}{3} V_S$$

$$I_L = I_{ph} = \frac{V_{ph}}{r} = \frac{\sqrt{2}\, V_S}{3\, r}$$

$$(I_T)_{rms} = \frac{I_{ph}}{\sqrt{2}} \qquad V_L = \sqrt{3}\, V_{ph}$$

①
$$\boxed{V_{ph} = \frac{\sqrt{2}}{3} V_s}$$

② $I_{ph} = \dfrac{V_{ph}}{r}$

③ $(I_T)_{rms} = \dfrac{I_{ph}}{\sqrt{2}}$

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④ $P = 3 I_{ph}^2 r = 3 \dfrac{V_{ph}^2}{r}$

⑤ $(V_L)_{rms} = \sqrt{3}\, V_{ph}$

fourier series for R phase v/g waveform.

$$\boxed{V_R = \sum_{n=6k\pm1}^{\infty} \frac{2V_s}{n\pi} \sin n\omega t}$$

NOTE : Even & triple harmonics are absent

$$V_{Rn} = \frac{2V_s}{n\pi} \sin n\omega t$$

$$(V_{Rn})_{rms} = \frac{\sqrt{2}\, V_s}{n\pi} \qquad (V_{R1})_{rms} = \frac{\sqrt{2}\, V_s}{\pi}$$

$$g = \frac{V_{R1}}{V_R} = \frac{\cancel{\sqrt{2}}\, \cancel{V_s}}{\pi} \Big/ \frac{\cancel{\sqrt{2}}\, \cancel{V_s}}{3} = \frac{3}{\pi}$$

$$\boxed{g = \frac{3}{\pi}}$$

THD = 31 %

Let us consider, RLC load in each phase.

$$V_{Rn} = \frac{2V_s}{n\pi} \sin n\omega t$$

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$$I_{Rn} = \frac{V_{Rn}}{Z_n} = \frac{V_{Rn}}{|Z_n| \angle \phi_n}$$

$$\boxed{I_{Rn} = \frac{2V_s}{n\pi |Z_n|} \sin(n\omega t - \phi_n)}$$

eg. I.M $\quad |Z_n|$ per phase $= \sqrt{R_1^2 + (n\omega L_1)^2}$

$$\phi_n = \tan^{-1} \frac{X_{Ln}}{R}$$

for $n = 1$ (F)

Diode conducts for $\phi = \tan^{-1} \frac{X_L}{R}$ for RL load

& $\phi = \tan^{-1} \frac{X_C - X_L}{R}$ for RLC load

—— x ——

Line v/g.

$$V_{RY} = \sum_{\substack{n=1,3,5 \\ n=6k\pm1}}^{\infty} \frac{4V_s}{n\pi} \sin \frac{n\pi}{3} \sin n\left(\omega t + \frac{\pi}{6}\right)$$

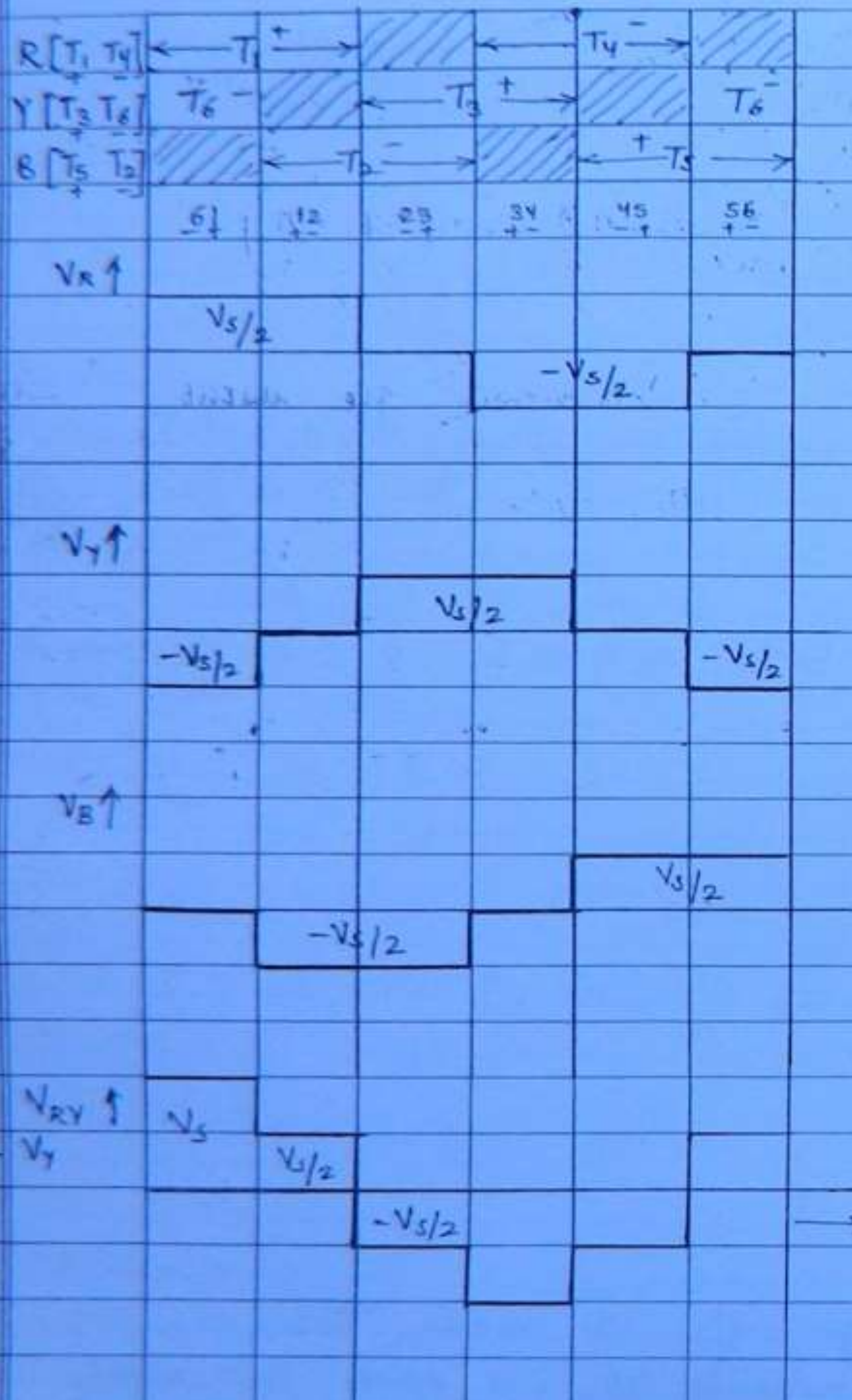
$$g = \frac{3}{\pi} \qquad THD = 31\%$$

Disadvantage of 180° mode VSI

Theres a possibility of S.C across the supply when incoming thyristor start conducting before the outgoing thyristor belonging to the same phase stops conducting

IES CONV

b) 120° mode → In 120° mode VSI we are alloting a conduction angle of 120° for each thyristor & the last 60° is alloted for commutation



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for Δ connection V_{RY} V_{YB} V_{BR} → displace this by 120° & 240° for V_{YB} & V_{BR}

$$(V_R)_{rms} = \frac{V_s}{2}\left(\frac{2\pi/3}{\pi}\right)^{1/2}.$$

$$= \frac{V_s}{2}\sqrt{\frac{2}{3}}$$

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$$(V_R)_{rms} \boxed{\frac{V_s}{\sqrt{6}} = V_{ph}}$$

phase v/g $\quad V_R = \sum_{\substack{n=6k\pm1 \\ n=1,3,5\ldots}}^{\infty} \frac{2V_s}{n\pi} \sin n\frac{\pi}{3} \sin n\left(\omega t + \frac{\pi}{6}\right)$

NOTE : Even & triple harmonics are absent

$g = \frac{3}{\pi} \qquad THD = 31\%$

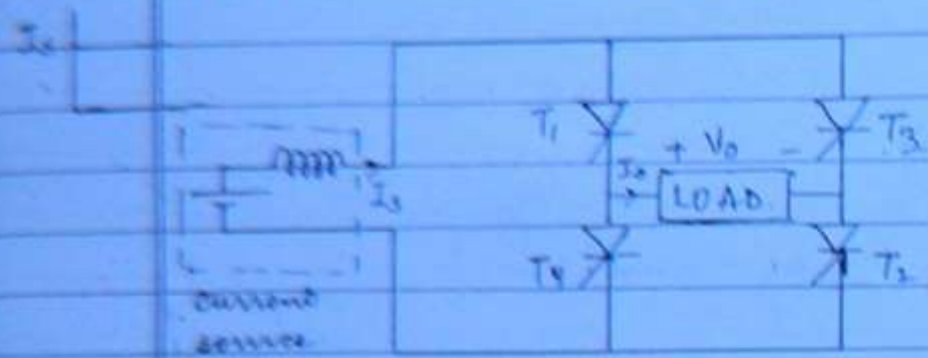
line v/g $\quad V_{RY} = \sum_{n=6k\pm1}^{\infty} \frac{3V_s}{n\pi} \sin n\left(\omega t + \frac{\pi}{3}\right)$

NOTE : Even & triple harmonics are absent

$g = \frac{3}{\pi} \qquad THD = 31\%$

for 120° & 180° mode
phase or line v/g
g & THD are same.

CSI



$T_1\, T_2 \rightarrow$ ON

$I_o = I_s$

$T_3\, T_4 \rightarrow$ ON

$I_o = -I_s$

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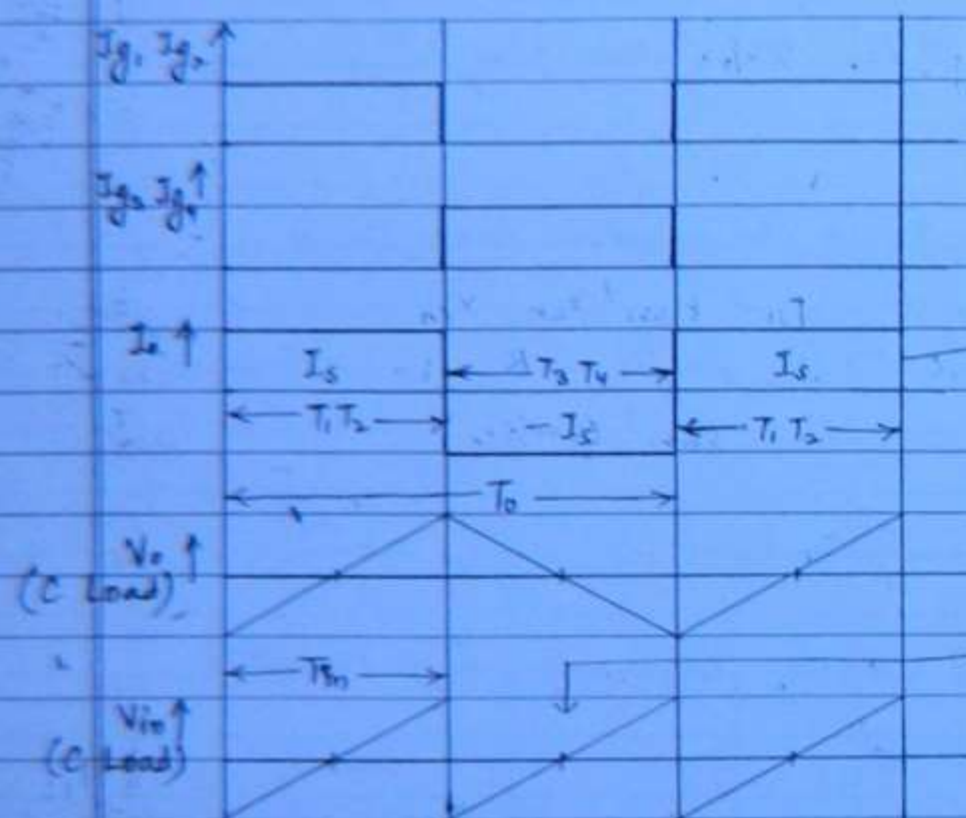

$I_{on} = I_s$ any load (CSI)

where $T_3\, T_4$ is ON V_{in} is $-$ve of V_0

$$I_0 = \sum_{n=1,3,5..}^{\infty} \frac{4 I_s}{n\pi} \sin n\omega t$$

$$g = \frac{2\sqrt{2}}{\pi} \qquad \text{THD} = 48.34\,\%$$

$$I_{on} = \frac{4 I_s}{n\pi} \sin n\omega t \quad [\text{for any load}]$$

Let us consider an RLC load -

$$V_{on} = I_{on} \cdot Z_n$$
$$= I_{on} |Z_n| \angle \phi_n$$

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$$V_{on} = \frac{4 I_s}{n\pi} |Z_n| \sin(n\omega t + \phi_n)$$

$$Z_n = R + j(X_{Ln} - X_{Cn})$$

For C load

$$|Z_n| = \frac{1}{n\omega C} \qquad \phi_n = \tan^{-1} \frac{X_{Ln} - X_{Cn}}{R}$$
$$= \tan^{-1} \frac{0 - X_{Cn}}{0}$$
$$\phi_n = -90^\circ$$

$$V_{on} = \frac{4 I_s}{n\pi} \left(\frac{1}{n\omega C} \right) \sin(n\omega t - 90^\circ)$$

$$V_{on} \propto \frac{1}{n^2}$$

$$T_0 = 2 T_{in}$$

$$\therefore \boxed{f_{in} = 2 f_0}$$

$f_{in} \rightarrow$ DC, $f_0 \rightarrow$ AC

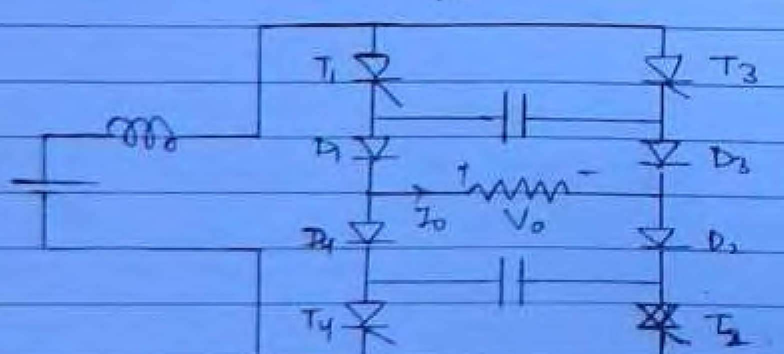
Advantages of CSI -

1. Feedback diodes are not required in CSI 152
2. Commutation is simple
3. For C loads theres a possibility of load commutation
4. Inherently theres a short ckt protection for the source when the incoming thyristors are switched on before the outgoing thyristor becomes off due to the presence of high inductance.

Disadvantage of CSI -

1. The commutating element (along with the load) applies high reverse v/g across the power device used in CSI ∴ the devices having low reverse v/g blocking capability such as GTO, IGBT & other transistors are not generally prefered in CSI. Here we prefer SCR because it has high reverse v/g blocking capability.

2. If commutating capacitor is directly connected across the load, then it will be continuously discharging through the load. To avoid it we must connect the diode as shown in figure.



CWB chapter 2

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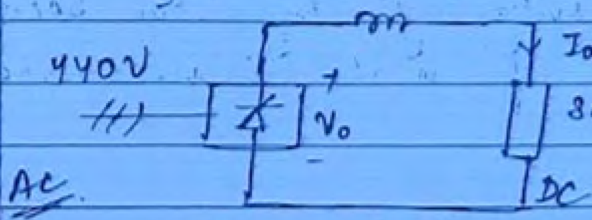
(5) $\alpha = 60$ $\quad \underset{(IDF)}{FDF} = \cos\alpha = \cos 60 = 0.5$

$PF = g\,(FDF)$

$= \frac{3}{\pi}(\cos\alpha) = \frac{3}{\pi}\cos 60 = 0.476$ (c)

(6) 440V



$E = 350$
$R = 0.5\Omega$

$P_{AC} \xleftarrow[(\alpha > 90)]{INV} P_{DC}$

$V_0 = -E + I_0R$

$= -350 + 10 = -340$

$\frac{3V_{ML}}{\pi}\cos\alpha = -E + I_0R$

$3 \cdot \frac{440\sqrt{2}}{\pi}\cos\alpha = -340$

$\alpha = 125°$

for $\alpha \geq 60°$ $\quad \omega t_c = \pi - \alpha = 180 - 125°$

$= 55°$ (d)

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5

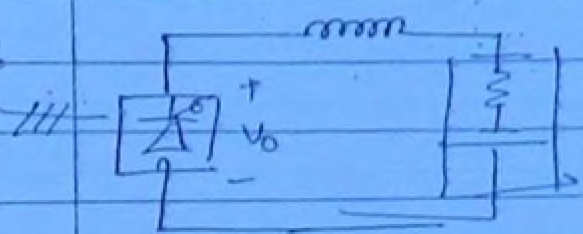
$\alpha = 60^\circ$, $FDF = \cos\alpha$

$= \cos 60 = 0.5$

$PF = g\,(FDF) = \frac{3 \times 0.5}{\pi}$

$= 0.478$

6



$P_{AC} \leftarrow P_{DC}$

inversion

$\alpha > 90^\circ$

$V_0 = -E + I_0 R$

$\cos\alpha \frac{3V_{mL}}{\pi \cos\alpha} = -350 + (20 \times 0.5)$

$\frac{3 \times 440\sqrt{2}}{\cos \pi} \cos\alpha = -350 + 10$

$\cos\alpha \Rightarrow \alpha = 125^\circ$

I $\alpha < 60^\circ$, $\omega t_c = 4\pi/3 - \alpha$

II $\alpha > 60^\circ$, $\omega t_c = \pi - \alpha/3$

$= 180^\circ - 125$

$= 55^\circ$

VOLTAGE Control of Inverter

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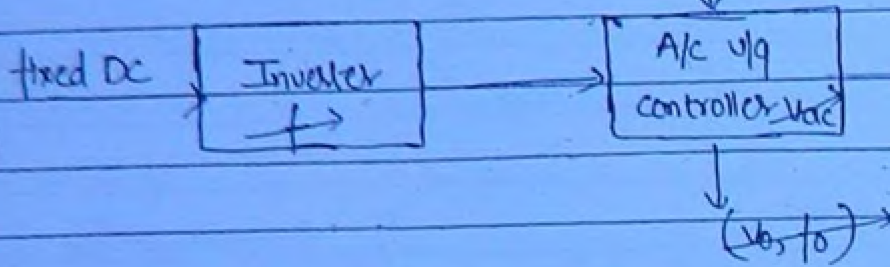
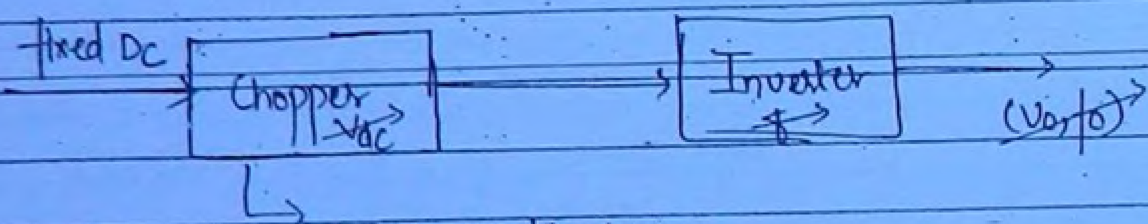
I. External control

It is like a i.e. change in v/g who change in frequency.

a)

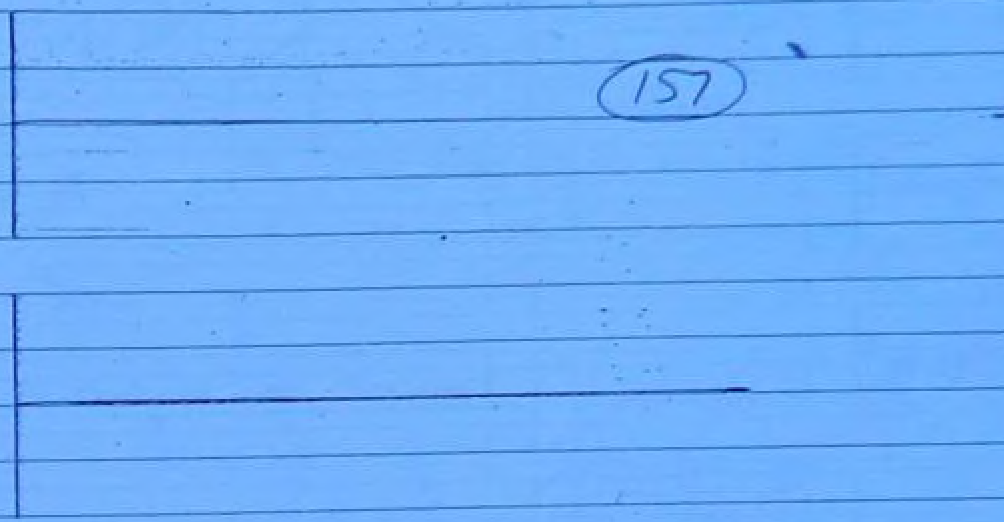


b)



DC level will change.

II Internal v/g control using PWM technique :-



Advantages of PWM technique :-

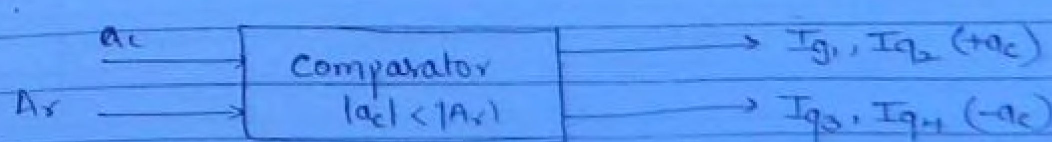
1. We can get variable v/g within the inverter without increasing no. of stages.

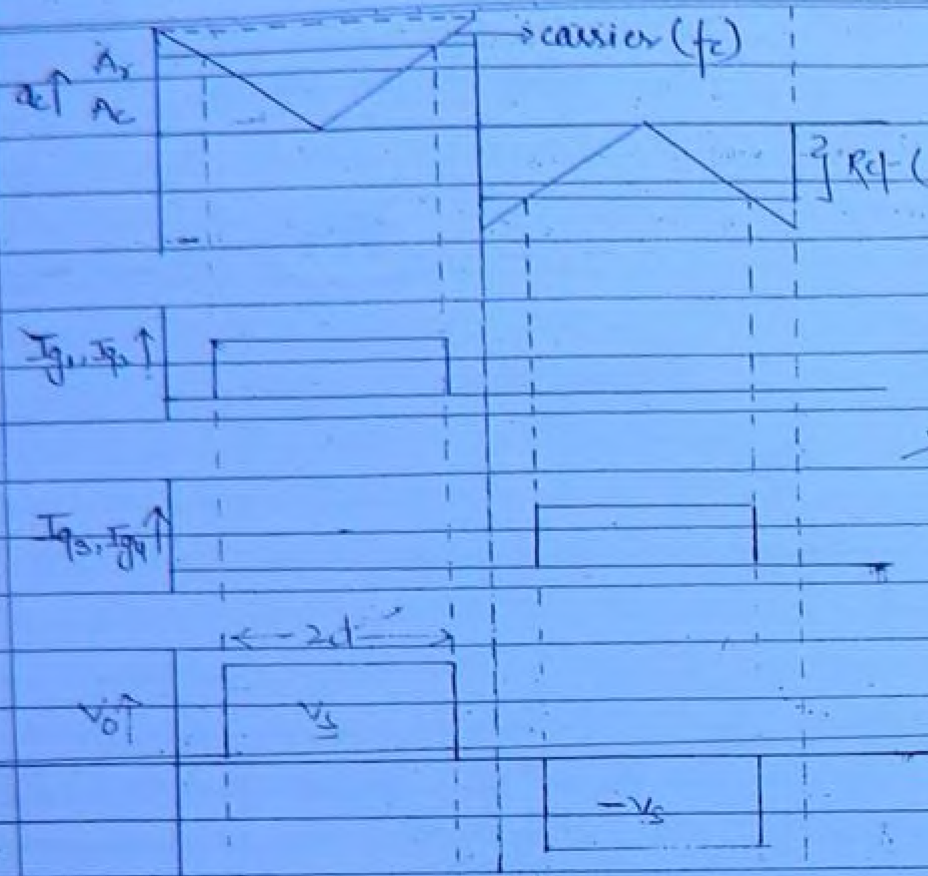
2. We can eliminate some of the lower order harmonics (higher order harmonics can be easily filtered).

Types of PWM techniques :-

1. Single PWM technique :-

Let us realise this modulation technique using full bridge inverter.





$2d \rightarrow$ Total pulse width

$$V_{rms} = V_s \left(\frac{2d}{\pi}\right)^{1/2}$$

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Fourier series for o/p v/g waveform :-

$$V_0 = \sum_{n=1,3,5}^{\infty} \frac{4V_s}{n\pi} \sin n\omega t \cdot \sin \frac{n\pi}{2} \sin nd$$

$$V_{on} = \frac{4V_s}{n\pi} \sin \frac{n\pi}{2} \cdot \sin nd \cdot \sin n\omega t$$

$$V_{on} = 0 \quad \text{if } nd = \pi, 2\pi, 3\pi, \ldots$$

$$d = \frac{\pi}{n}, \frac{2\pi}{n}, \frac{3\pi}{n}, \ldots$$

$2d = \frac{2\pi}{n}, \frac{4\pi}{n}, \frac{6\pi}{n} \longrightarrow$ valid only if $2d < \pi$

$\rightarrow$ condition to eliminate nth harmonic

To eliminate 3rd harmonic

$$2d = \frac{2\pi}{3}, \underbrace{\frac{4\pi}{3}, \frac{6\pi}{3}, \ldots}_{>\pi}$$

$\therefore$ since it is less than π

$$2d = \frac{2\pi}{3} = 120^\circ$$

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To eliminate 5th harmonic

$$2d = \underbrace{\frac{2\pi}{5}, \frac{4\pi}{5}}, \frac{6\pi}{5}, \ldots$$

both less than π

$$2d = 72^\circ, 144^\circ$$

$$V_{on} = \frac{4V_s}{n\pi}\left|\sin\left(\frac{n\pi}{2}\right)\right| \sin nd \cdot \sin n\omega t$$

for $n = 1, 3, 5 \ldots$ it is ± 1, for rms value calculation it is always 1

$$(V_{on})_{rms} = \frac{4V_s}{n\pi}\frac{\sin nd}{\sqrt{2}} = \frac{2\sqrt{2}V_s}{n\pi}\sin nd$$

$$\boxed{\begin{aligned} (V_{o1})_{rms} &= \frac{2\sqrt{2}V_s}{\pi}\sin d \\ g &= \frac{(V_{o1})_{rms}}{V_{or}} = \frac{\frac{2\sqrt{2}V_s}{\pi}\sin d}{V_s\left(\frac{2d}{\pi}\right)^{1/2}} \\ g &= \frac{2\sqrt{2}\cdot \sin d}{\sqrt{(2d)\cdot\pi}} \\ THD &= \left(\frac{1}{g^2} - 1\right)^{1/2} \end{aligned}}$$

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1

$2\alpha = \alpha = 120^\circ$

$V_s = 1V$

$$(V_{01})_{rms} = \frac{2\sqrt{2}V_s}{\pi} \sin\alpha$$

$$= \frac{2\sqrt{2}}{\pi} \sin 60^\circ$$

$$= 0.78$$

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2. $2\alpha = 72^\circ$ or 144°

3

$2\alpha = 144^\circ$

$$(V_{03})_{rms} = \frac{2\sqrt{2}V_s}{3\pi} \sin 3\alpha$$

$$(V_{01\,max}) = \frac{2\sqrt{2}V_s}{\pi}$$

$$\frac{(V_{03})_{rms}}{(V_{01})_{max}} = \frac{\sin 3\alpha}{3}$$

$$= 19.6\%$$

5

$2\alpha = 150^\circ$

$$(V_{01})_r = \frac{2\sqrt{2}V_s}{\pi} \sin\alpha$$

$$g = \frac{2\sqrt{2} \cdot \sin\alpha}{\sqrt{(2\alpha) \cdot \pi}}$$

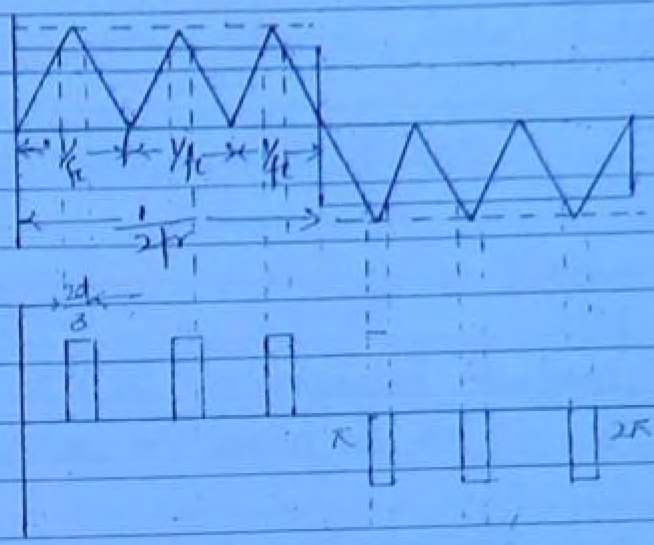
$$= 0.95$$

$THD = 0.32$

$= 31.83\%$

Multiple pulse PWM technique :-

(161)



$$\frac{3}{f_c} = \frac{1}{2f_r}$$

$$3 = \frac{f_c}{2f_r}$$

$2d \longrightarrow$ Total PW in each $1/2$ cycles

$$V_{or} = V_s\left(\frac{2d}{\pi}\right)^{1/2}$$

$$\boxed{n = \frac{f_c}{2f_r}}$$

Pulse width (Pulse length) $= \dfrac{2d}{N} = \left(1 - \dfrac{V_r}{V_c}\right)\dfrac{\pi}{N}$

→ Height of pulse is decided by supply, Only by changing supply, height of the pulse can be varied.

III. Sinusoidal PWM technique

In sinusoidal PWM technique, the reference signal is taken as sine waveform.

Here we've two cases :-

I case :- Peak value of carrier coincident with zero of Ref. signal

II case :- Zero of carrier co-incident with zero of reference.

Case-1 :- $N = \frac{f_c}{2f_r}$, Case-2 :- $N = \frac{f_c}{2f_r} - 1$

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NOTE :-

Dominant harmonics = $2N \pm 1$.

$N \longrightarrow$ No. of pulses in each half cycle

for $N = 3$, Dominant harmonic $= 6 \pm 1 = 5, 7$-th.

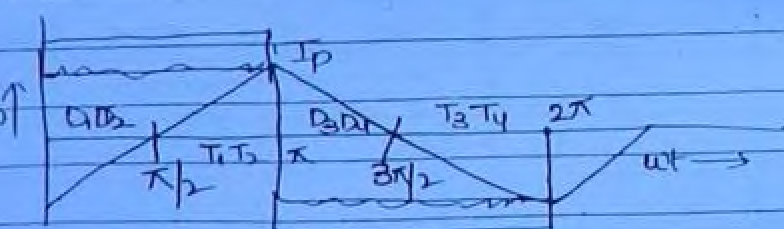
for $N = 9$, " " $= 18 \pm 1 = 17, 19$-th.

→ If Domination is shown by lower order, it is difficult to filter them, we require big size filter to remove lower order harmonics.

→ If Domination is shown by higher order, we can easily filter them.

In sinusoidal PWM tech., we ↑ the No. of pulses in each half cycle & ↑ the order of dominant harmonics so that they can be easily filtered.

7 $V_{0r} = \frac{\sqrt{2} V_s}{3\pi}$



$\frac{\pi}{2}$ to $\pi \longrightarrow V_0 = V_s = L\frac{dI}{dt} = V_s$

$\omega\, di = \frac{\omega V_s}{L} dt \Rightarrow \omega\, di = \frac{V_s}{L} d(\omega t)$

$$\int_{0}^{I_p} di = \frac{V_s}{\omega L} \int_{\pi/2}^{\pi} d(\omega t)$$

$$i_p = \frac{200\cancel{\pi}}{100\pi \times 0.1} \times \frac{\pi}{\cancel{\pi}}$$

$$= 10A$$

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13 Since, Inductor $\longrightarrow \phi = 90°$

$$\omega t_c = \phi$$

$$t_c = \frac{\phi}{\omega} = \frac{90°/180° \times \pi}{2\pi f} = 5ms$$

12 3ϕ VSI

$\longrightarrow$ pure inductive Load, $|Z_n| = n\omega L$

$$V_{on} = \alpha V_{a} \quad (\alpha_n < 1)$$

$$I_{on} = \frac{\alpha V_{01}}{|Z_n|} = \frac{\alpha_n}{n} \frac{V_{01}}{\omega L}$$

$$\frac{0.5 \times \pi}{2\pi f \times 5c}$$

$$= \frac{\alpha_n}{n} I_{01}$$

1. $R = 3\Omega$, $X_L = 12\Omega$, $X_C = ?$

$$f = \frac{10^3}{0.2} = \frac{10^3 \times 10}{2} = 5000\,Hz$$

$$t_q = 12 \times 10^{-6} s, \quad SF = 2$$

$$t_c = 12 \times 2 \times 10^{-6}$$

$$\phi = \tan^{-1}\left(\frac{X_L - X_C}{R}\right)$$

$$24 \times 10^{-6} \times \frac{2 \times \pi \times 5000 \times 180°}{\pi} \tan^{-1}\left(\frac{12 - X_C}{3}\right)$$

$$X_C = 14.182 \qquad \frac{1}{2\pi f C} = X_C$$

$$C = 2.15\,\mu F$$

CHOPPER

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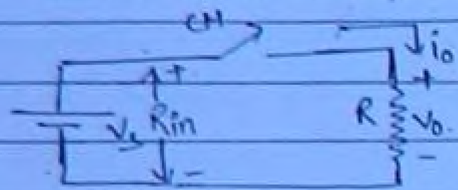
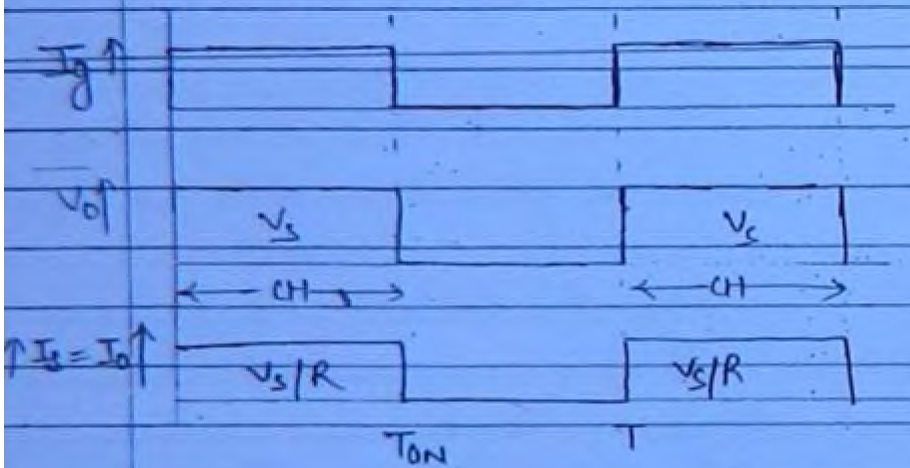
fixed DC $\longrightarrow$ variable DC

1. Step-down Chopper :- ($V_0 < V_s$) $\rightarrow$ without filter



Switch can be replaced by GTO or any power device for low power appl^n but for High-power, it should be SCR with forced commutation.



$$\alpha = \frac{T_{ON}}{T}$$

= Duty cycle.

$$V_0 = V_s\left(\frac{T_{ON}}{T}\right) = \alpha V_s$$

$$V_{rms} = \sqrt{\alpha}\, V_s$$

$$I_0 = \frac{V_0}{R} = \frac{\alpha V_s}{R} = I_{s0}$$

$$R_{in} = \frac{V_s}{I_s} = \frac{V_s}{\alpha \frac{V_s}{R}} = \frac{R}{\alpha}$$

$$\boxed{R_{in} = \frac{R}{\alpha}}$$

$$V_0 = \alpha V_s + \sum_{n=1}^{\infty} \frac{2V_s}{n\pi} \sin(n\omega t + \phi_n) \sin n\pi\alpha$$

Date / /

where $\phi_n = \tan^{-1}\left[\dfrac{\cos n\pi\alpha}{\sin n\pi\alpha}\right]$

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$$v_{0n} = \alpha V_s + \frac{2V_s}{n\pi}\sin n\pi\alpha \cdot \sin(n\omega t + \phi_n)$$

$v_{0n} = 0$ if $n\alpha = 1$

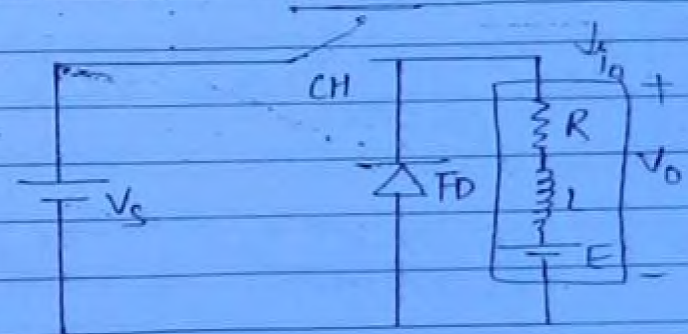
$$\boxed{\alpha = \frac{1}{n} \rightarrow \text{cond}^n \text{ to eliminate } n\text{th harmonic}}$$

$$FF = \frac{V_{or}}{V_0} = \frac{\sqrt{\alpha}\,V_s}{\alpha V_s} = \frac{1}{\sqrt{\alpha}}$$

$$\downarrow VRF = \sqrt{FF^2 - 1} = \sqrt{\frac{1}{\alpha\uparrow} - 1}$$

For High values of Duty cycle, harmonics are lesser.

RLE Load $\longrightarrow$ Cont. Conduction (waveform same for RL & RLE)

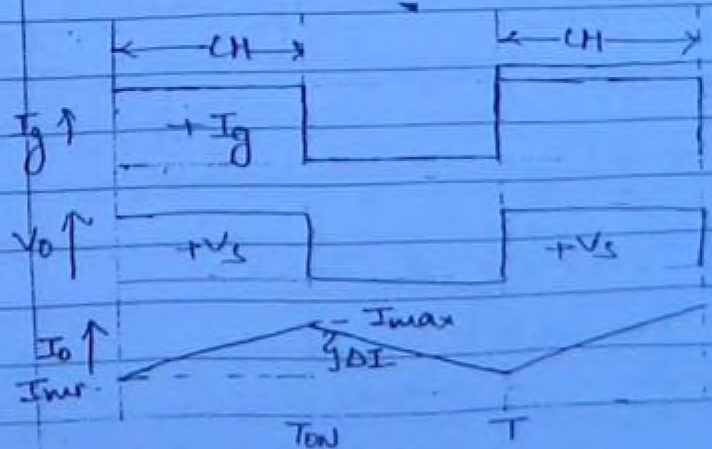


$V_0 = \alpha V_s$

$V_{or} = \sqrt{\alpha}\,V_s$

$$I_0 = \frac{\alpha V_s - E_b}{R}$$

Av. o/p will not depend on value of L but ripple will depend.



$$I_{max} = \frac{V_s}{R_a}\frac{(1-e^{-T_{on}/T_a})}{(1-e^{-T/T_a})} - \frac{E_b}{R_a} \quad \text{—— (1)}$$

where $T_a = \frac{L_a}{R_a} \longrightarrow$ w/c time constt

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$$I_{min} = \frac{V_s}{R_a}\left[\frac{e^{T_{on}/T_a}-1}{e^{T/T_a}-1}\right] - \frac{E_b}{R_a} \quad \text{—— (2)}$$

Ripple current $= \Delta I_o = I_{max} - I_{min}$

$$= \frac{V_s}{R_a}\left\{\frac{(e^{T_{on}/T_a}-1)}{(e^{T/T_a}-1)}\left(\frac{e^{T/T_a}}{e^{T_{on}/T_a}}\right) - \left(\frac{e^{T_{on}/T_a}-1}{e^{T/T_a}-1}\right)\right\}$$

$$= \frac{V_s}{R_a}\left\{\left(\frac{e^{T/T_a}-1}{e^{T_{on}/T_a}-1}\right)\left(\frac{e^{T_{on}/T_a}-1}{e^{T/T_a}-1}\right)\right\}$$

$$= \frac{V_s}{R_a}\left[\frac{(1-e^{-T_{on}/T_a})(1-e^{-T_{off}/T_a})}{(1-e^{-T/T_a})}\right]$$

as α varies, ΔI_o varies.

$$T_{on} = \alpha T, \quad T_{off} = (1-\alpha)T$$

$$\Delta I_o = \frac{V_s}{R_a}\left[\frac{(1-e^{-\alpha T/T_a})\cdot(1-e^{-(1-\alpha)T/T_a})}{(1-e^{T/T_a})}\right]$$

$$\frac{d}{d\alpha}\Delta I_o = 0 \Rightarrow \frac{V_s}{R_a(1-e^{-T/T_a})}\left[(1-e^{-\alpha T/T_a})e^{-(1-\alpha)T/T_a} \times \frac{T}{T_a} + (1-e^{-(1-\alpha)T/T_a})e^{-\alpha T/T_a} \times -\frac{T}{T_a}\right]$$

$$= (1-e^{-\alpha T/T_a})e^{-(1-\alpha)T/T_a} - (1-e^{-(1-\alpha)T/T_a})e^{-\alpha T/T_a} = 0$$

$$= e^{-(1-\alpha)T/T_a} - e^{-\alpha T/T_a}$$

$$(1-\alpha)\frac{T}{T_a} = \alpha\frac{T}{T_a}$$

$$\alpha = 1-\alpha \Rightarrow 2\alpha = 1$$

$$\boxed{\alpha = 0.5 \,\Big|_{\Delta I_o max}}$$

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$$\Delta I_0\Big|_{max} = \frac{V_s}{R_a}\frac{\left(1-e^{-0.5T/T_a}\right)^2}{\left(1-e^{T/T_a}\right)}$$

$$= \frac{V_s}{R_a}\tanh\left(\frac{T}{4T_a}\right)$$ (167)

$$\Delta I_0\Big|_{max} \simeq \frac{V_s}{R_a}\times\frac{T}{4T_a} = \frac{V_s}{R_a}\times\frac{1}{f\times 4\times\frac{L_a}{R_a}}$$

$$\boxed{\Delta I_0\Big|_{max} \simeq \frac{V_s}{4fL_a}} \longrightarrow \text{at } \alpha = 0.5$$

from this Equ^n $\downarrow$ Ripple current $\alpha \frac{1}{\uparrow f L_a \uparrow}$

for $L_a = \infty \Rightarrow \Delta I_0|_{max} = 0$

i.e. for Highly inductive load $\longrightarrow$ Const current.

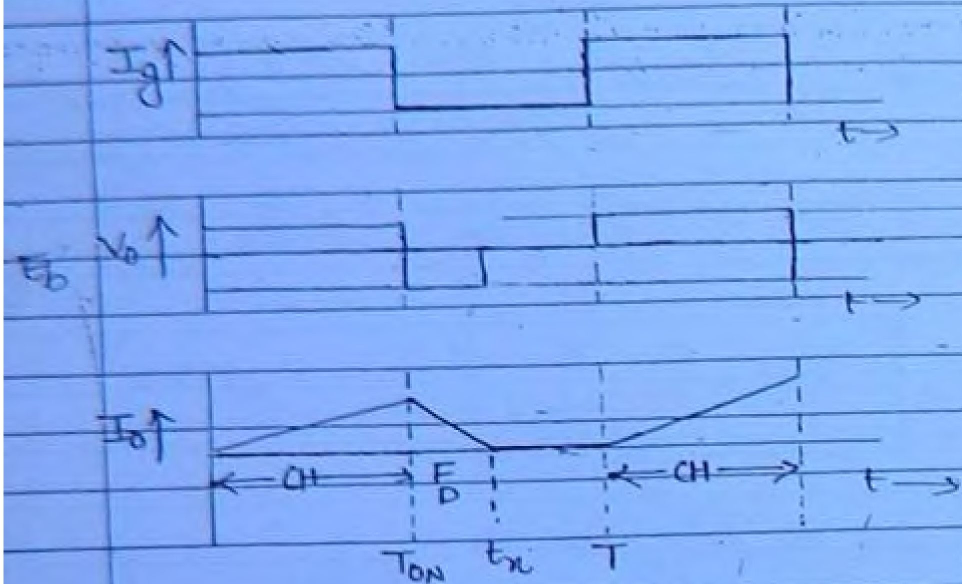
for High frequency $\longrightarrow$ Ripple will $\downarrow$ $\therefore$ chopper is being operated at very high frequency i.e. in range of KHz.

$\Rightarrow$ At very high switching frequency we can reduce the ripple in the curr without $\uparrow$ing the size of filter. SMPS operates on the same chopper principle.

Reasons for Discont. conduction

Rectifier	Chopper
$\uparrow\alpha \longrightarrow$ firing angle.	$\downarrow\alpha \rightarrow$ Duty cycle
$\downarrow L$	$\downarrow L$
$\downarrow I_0$	$\downarrow I_0$

RLE Load → Discontinuous conduction



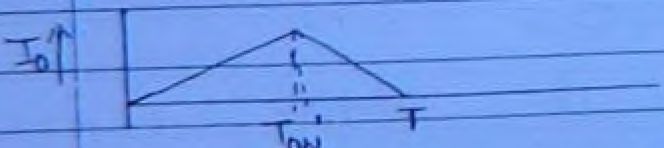
$T_x \rightarrow$ Extinction Time

$$V_0 = V_s\left(\frac{T_{ON}}{T}\right) + E_b\left(\frac{T - t_x}{T}\right)$$

$$V_0 = \alpha V_s + E_b\left(1 - \frac{t_x}{T}\right)$$

Duty cycle limit for continuous conduction :-

Let α' be the duty cycle at the boundary b/w continuous & discontinuous conduction.



at $t = T$, $I_{min} = 0$

$$I_{min} = \frac{V_s}{R}\left(\frac{e^{T_{ON}/T_a} - 1}{e^{T/T_a} - 1}\right) - \frac{E_b}{R_a} = 0$$

$$\frac{e^{T_{ON}/T_a} - 1}{e^{T/T_a} - 1} = \frac{E_b}{V_s}$$

at boundary, $e^{T_{ON}'/T_a} - 1 = \frac{E_b}{V_s}\left(e^{T/T_a} - 1\right)$

$$e^{T_{ON}'/T_a} = \frac{E_b}{V_s}\left(e^{T/T_a}-1\right)+1$$

$$T_{ON}' = T_a \ln\left[\frac{E_b}{V_s}\left(e^{T/T_a}-1\right)+1\right]$$

$$\alpha' = \frac{T_{ON}'}{T} = \frac{T_a}{T}\ln\left[\frac{E_b}{V_s}\left(e^{T/T_a}-1\right)+1\right]$$

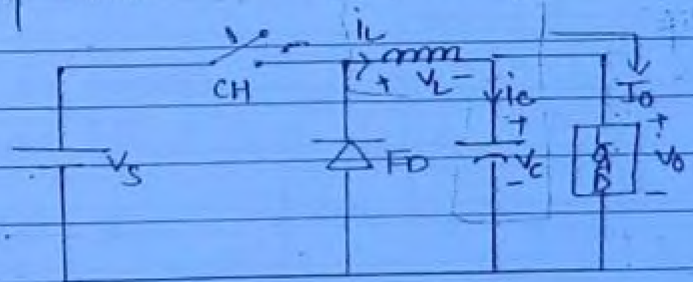
$\alpha < \alpha' \longrightarrow$ Discontinuous conduction.

$\alpha > \alpha' \longrightarrow$ Continuous conduction.

Step-down chopper ----→ with filters

Buck Regulator

In the step-down chopper, we could ↓ the ripple in current by ↑ L or f but Ripple in v/g was uneffected ∴ to reduce Ripple in v/g, step-down chopper with filter is being used.

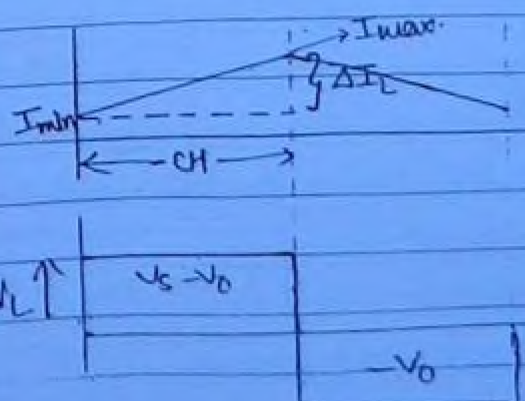


$I_o <$ (I_L)avg. / ΔI_L

Assume high value of capacitor across the load which maintains almost constt v/g across the load.

(I_L)avg $\longrightarrow$ Dc component $= I_o$

$\Delta I_L = \Delta I_c$



$0 < t < T_{ON}$

$CH \longrightarrow ON$

$-V_S + V_L + V_0 = 0$

$V_L = V_S - V_0$

$L\dfrac{dI_L}{dt} = V_S - V_0$

$dI_L = \dfrac{V_S - V_0}{L}\, dt$

$$\int_{I_{min}}^{I_{max}} dI_L = \frac{V_S - V_0}{L}\int_0^{T_{ON}} dt$$

$$(I_{max} - I_{min}) = \frac{(V_S - V_0)\,T_{ON}}{L}$$

$$\Delta I_L = \left(\frac{V_S - V_0}{L}\right) T_{ON}$$

$$= \frac{V_S(1-\alpha)\,T_{ON}}{L}$$

$$\boxed{\begin{array}{l} \Delta I_L = \dfrac{V_S(1-\alpha)\,\alpha}{fL} \\ \Delta I_L \propto \dfrac{1}{fL} \end{array}}$$

$$\frac{d\Delta I_L}{d\alpha} = \frac{V_S}{fL}(1 - 2\alpha) = 0$$

$\alpha = 0.5$ for $\Delta I_L\big|_{max}$.

$$\Delta I_{L\,max}\Big|_{\alpha=0.5} = \frac{V_S}{4fL}$$

$$\boxed{I_S = I_{CH}}$$

II $T_{ON} \leq t \leq T$

$CH \rightarrow off$, $FD \rightarrow ON$

120

$+V_L + V_0 = 0$

$\therefore V_L = -V_0$

$(V_L)_{avg} = 0$

Ar. of (+)ve spike = Ar. of (−)ve spike

$(V_S - V_0)\,T_{ON} = V_0\,T_{off}$

$V_0\,(T_{off} + T_{ON}) = V_S\,T_{ON}$

$V_0\,T = V_S\,T_{ON}$

$V_0 = V_S\dfrac{T_{ON}}{T}$

$$\boxed{V_0 = \alpha V_S}$$

Assuming No loss in chopper
I/p power = o/p power

$V_0 I_0 = V_S I_S$

$$\frac{V_0}{V_S} = \frac{I_S}{I_0} = \alpha$$

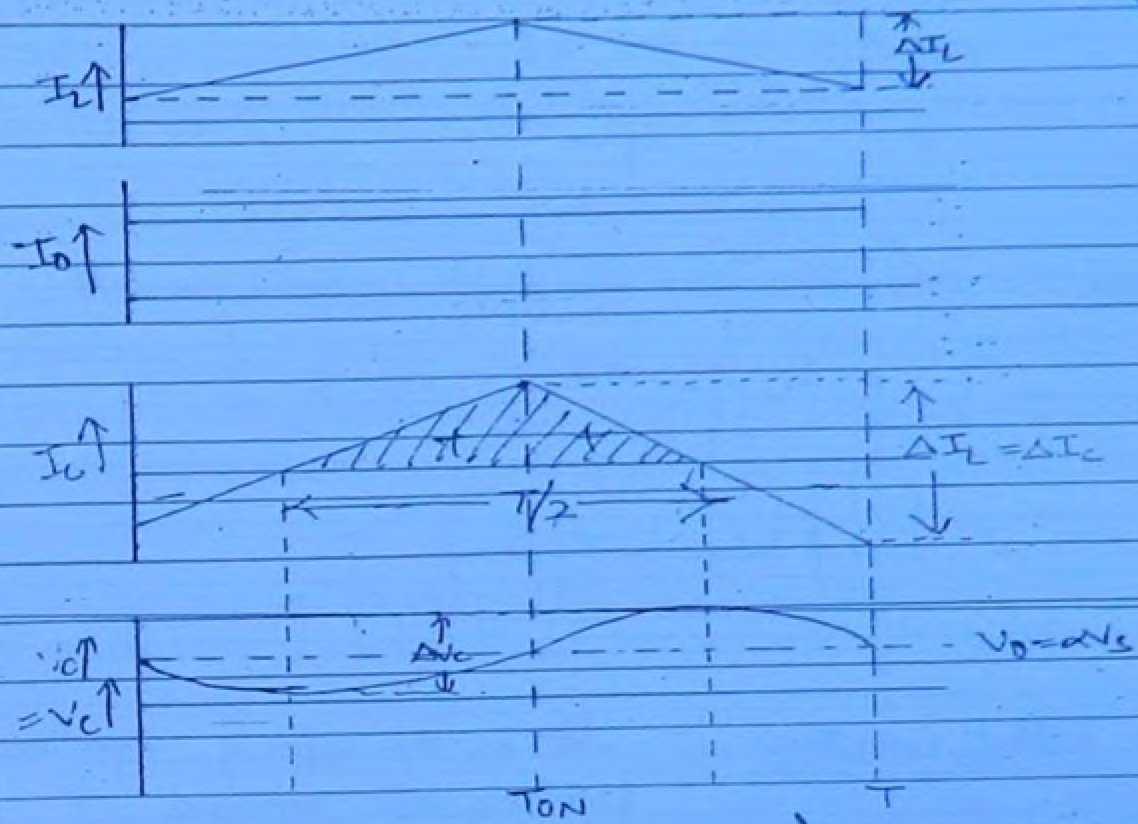
CHOPPER works as DC Transformer.

$$\boxed{\begin{array}{l} I_S = \alpha I_0 \\ I_0 = \dfrac{I_S}{\alpha} \end{array}}$$

$$\boxed{(I_{FD})_{avg} = (1-\alpha)\,I_0}$$

Ripple in capacitor voltage ($\Delta V_C = \Delta V_0$)

(171)



$$\Delta Q = C\Delta V_C \Rightarrow \Delta Q = \frac{\Delta I_L}{2} \times \frac{1}{2} \times \frac{T}{2} = \frac{\Delta I_L T}{8}$$

$$\boxed{\Delta V_C = \frac{\Delta I_L}{8fC}} \qquad \Delta V_C \propto \frac{1}{fC}$$

$$\Delta V_C = \frac{\alpha(1-\alpha)V_s}{8f^2LC}$$

$$\Delta V_C\Big|_{max} \text{ at } \alpha = 0.5 = \frac{V_s}{32f^2LC}$$

Critical Inductance :- It is the value of inductance at which the inductance current waveform is just discontinuous.

Symbol $\rightarrow L_C$

At the boundary b/w continuous & discontinuous condⁿ of I_L

$$I_0 = \frac{\Delta I_L}{2} = (I_L)_{avg}$$

(172)

$$I_0 = \frac{\alpha(1-\alpha)V_s}{2fL_c}$$

$$L_c = \frac{\frac{1}{2f I_0}}{\alpha(1-\alpha)V_s}$$

$$L_c = \frac{\frac{1}{2f \alpha V_s/R}}{\alpha(1-\alpha)V_s} =$$

$$\boxed{L_c = \frac{(1-\alpha)R}{2f}}$$

Critical Capacitance :- It is the value of capacitance at the boundary b/w continuous & discontinuous conduction for the capacitance o/p waveform.

At the boundary :- $V_0 = \frac{\Delta V_c}{2} = (V_c)_{avg}$

$$V_0 = \frac{\alpha(1-\alpha)V_s}{8f^2 L_c C_c} \times \frac{1}{2}$$

$$\alpha V_s = \frac{\alpha(1-\alpha)V_s}{8f^2 L_c C_c} \cdot \frac{1}{2}$$

$$\boxed{C_c = \frac{(1-\alpha)}{16f^2 L_c}}$$

Sol

$\alpha = 0.5$, $\Delta I_L = 1.6$, $I_0 = 5A$

$(I_s)_{max} = (I_L)_{max}$

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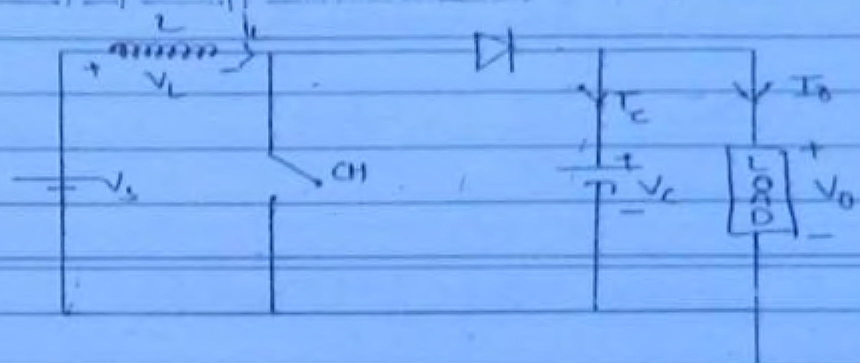
$$(I_L)_{max} = I_o + \frac{(\Delta I_C)}{2}$$

$$= \frac{1.6}{2} + 5 = 5.8 \text{ Amp}$$

3. Step-up Chopper $(V_0 > V_s)$ (with filter) ——→ Boost Regulator



(I) $0 \leq t \leq T_{ON}$ → $I_C = -I_0$

CH → ON, D → OFF

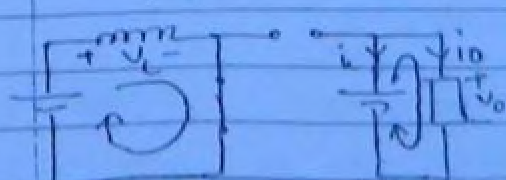
$V_L = V_s$

$$L\frac{dI_L}{dt} = V_s$$

$$\int_{I_{min}}^{I_{max}} dt = \frac{V_s}{L}\int_0^{T_{ON}} t$$

$$\Delta I_L = \frac{V_s}{L} T_{ON}$$

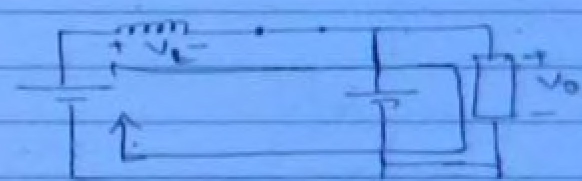
$$\boxed{\Delta I_L = \frac{\alpha V_s}{fL}}$$



$I_C = -I_0$

II $T_{ON} \leq t \leq T$

CH → OFF, D → ON



$V_S = V_L + V_0$

$V_L = V_S - V_0 = -(V_0 - V_S)$

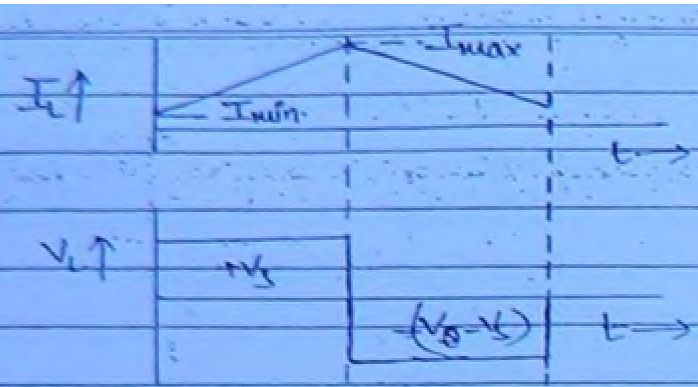
Since it is step-up chopper,

$V_0 > V_s \Rightarrow V_L \Rightarrow$ (−)ve

L is relasing energy

$I_L = I_C + I_0$

$I_C = I_L - I_0$



$(V_L)avg = 0$

$V_s T_{ON} - (V_0 - V_s) T_{off} = 0$

$V_s (T_{ON} + T_{off}) = T_{OFF} V_0$

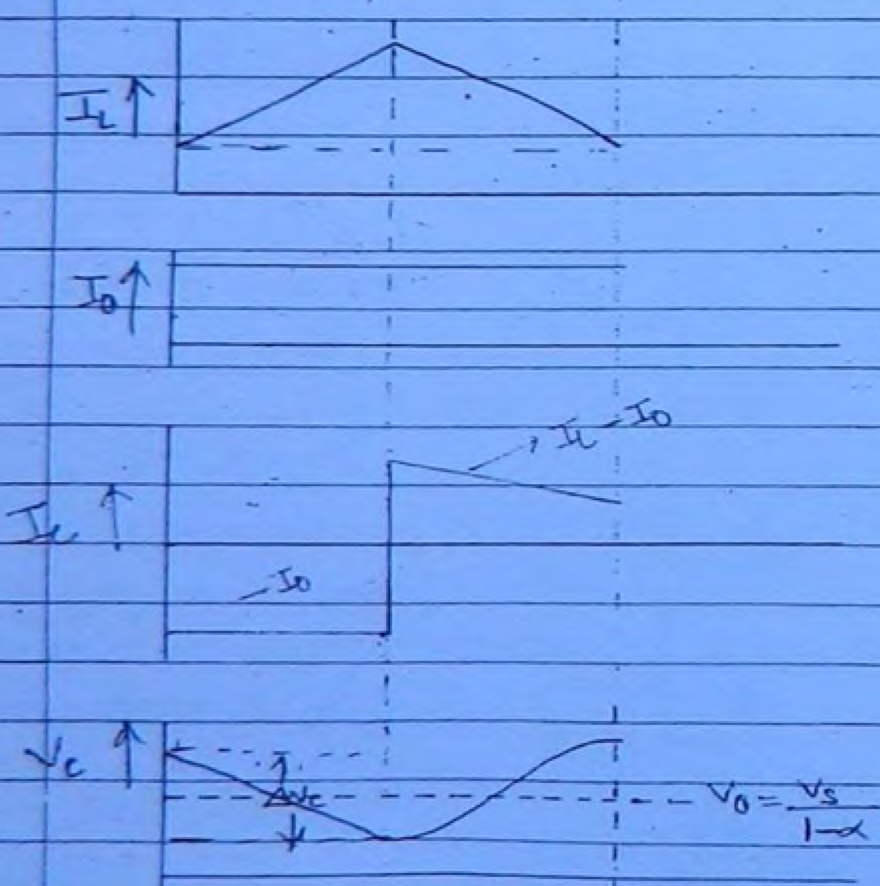
$V_s T = V_0 T_{off}$

$V_s T = (1-\alpha) V_0 T_{OFF}$

$$\boxed{V_0 = \frac{V_s}{1-\alpha}}$$

$$\frac{V_0}{V_s} = \frac{I_s}{I_0} = \frac{1}{1-\alpha}$$

Ripple in Capacitor voltage :- $(\Delta V_c = \Delta V_0)$.



$$\Delta V_c = \frac{\Delta Q}{C}$$

$$\Delta V_c = \frac{I_0 T_{ON}}{C}$$

$$\boxed{\Delta V_c = \frac{\alpha I_0}{fC}}$$

Critical Inductance (L_c) :-

At the boundary cond^{n}s of I_L

$$I_0 = \frac{\Delta I_L}{2} = (I_L)_{av.}$$

125

$$\frac{V_s}{R(1-\alpha)} = \frac{\alpha V_s}{2 f L_c}$$

$$\boxed{L_c = \frac{R(1-\alpha)\alpha}{2f}}$$

Critical Capacitance (C_c) :-

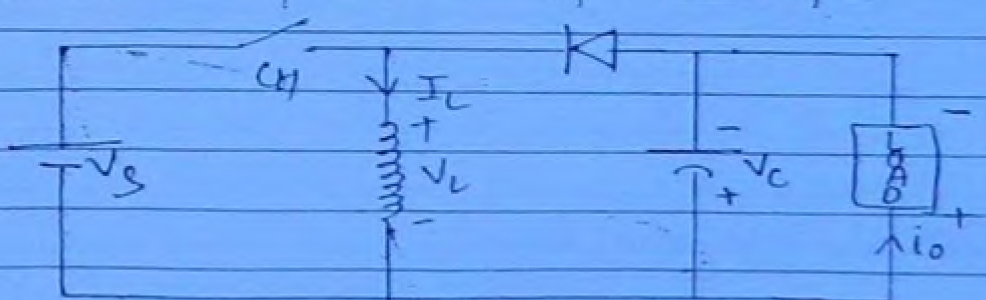
At the boundary cond^{n} of V_c waveform.

$$V_0 = \frac{\Delta V_c}{2} = (V_c)_{av.}$$

$$I_0 R = \frac{\alpha I_0}{2 f C_c}$$

$$\boxed{C_c = \frac{\alpha}{2 f R}}$$

Buck - Boost Regulator (Step up / Step down Chopper)

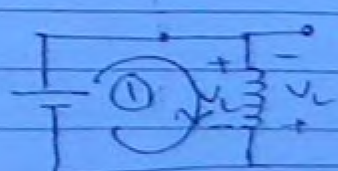


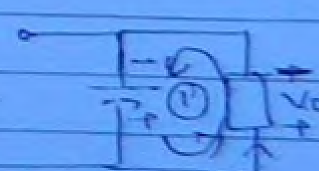
$\alpha > 0.5 \rightarrow$ step up

$\alpha < 0.5 \rightarrow$ step down

Mode - I $\quad 0 \leq t < T_{ON}$

Same as step-up chopper





Mode-2 CH → OFF, D → ON $T_{ON} \le t \le T$

$V_S = V_L$

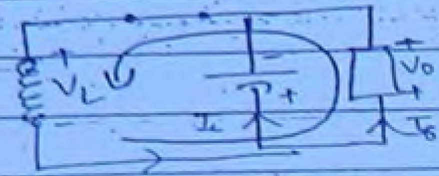
$L\frac{di}{dt} = V_S$

$$\int_{I_{min}}^{I_{max}} di = \frac{V_S}{L}\int_0^{T_{ON}} dt$$

$$\Delta I_L = \frac{V_S}{L} T_{ON}$$

$$\boxed{\Delta I_L = \frac{\alpha V_S}{fL}}$$

$$\boxed{I_C = -I_0}$$

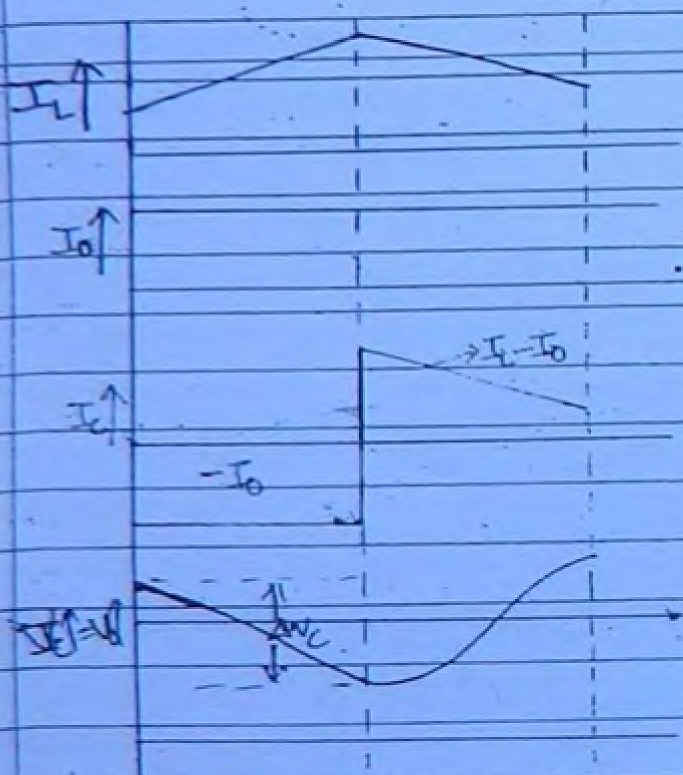


$+V_L + V_0 = 0$

$\therefore V_L = -V_0$

$I_L = I_C + I_0$

$$\boxed{I_C = I_L - I_0}$$



$V_0 = \frac{\alpha V_S}{1-\alpha}$

$(V_L)_{avg} = 0$

$V_S T_{ON} - V_0 T_{off} = 0$

$V_S T_{ON} = V_0 T_{off}$

$V_S \alpha T = V_0 (1-\alpha) T$

$$\boxed{V_0 = \frac{V_S \alpha}{1-\alpha}}$$

$$\frac{V_0}{V_S} = \frac{I_S}{I_0} = \frac{\alpha}{1-\alpha}$$

$$\Delta V_C = \frac{\Delta Q}{C}$$

$$= \frac{I_0 T_{ON}}{C} = \frac{\alpha I_0}{fC}$$

Critical Inductance (L_c):-

$$I_0 = \frac{\Delta I_L}{2}$$

$$\frac{\alpha V_S}{(1-\alpha)R} = \frac{\alpha V_S}{2fL_c} \Rightarrow \boxed{L_c = \frac{(1-\alpha)R}{2f}}$$

Critical Capacitance :-

$$V_0 = \frac{\Delta V_c}{2}$$

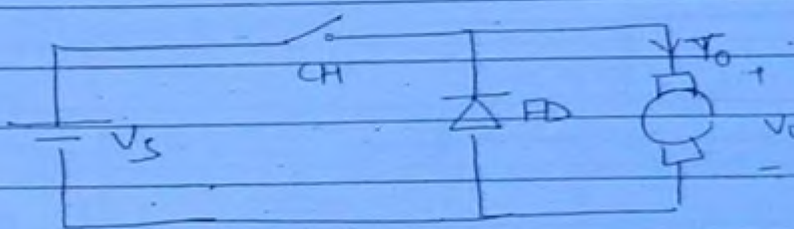
(177)

$$\frac{\alpha V_s}{1-\alpha} = \frac{\alpha I_0}{2 f C_c} = I_0 R$$

$$\boxed{C_c = \frac{\alpha}{2 f R}}$$

Classification of chopper based on quadrant operation

I First quadrant chopper (Type A) — (Stepdown choppers)



(I) $CH_1 \rightarrow ON$, $V_0 = V_s$

$\therefore V_0$ (+)ve, E_b (+)ve get

II $CH \rightarrow OFF$
$D \rightarrow ON$ [FWP] I_0 (+)ve

$\therefore E_a \rightarrow$ (+)ve

E_a +ve

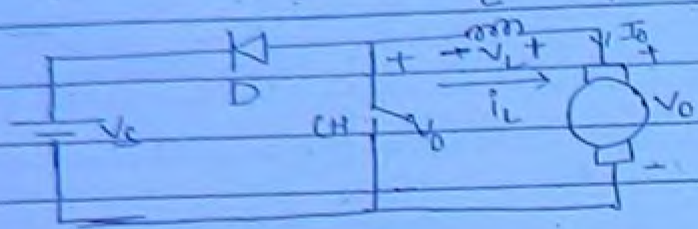
V_0 (+ve) $\boxed{\alpha V_s = E_b + I_0 R_a}$

P(+)ve, ϕ(+)ve

I_0 (+)ve

II Second quadrant operation (Type B Chopper)

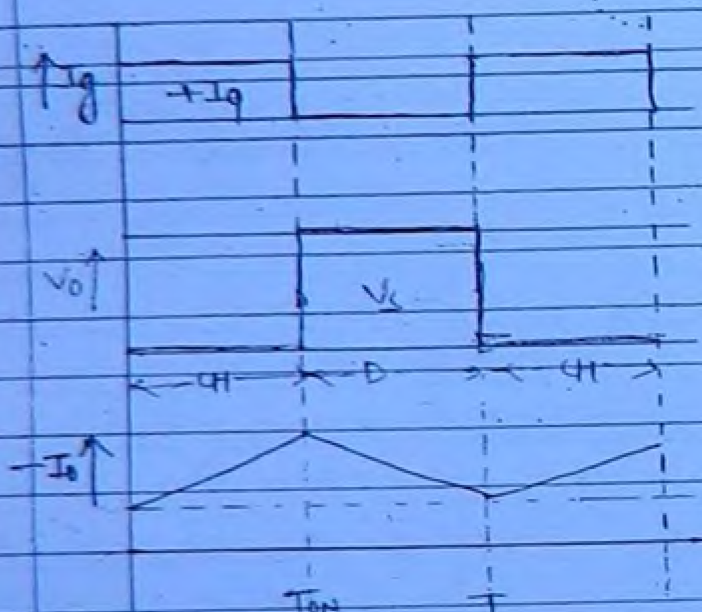
Regenerating Braking of DC Motor :-



Assume that m/c is operating at rated speed b/t T = 0 seconds

Mech Energy $= \frac{1}{2} J\omega^2 \longrightarrow$ Brake Energy

$$I_0 = \frac{V_0 - E_b}{R_a}$$

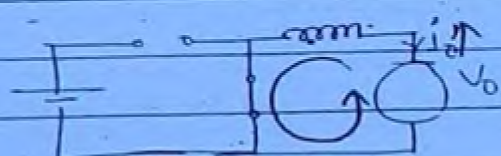


Mode - I

$CH_2 \longrightarrow ON$, $D \longrightarrow off$

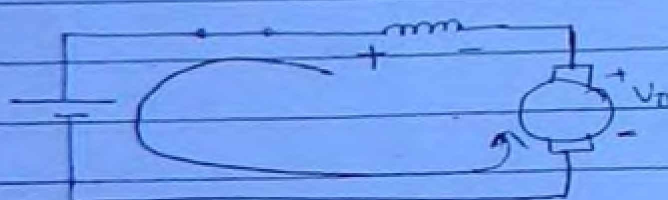
$0 < t < T_{ON}$

$\therefore I_0$ (-)ve, $\therefore I_a \longrightarrow \phi I_g$ (-)ve



$\frac{1}{2} J\omega^2 \longrightarrow \frac{1}{2} Li^2$

Mode - II , $T_{ON} \leq t \leq T$, $CH \longrightarrow OFF$, $D \longrightarrow ON$

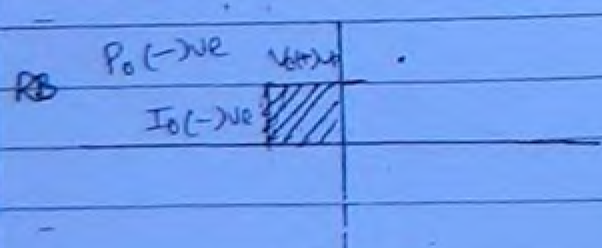


$V_0 = V_s$

$\frac{1}{2} Li^2 \longrightarrow$ Source

Brake Energy to source

Regenerative Braking



$V_0 = V_s (T_{off}/T) = (1-\alpha) V_s$

Regenerated power $= V_0 I_0$

$= V_s(1-\alpha) \cdot I_0$

Date / /

III → Two quadrant operation (Type C Chopper)

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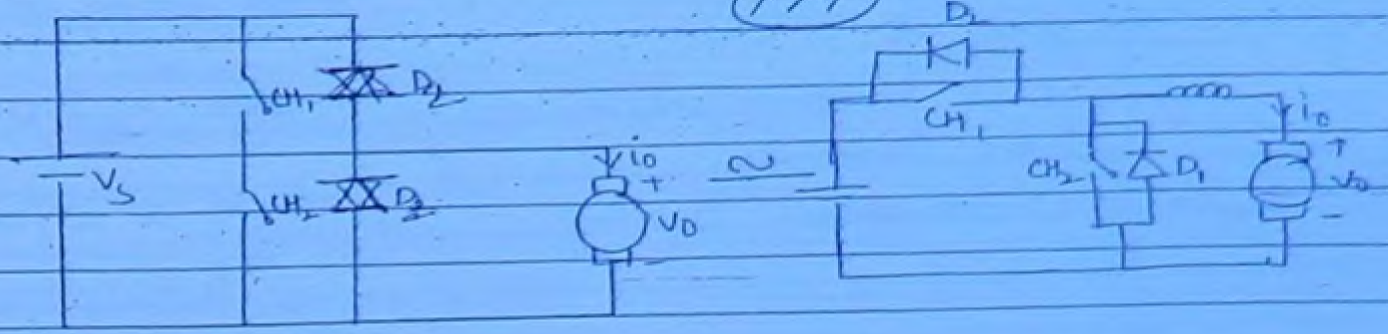


V_0

I $CH_2 \rightarrow ON$

$I_0 \rightarrow (-)ve$, T_a (-)ve

$\frac{1}{2}J\omega^2 \rightarrow \frac{1}{2}Li^2$

II $CH_2 \rightarrow OFF$

$D_2 \rightarrow ON$

$\frac{1}{2}Li^2 \rightarrow$ source

RB

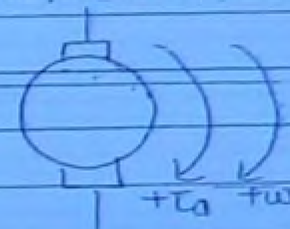
I → $CH_1 \rightarrow ON$, $D_2 \rightarrow off$

I_0 (+)ve, V_0 (+)ve, E_b (+)ve, T_a (+)ve, ω (+)ve

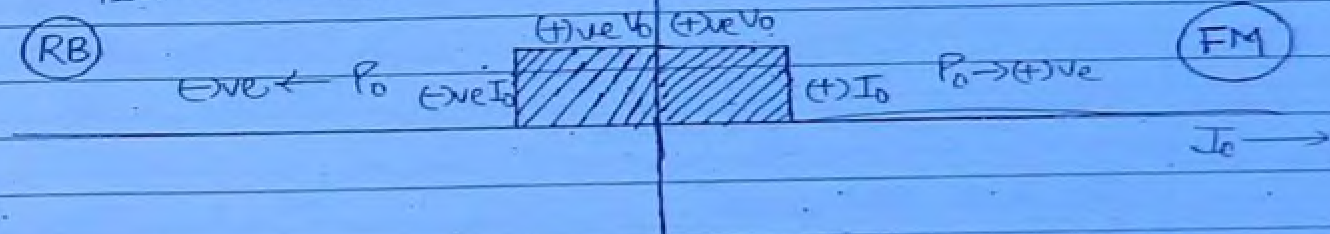
II $CH_1 \rightarrow OFF$

$D_1 \rightarrow ON$ (FWD)

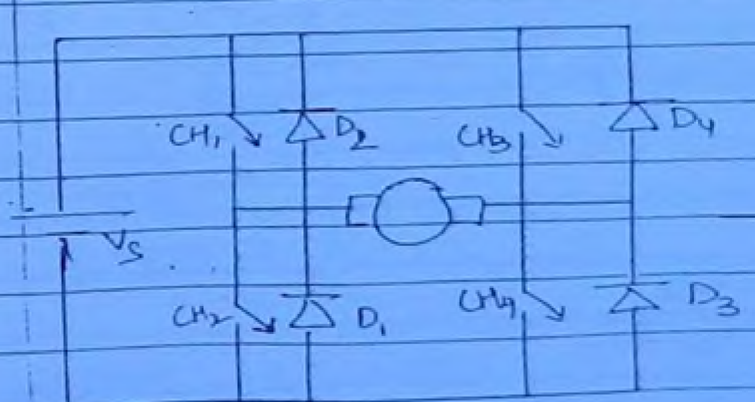
$\alpha V_s = E_b + I_0 R_a$

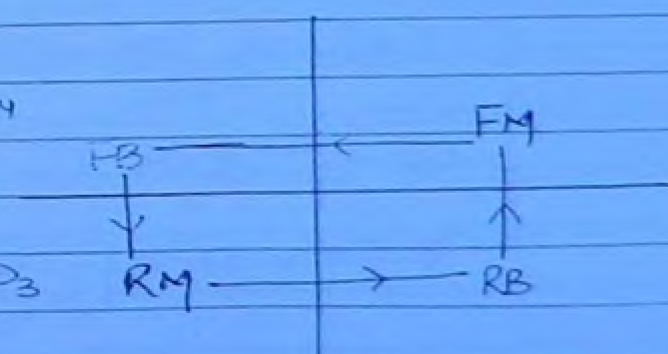


FM



IV → Four quadrant Chopper (Type-D)





Converter 17

3-ϕ semiconverter

$I_a \neq 0 \Rightarrow I_0 \neq 0$

$I_0(\omega t) = 0$

at $\omega t = \beta$, $I_0(\omega t) = 0$

i.e. it has discont conduction.

During discontinuity,

Load voltage = Back Emf

180

18 $R_{me} = 100\Omega$, FSD → 1mA

PMMC → measure Av. value ∴

It will measure DC v/g only.

$$V_{d0} = \frac{2V_m}{\pi} = I_f(100 + R_s)$$

$$\frac{2 \times 100\sqrt{2}}{\pi} = 10^{-3}(100 + R_s)$$

$$R_s = 89.9K\Omega$$

19. $\alpha > 90° \rightarrow$ Inversion mode

$P_{AC} \leftarrow P_{DC}$

$V_0 = -E + I_0 R_a$ — (1)

$\Delta V_{d0} = 4fL_c I_0$

$V_0 = V_{d0}\cos\alpha - 4fL_c I_0$

$$V_{d0} = \frac{2V_m}{\pi}$$

$$= \frac{2 \times 120\sqrt{2}}{\pi}$$

$$V_0 = \frac{240\sqrt{2}}{\pi}\cos 110 - \frac{4 \times 50 \times 10^{-3} \times I_0}{200}$$

$$-80 + I_0 \times 1 = \frac{240\sqrt{2}}{\pi}\cos 110°$$

$$I_0(1.2) = \frac{240\sqrt{2}}{\pi}\cos 110° + 80$$

$$I_0 = 85.87A$$

$$\Delta V_{d0} = \frac{V_{d0}}{2}\left[\cos\alpha - \cos(\alpha + u)\right] = 4fL_s I_0$$

$$u = 8.35°$$

23 Pulse No ↑ THD ↓

L filter, smoothness of waveform ↑

THD ↓

C filter, VRF ↓

but THD ↑

1 $V_0 = 50\%\ (V_0)_{max}$

$(V_0)_{max}$ when $\alpha = 0°$

$$(V_0)_{max} = V_{d0} = \frac{3V_{mL}}{2\pi}$$

$$V_0 = \frac{1}{2}\left[\frac{3V_{mL}}{2\pi}\right](1+\cos\alpha)$$

$$V_0 = V_{d0}(1+\cos\alpha)$$

$$\frac{1}{2}\left[\frac{3V_{mph}}{2\pi}\right] = \frac{3V_{mph}}{2\pi}\left[1+\cos(\alpha+30°)\right]$$

$$V_{ml} = \sqrt{3}V_{mph}$$

$$\frac{\sqrt{3}}{2} - 1 = \cos(\alpha + 30°)$$

$$\alpha = 67.7°$$

$$I_0 = \frac{V_0}{R} = 7.77\,Amp.$$

$$I_{0r} = \frac{V_{0r}}{R} = 10.48\,Amp.$$

$$\text{Rectification Efficiency} = \frac{V_0 I_0}{V_{0r} \cdot I_{0r}}$$

$$= 55.05\%$$

2. $\alpha_{max} = 180° - (\omega t_q + \mu_0)$

$\mu_0 \rightarrow$ overlap ∠ at $\alpha = 0°$

$V_0 = E_b + I_0 R_a$

$V_0 = V_{d0}\cos\alpha - 3 f L_s I_0$

$E_b + I_0 R_a = V_{d0}\cos\alpha - 3 f L_s I_0$

$$I_0 = \frac{V_{d0}\cos\alpha - E_b}{R_a + 3 f L_s}$$

$$V_{d0} = \frac{3V_{mL}}{2\pi} = 202.51V$$

$I_0 = 14.77A$

$\frac{V_{d0}}{2}(\cos\alpha - \cos(\alpha+\mu)) = 3 f L_s I_0$

$\mu|_{\alpha=0°} = 17°$

$\omega t_q = 2\pi f \cdot 250\times10^{-6} \cdot \frac{180}{\pi}$

$= 4.5°$

$\alpha_{max} = 180° - 4.5 - 17°$

$= 158.5°$

3. at $\alpha = 0°$ = $\mu = 30°$

$$\frac{I_0}{I_{0p}} = \cos\alpha - \cos(\alpha+\mu)$$

$= 0.134$

at $\alpha = 30°$

$0.134 = \cos 30° - \cos(30° + \mu)$

$\mu = 12.94°$

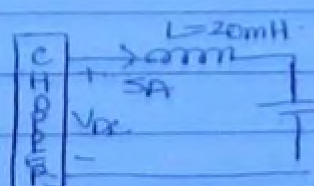
at $\alpha = 60°$

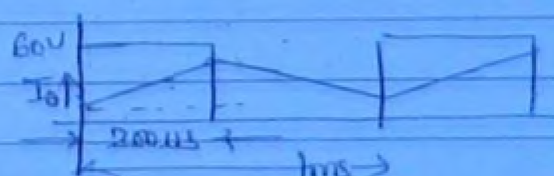
$0.134 = \cos 60° - \cos(60° + \mu)$

$\mu = 8.53°$

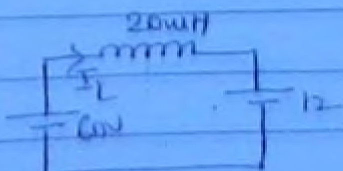
Choppers :-

1.





CH → ON, $V_{DC} = 60V$



$60 - 12 = L\frac{di}{dt}$

$$\Delta I_L = \frac{48}{20\times10^{-3}} \times T_{ON}$$

$$\Delta I_L = \frac{48}{20\times10^{-3}} \times 200\times10^{-6}$$

$= 480\times10^{-3} = 0.48A$

$V_s = 100V$

$\alpha = 0.8$

$(I_{FD}) = I_0 \frac{(T_{off})}{T}$

$= I_0(1-\alpha)$

$I_0 = \frac{\alpha V_s}{R}$

$= \frac{0.8 \times 100}{10} = 8$

$(I_{FD})_{av} = 8(1-0.8)$

$= 1.6A$

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3 CLASS-D Commutation → v/g commutation

$(T_{ON})_{min}$ of chopper $= \pi\sqrt{LC}$

$= 140\mu s$

4 $V_0 = V_s\left[\frac{T_{ON} + 2t_{CM}}{T}\right]$

$(V_0)_{min} = 250\left[\frac{140\times10^{-6} + 2t_{cm}}{}\right] \times 10^3$

$t_{CM} = \frac{CV_s}{I_0} = \frac{10^{-6} \times 250}{10}$

$= 25\times10^{-6}$

$(V_0)_{min} = 47.5V$

5 $(I_{TM})_{peak} = I_0 + V_s\sqrt{\frac{C}{L}} = 10 + 200\sqrt{\frac{10^{-7}}{10^{-3}}} = 10 + 200\times10^{-2} = 12A$

$(I_{TA})_{peak} = I_0 = 10A$

Step-down chopper → $\alpha = 0.5$

$I_0 = \frac{\alpha V_s}{R} = \frac{0.5\times60}{3} = 10A$

Av. value will not depend on Inductance.

$T_{ON}' = T_A \ln\left[1 + \frac{E_D}{V_s}(e^{T/T_a} - 1)\right]$

$T_a = \frac{L}{R} = \frac{10^{-3}}{0.25} = 4\times10^{-3}$

$\frac{T}{T_a} = \frac{2.5\times10^{-3}}{4\times10^{-3}} = 0.625$

$T_{ON}' = 332.5\mu\,sec$

Date / /

9. $\alpha = \frac{1}{5} \Rightarrow VRF = \sqrt{\frac{1}{\alpha} - 1} = \sqrt{5-1} = 2$

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10. $T_{min} = \pi\sqrt{LC}$

$t_{max} = \pi\sqrt{LC} + \sqrt{LC}\sin^{-1}\left(\frac{I_o}{I_p}\right)$

$I_p = V_s\sqrt{\frac{C}{L}} = 230 \times \sqrt{\frac{10\times10^{-6}}{25.28\times10^{-6}}}$

$= 144.66$

$t_{max} \simeq 75\mu s.$

$t_{min} = 50\mu s$

11. $L = 64\mu H.$

$(I_M)_{peak} = I_0 + V_s\sqrt{\frac{C}{L}} = 70.7 + 200\sqrt{\frac{64\times10^{-6}}{16\times10^{-6}}}$

$= 170.7A.$

1 c) $t_{cM} = \frac{CV_s}{I_0} = \frac{50\times10^{-6}\times220}{80} = 137.5\mu sec.$

$t_{cA} = \frac{\pi}{2}\sqrt{LC} = \frac{\pi}{2}\sqrt{20\times50}\times10^{-6} = 49.67\mu sec.$

a) $T_{ON}|_{effective} = T_{ON} + 2t_{cM}$

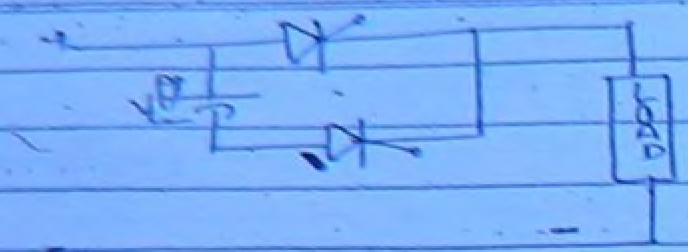
$= 800 + 2\times137.5 = 1075\mu s$

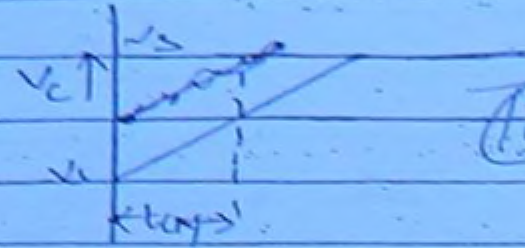
$(I_{TM})_p = I_0 + V_s\sqrt{\frac{C}{L}} = 427.85$ Amp.

$(I_{TA})_p = I_0 = 8$ Amp.

Total Commutation interval $= 2t_{cM}$

$= 2\times137.5 = 275\mu s.$





$$V_c = \frac{V_s}{t_{cm}} \cdot t - V_s \Rightarrow V_c = \frac{220}{137.5} \cdot 150 - 220$$

$$= 20V$$

13-12-12

AC VOLTAGE CONTROLLERS

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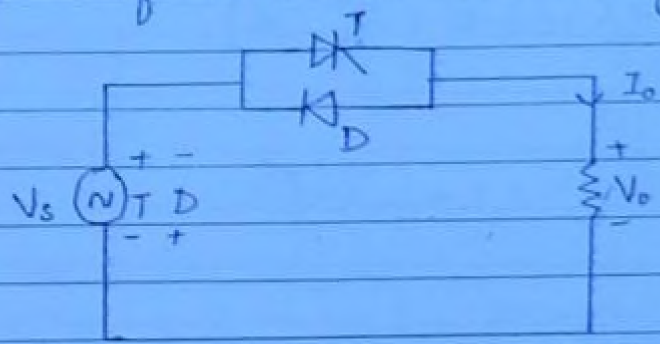
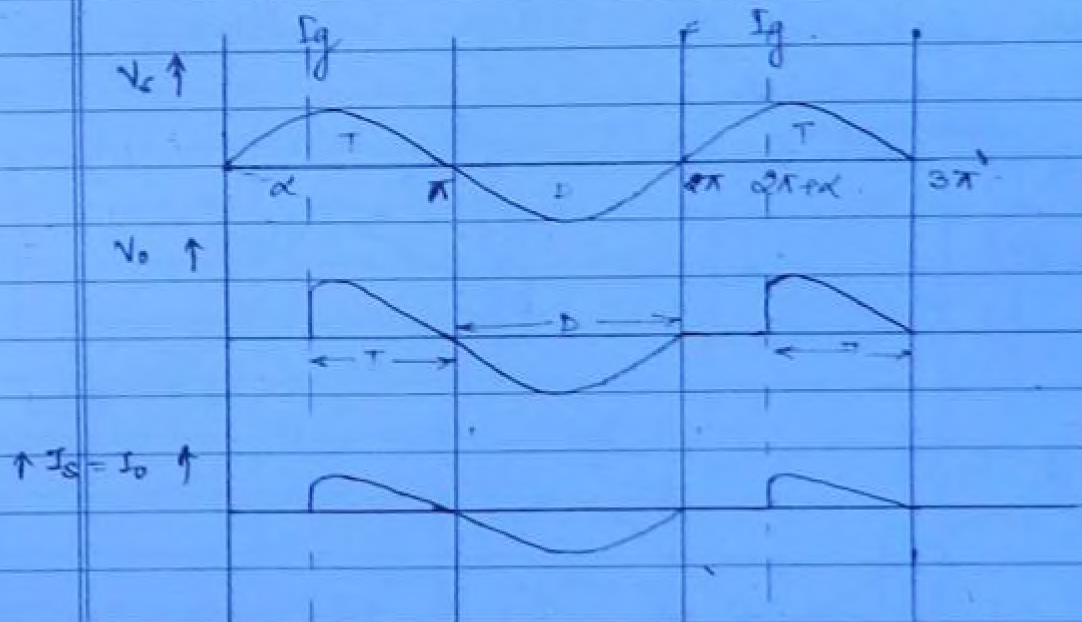
fixed AC $\longrightarrow$ variable AC $(\vec{V_0}, f_0)$

I Phase Control Technique -

a) 1-ϕ Half Controlled AC Voltage Controllers -





$$V_0 = \frac{1}{2\pi}\int_{\alpha}^{2\pi} V_m \sin\omega t \, d(\omega t) = \frac{V_m}{2\pi}(\cos\alpha - 1)$$

$$(I_s)_{avg} = I_0 = \frac{V_m}{2\pi R}(\cos\alpha - 1)$$

DC component

Drawback -
Source current contains DC component & saturates the supply transformer core.

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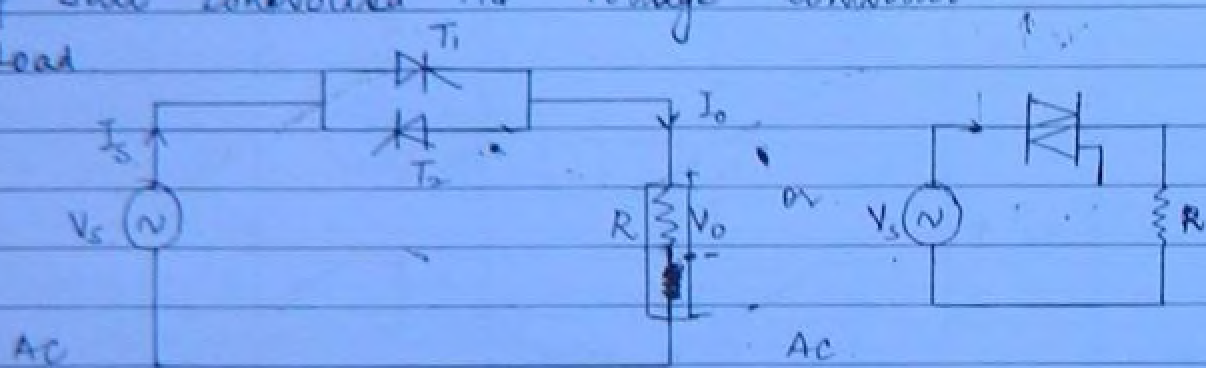
$$V_{or} = \left\{ \frac{1}{2\pi} \int_{\alpha}^{2\pi} V_m^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

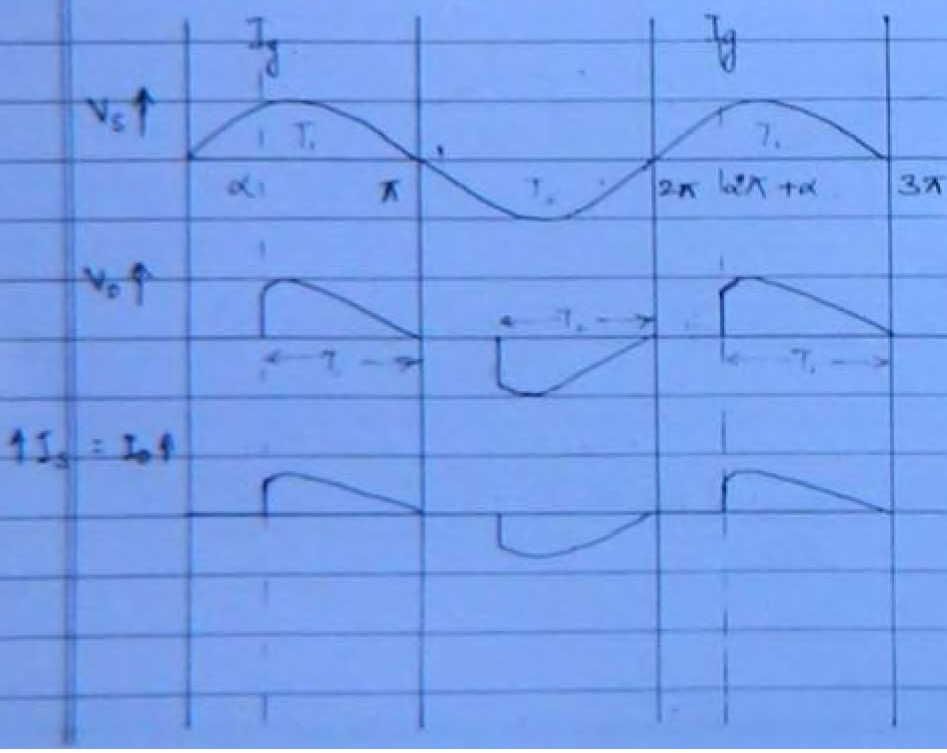
$$V_{or} = \frac{V_m}{2\sqrt{\pi}} \left[(2\pi - \alpha) + \frac{1}{2} \sin 2\alpha \right]^{1/2}$$

$$PF = \frac{V_{or}}{V_{sr}} = \frac{1}{\sqrt{2\pi}} \left[(2\pi - \alpha) + \frac{1}{2} \sin 2\alpha \right]^{1/2}$$

(b) 1-ϕ Full Controlled AC Voltage Controller -

1. R Load



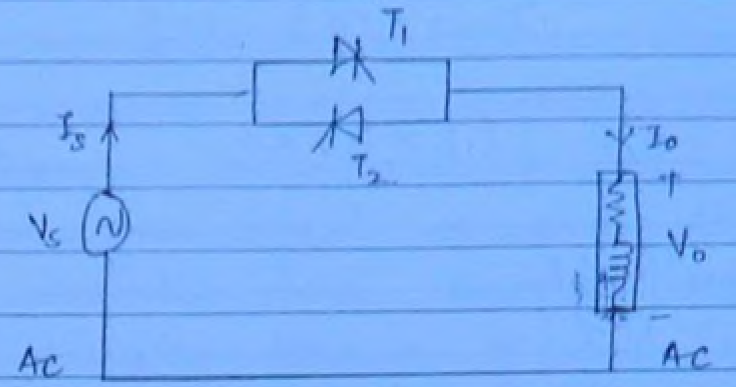


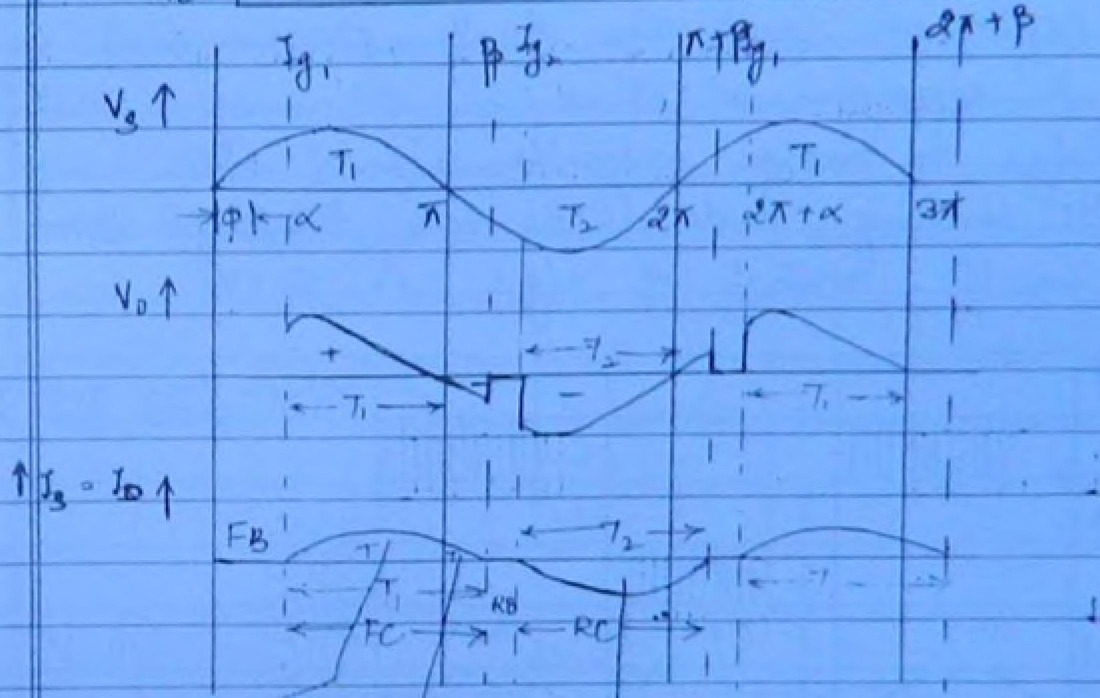
2. RL load.

After reaching steady state I_0 lags V_0 by $\phi = \tan^{-1}\frac{\omega L}{R}$

(I) $\alpha > \phi$, V_0 is controlled

(II) $\alpha \le \phi$, V_0 is uncontrolled.





P(+) power flows from source to load

(I) $\alpha > \phi$ V_0 is controlled → (By α ↗ O/p v/g can be controlled)

P(+) power again flows from source to load

P(−) load → source

(Inductor changes polarity)

$$V_{or} = \left\{ \frac{1}{\pi} \int_{\alpha}^{\beta} V_m^2 \sin^2 \omega t \, d(\omega t) \right\}^{1/2}$$

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$$V_{or} = \frac{V_m}{\sqrt{2\pi}} \left[(\beta - \alpha) + \frac{1}{2} (\sin 2\alpha - \sin 2\beta) \right]^{1/2}$$

II $\phi > \alpha$

$L \uparrow \quad T \uparrow \quad \beta \uparrow \quad (\beta > \pi + \alpha)$

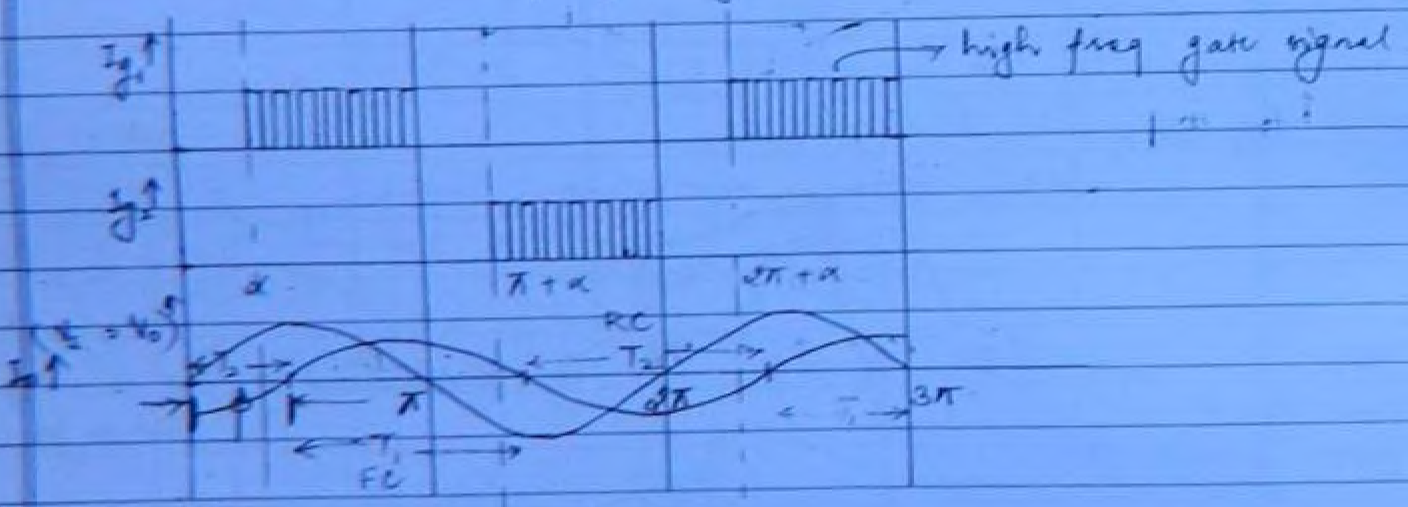
$\uparrow \phi = \tan^{-1} \left(\frac{\omega L}{R} \right)$

⇒ continuous conduction of V_0 occurs as β approaches $\pi + \alpha$

Only 2 modes available FC & RC

No blocking mode (SCRs are short ckted)

∴ No control of vlgs ∴ $V_0 = V_s$



When T_1 stops conducting, but there is no gate signal for T_2 (I_{g2}), T_2 fails to turn-on & it behaves as a rectifier

To avoid this continuous gate signal is given but that ↑ power loss & also may saturate pulse Trans

We get max^m output v/g when its uncontrolled

$$(V_0)_{rms} = (V_s)_{rms} = \frac{V_m}{\sqrt{2}}$$

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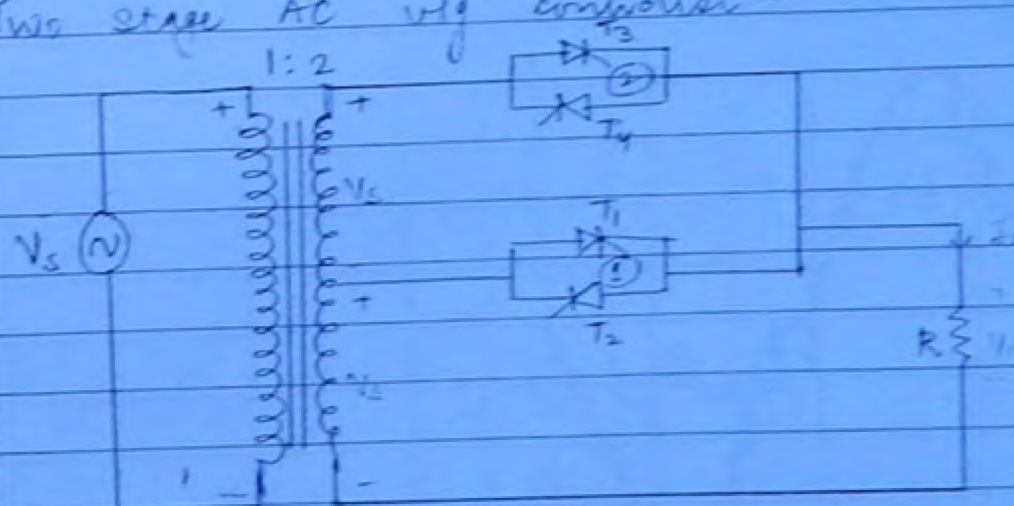
$$(I_{or})_{max} = I_{or} = \frac{(V_{or})_{max}}{|Z|}$$

$$|Z| = \sqrt{R^2 + (\omega L)^2}$$

NOTE.

If pulse gate signal is given to AC v/g controllers with inductive load, then it may behave as half wave rectifier. because of one of the SCRs fails to turn on due to the absence of gate signal at that instant. ~~we~~ To avoid this problem we can give either continuous gate signal or high frequency gate signal

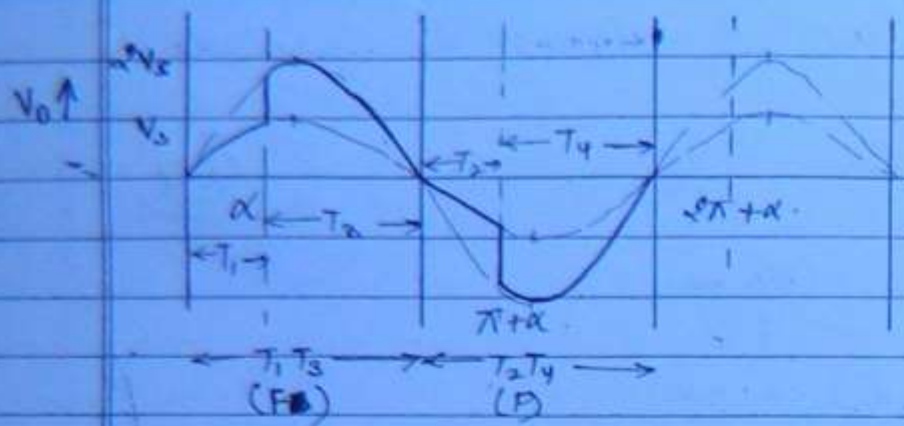
c) Two stage AC v/g controller —



+ $T_1 T_3$ (FB)

− $T_2 T_4$ (FB)

fwd biase



Since waveform is approaching sine wave, the smoothness ↑ with no. of stages
Thus harmonic distortion ↓.

$$V_{or} = \left\{ \frac{1}{\pi} \left[\int_0^{\alpha} V_m^2 \sin^2 \omega t \, d(\omega t) + \int_{\alpha}^{\pi} 4V_m^2 \sin^2 \omega t \, d(\omega t) \right] \right\}^{1/2}$$

$$= \frac{V_m}{\sqrt{2\pi}} \left\{ \left[\alpha - \frac{1}{2} \sin 2\alpha \right] + 4 \left[(\pi - \alpha) + \frac{1}{2} \sin 2\alpha \right] \right\}.$$

d) Multistage Regulator –

II Integral Cycle control (ON/OFF)

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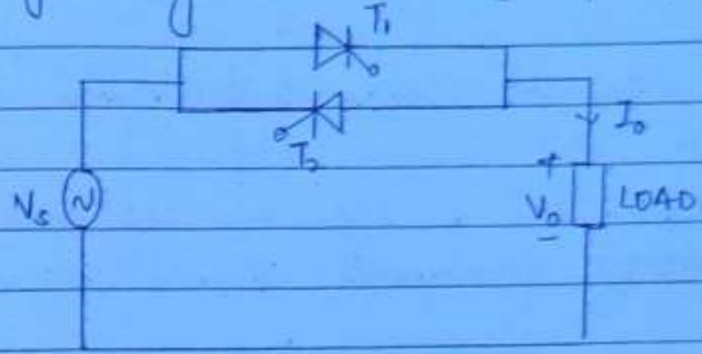
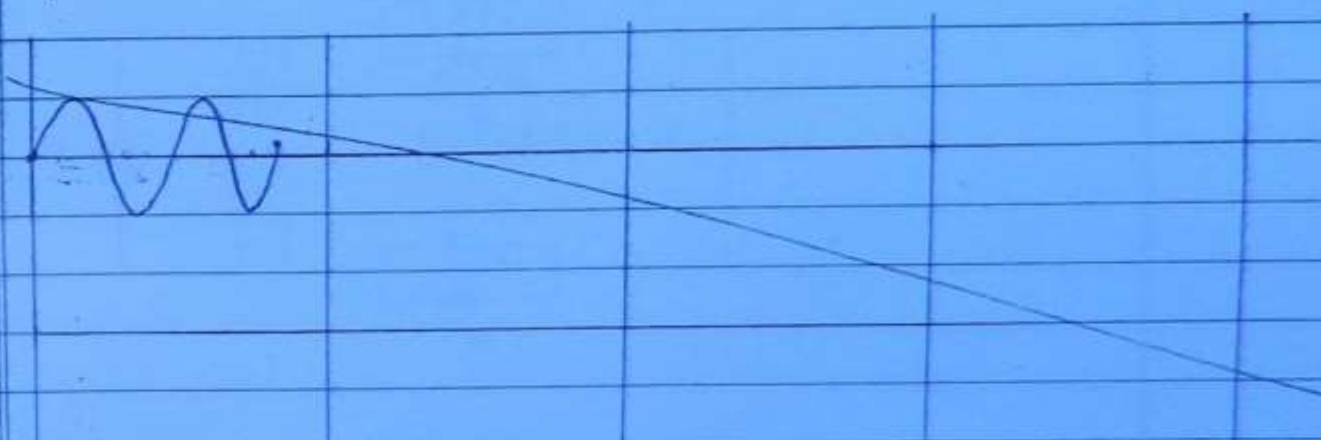


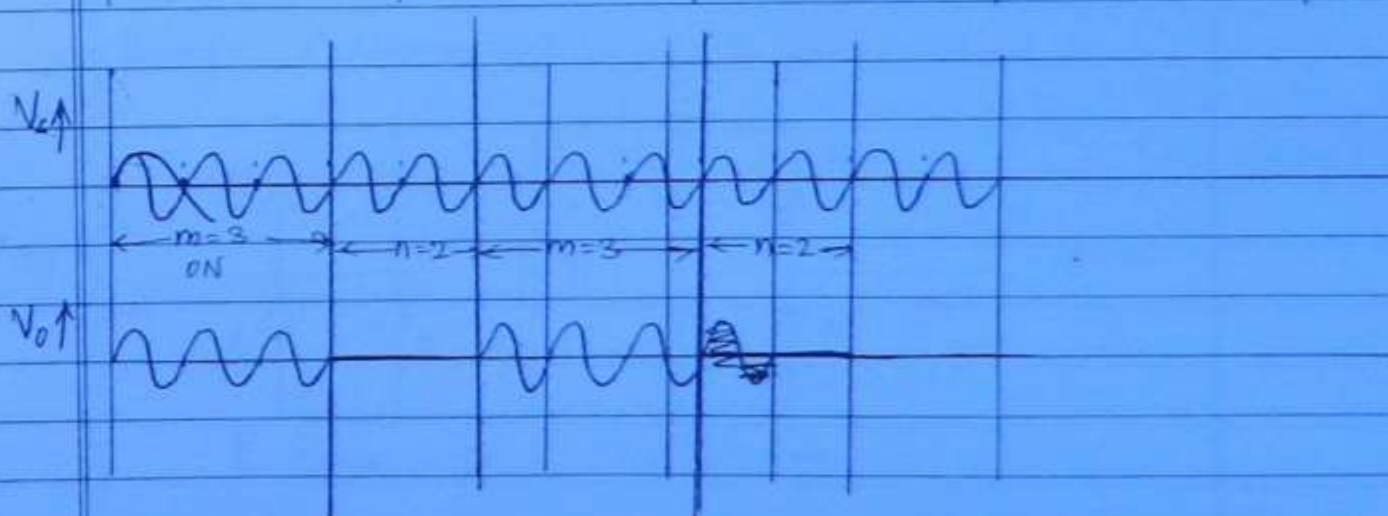
m cycles (ON) [m = 3]

n cycles (OFF) [n = 2]

I_{g1} $(0, 2\pi, 4\pi, \cancel{6\pi, 8\pi}, 10\pi, 12\pi, 14\pi, \cancel{16\pi, 18\pi} \ldots\ldots)$

I_{g2} $(\pi, 3\pi, 5\pi, \cancel{7\pi, 9\pi}, 11\pi, 13\pi, 15\pi, \cancel{17\pi, 19\pi} \ldots)$





$$V_{or} = V_{sr}\left(\frac{m}{m+n}\right)^{1/2}$$

$$V_{or} = \sqrt{K}\, V_{sr} \quad \text{where} \quad K = \frac{m}{m+n}$$

Applications –

It can be used for AC loads with ~~[illegible]~~ high time constant

eg It can be used for a big size of motor with high MOI & mech time constant.

Limitations –

We cannot get wide range of control ie control is limited.

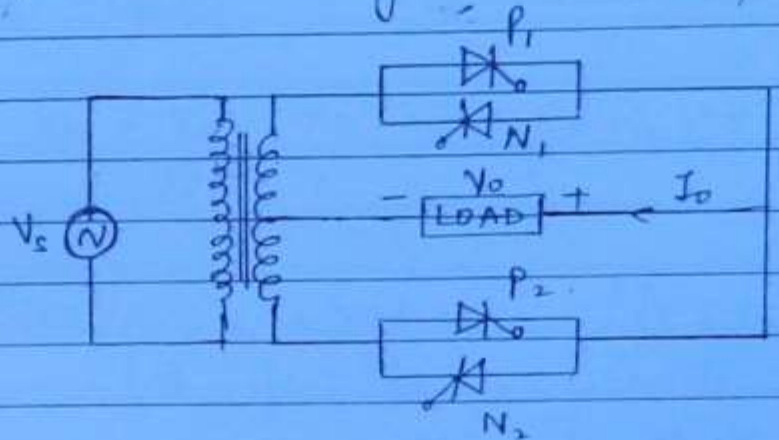
CYCLOCONVERTER

Fixed Ac ⟶ Variable Ac
(V_s, f_s) (V_0, f_0)

$f_0 < f_s \Rightarrow$ step-down cycloconverter

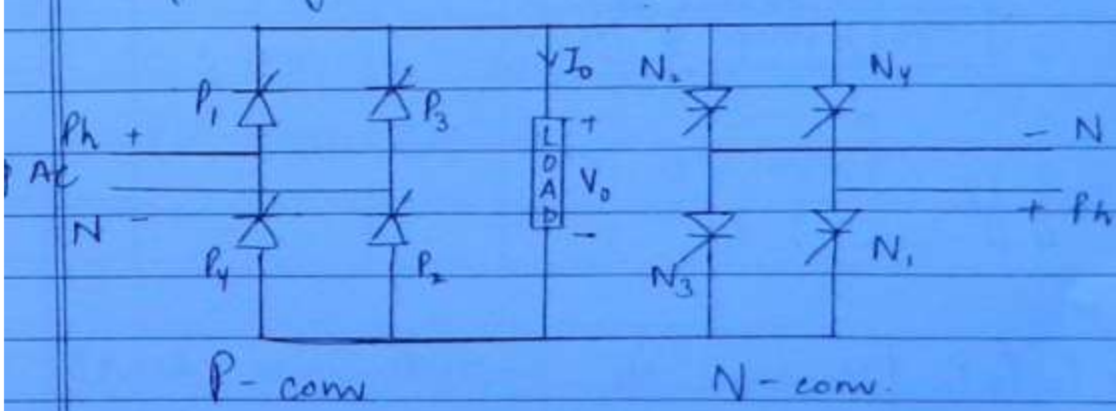
$f_0 > f_s \Rightarrow$ step up cycloconverter

Mid - Point Cycloconverter -



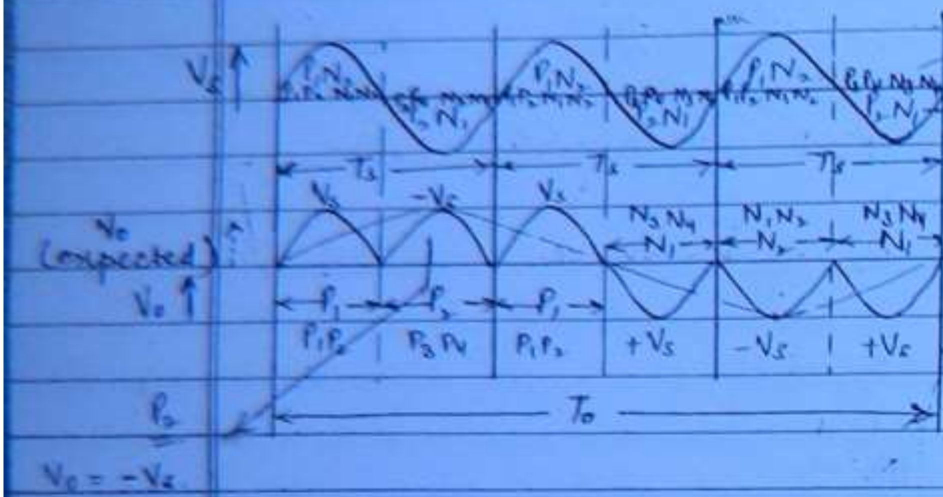
Bridge Cycloconverter

+ $P_1 P_2$ $N_1 N_2$ (F
− $P_3 P_4$ $N_3 N_4$ (F



Stepdown cycloconverter ($f_0 < f_s$)

let $f_0 = \frac{1}{3} f_s$ i.e. $T_0 = \frac{3}{8} T_s$



→ Bridge cycloconv

→ Mid pt cycloconv

For +ve cycle of expected consider P thyristors of V_s.

Shortcut →

⇒ for $f_0 = \frac{1}{3} f_s$

$P_1\ P_2\ P_1 \mid N_1\ N_2\ N_1$

$+\ -\ +\ \mid -\ +\ -$ → supply

[for -ve supply]

$\boxed{\text{-ve group is } N_1}$

⇒ for $f_0 = \frac{1}{4} f_s$

$P_1\ P_2\ P_1\ P_2 \mid N_2\ N_1\ N_2\ N_1$

$+\ -\ +\ -\ \mid +\ -\ +\ -$

for +ve supply

$\boxed{\text{-ve group is } N_2}$

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Here only frequency is varied as α is kept const but if variation in both V_0 & f_0 is required with this scheme vary α.

R-load -

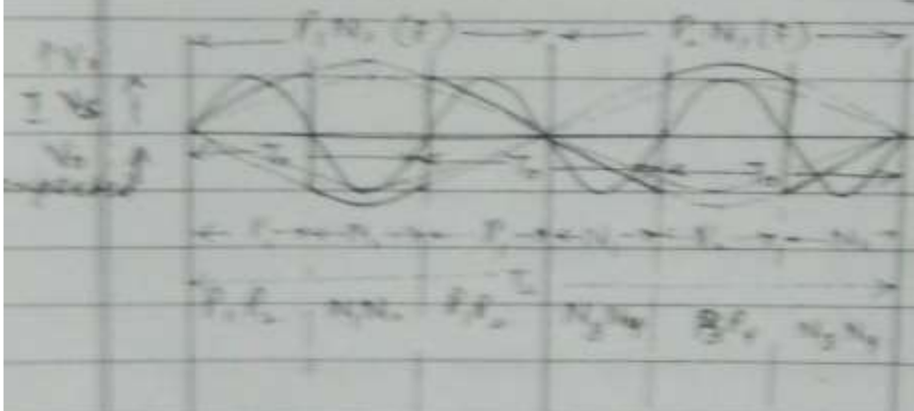
$$V_{0r} = \frac{V_m}{\sqrt{2\pi}}\left[(\pi-\alpha) + \frac{1}{2}\sin 2\alpha\right]^{1/2}$$

RL - Load -

$$V_{0r} = \frac{V_m}{\sqrt{2\pi}}\left[(\beta-\alpha) + \frac{1}{2}(\sin 2\alpha - \sin 2\beta)\right]$$

Step Up cycloconverter - ($f_0 > f_s$)

Let $f_0 = 3f_s$ $\therefore T_0 = \frac{1}{3}T_s$ $\therefore T_s = 3T_0$



- Forced commutation is required in step up cycloconverter.

- Harmonic distortion is more in cycloconverters hence they operate at low pf.

Applications -

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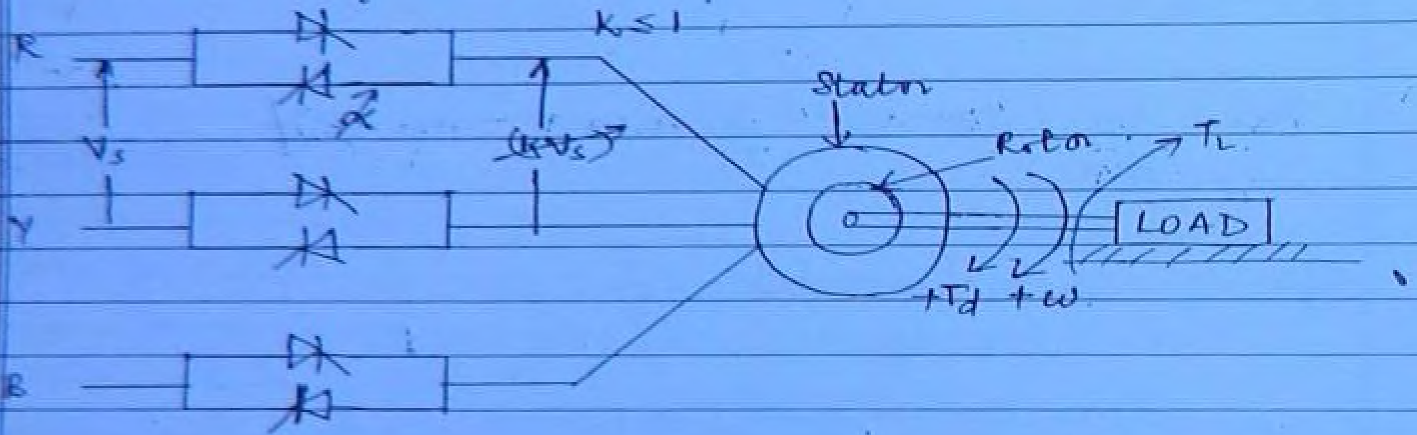
Its used for high ~~spee~~ power, low speed and reversible AC drives
eg In 3ϕ I.M for controlling speed.

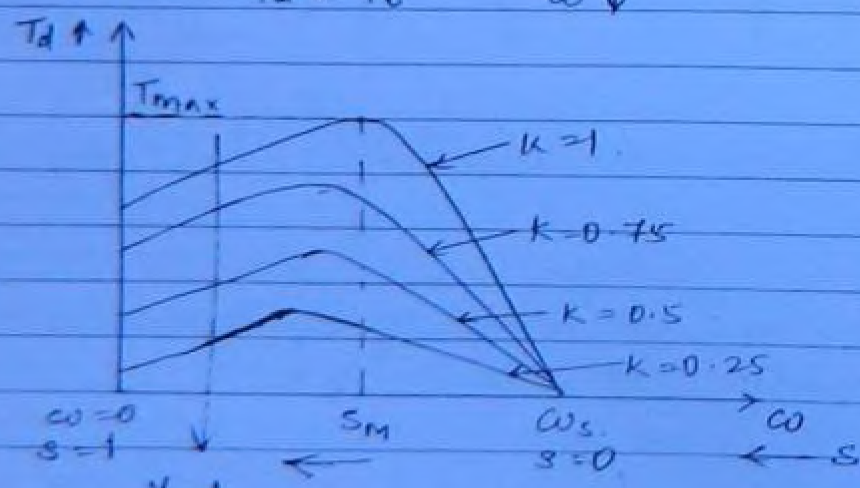
AC DRIVES -

1. Stator voltage control of I.M -



At starting $T_d > T_L$ $\omega \uparrow$
After reaching steady state speed $T_d = T_L$

$T_d < T_L$ $\omega \downarrow$



$$T_d = \frac{3}{\omega_s} \frac{(KV_s)^2 \; R_r'/s}{\left(R_s + \frac{R_r'}{s}\right)^2 + (X_s + X_r')^2}$$ 197

$$s = \frac{N_s - N}{N_s} = \frac{\omega_s - \omega}{\omega_s}$$

Mech Loads –

1. $T_L = \text{const.}$
2. $T_L \propto \omega$
3. $T_L \propto \omega^2$
4. $T_L \propto \frac{1}{\omega}$

1.⇒ Check whether constant load torque is suitable or not suitable for the given electrical drive

Let us consider the m/c is running at rated speed

$\therefore T_d = T_L$

$(KV_s)\downarrow$, $T_d \downarrow$ $(T_d < T_L)$

$\therefore \omega \downarrow \; S \uparrow \; I \uparrow \; \therefore T_d \uparrow$

Here m/c will slow down & ↑ the T_d until it balances the constant T_L.
Here the m/c draws more current from the supply mains in order to ↑ the T_d for balancing the T_L ∴ w/g's gets overheating
Hence this mech load is not suitable for the drive.

$\Rightarrow \quad T_L \propto \omega^2$

$(\downarrow V_s)\downarrow \quad T_d \downarrow \quad (T_d < T_L)$

$\omega \downarrow \quad T_L \downarrow$

Here speed will slow down until the decreased T_L balances the T_d. Here the m/c will not draw more current from supply line.

So this type of mech load is suitable for given electrical drive.

Applications -

Used where $T_L \propto \omega^2$

eg fan loads, reciprocating pumps, compressors etc.

2 Stator frequency control of I.M -

(a) $\frac{V}{f}$ control $(\omega < \omega_s)$

To maintain const flux we go for V/f control.

If ϕ is not maintained const

overfluxing occurs

- $\rightarrow I_m \uparrow$
- $\rightarrow$ harmonic distortion $\uparrow$
- $\rightarrow$ losses $\uparrow$
- $\rightarrow$ pf $\downarrow$

$\rightarrow$ We can ~~not~~ realise V/f control by using cycloconverters but its used only for high power low speed. With cycloconverter the maximum speed is limited to 40% of rated speed.

→ We can also realise the V/f control by using PWM inverters. With PWM inverters —
↳ smooth rotation is possible ag some lower order harmonics can be eliminated

(b) Constant V (for $\omega > \omega_r$)

At rated speed $V_s = V_{rated}$

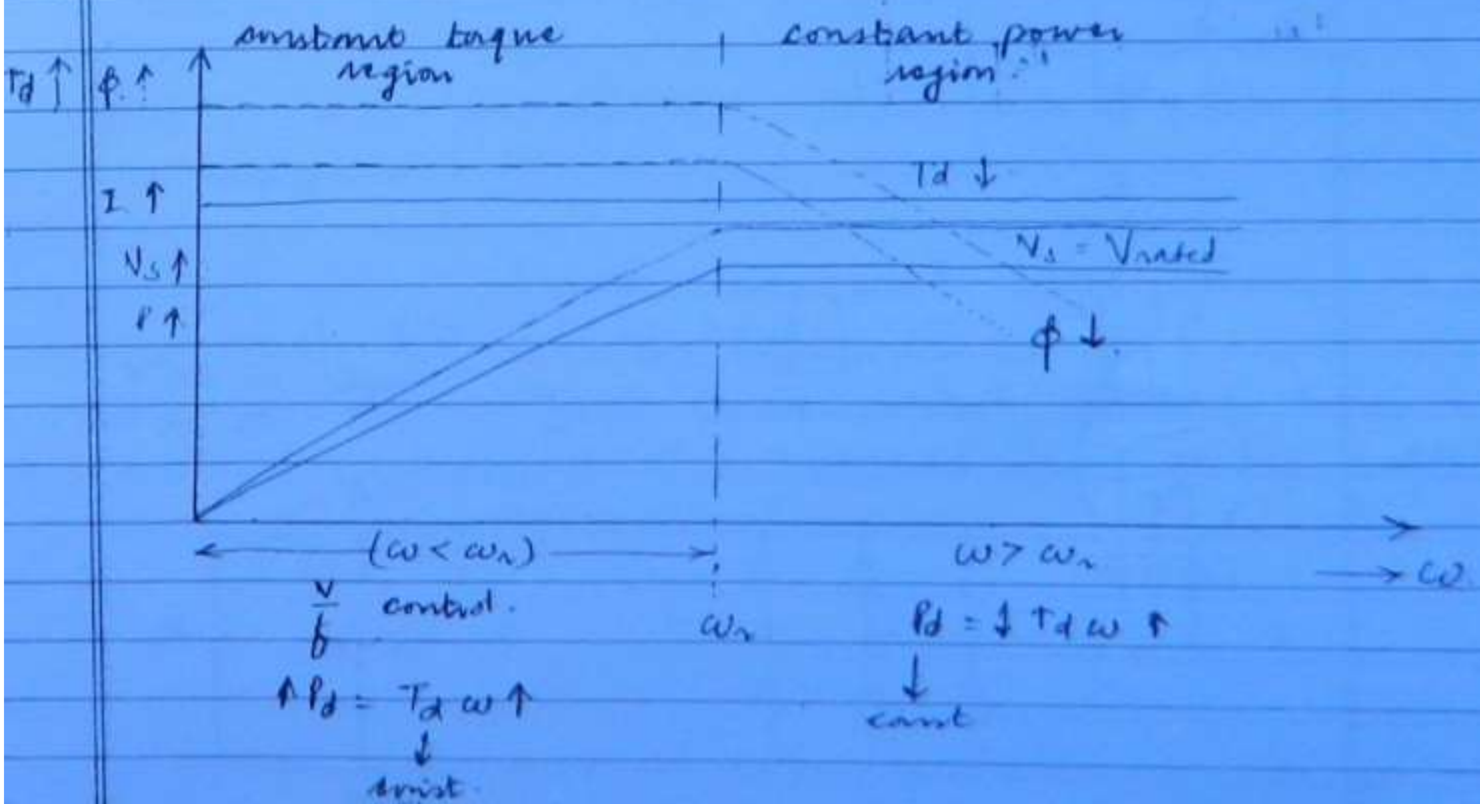
$V_s \propto \downarrow \phi \, f \uparrow$

$N \uparrow \quad f \uparrow \; \therefore \phi \downarrow$

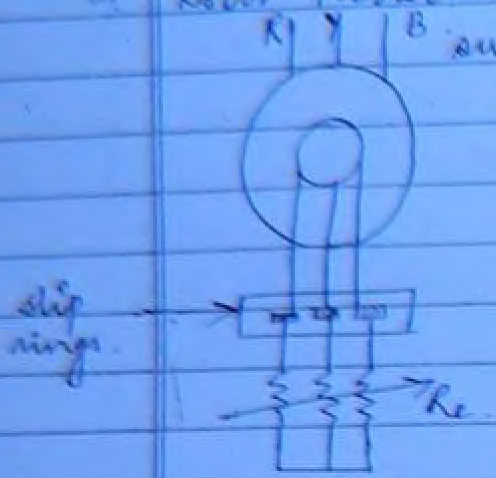
Here as $f \uparrow \Rightarrow \phi \downarrow$ because we cannot ~~exceed~~ ↑ the stator vlg beyond rated value ∴ stator vlg is fixed at rated value in this method

We can realise this method by:

→ Square Wave Inverter
→ PWM Inverter



Rotor Resistance Control -

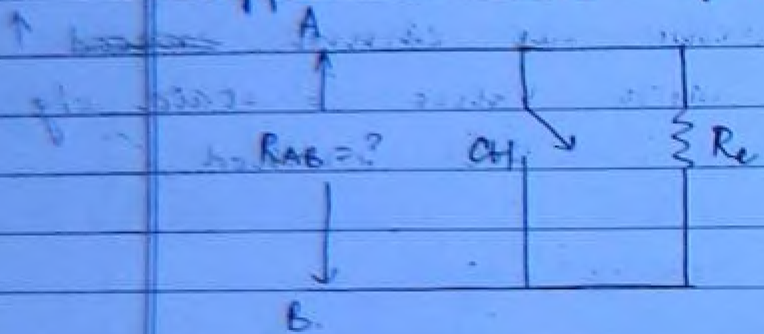


Total Cu loss = sP_g

$$3I_r^2[R_r + \uparrow R_e] = \uparrow sP_g$$

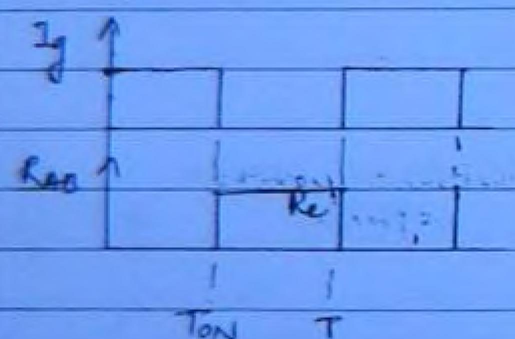
$s\uparrow \therefore \omega\downarrow$

Chopper Controlled Resistance -



$$R_{AB} = R_e\left(\frac{T_{OFF}}{T}\right)$$

$$R_{AB}\nearrow = R_e(1-\alpha\nearrow)$$



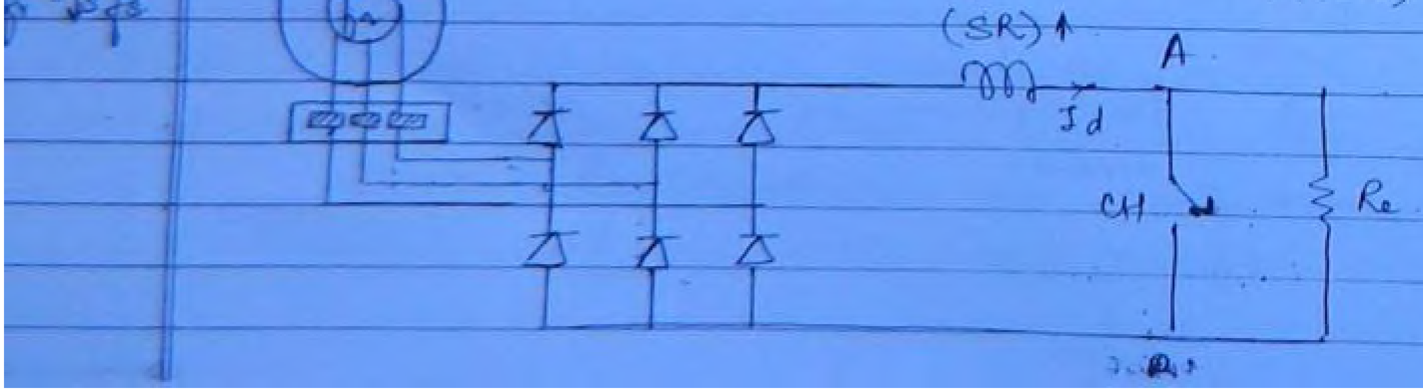
Stator Rotor Resistance Control

$\boxed{f_0 = 6.sf_s}$

$S\downarrow$ $f_0\downarrow$

ripple in $I_d\uparrow$

$\therefore (SR)\uparrow$



Q What's the effective resistance connected in series per phase with the rotor wdgs per phase for the above system? (201)

$$\text{Total Cu loss} = sP_g$$

$$3I_r^2 R_r + I_d^2 R_e(1-\alpha) = sP_g$$

$$I_{sr} = I_0\sqrt{\frac{2}{3}} \Rightarrow I_r = I_d\sqrt{\frac{2}{3}}$$

$$I_d = I_r\sqrt{\frac{3}{2}}$$

$$\Rightarrow 3I_r^2 R_r + \frac{3}{2}I_r^2 R_e(1-\alpha) = sP_g$$

$$\Rightarrow 3I_r^2 [R_r + \underbrace{0.5R_e(1-\alpha)}_{}] = sP_g$$

$$\cancel{3I_r^2 [R_r + (C)]}$$

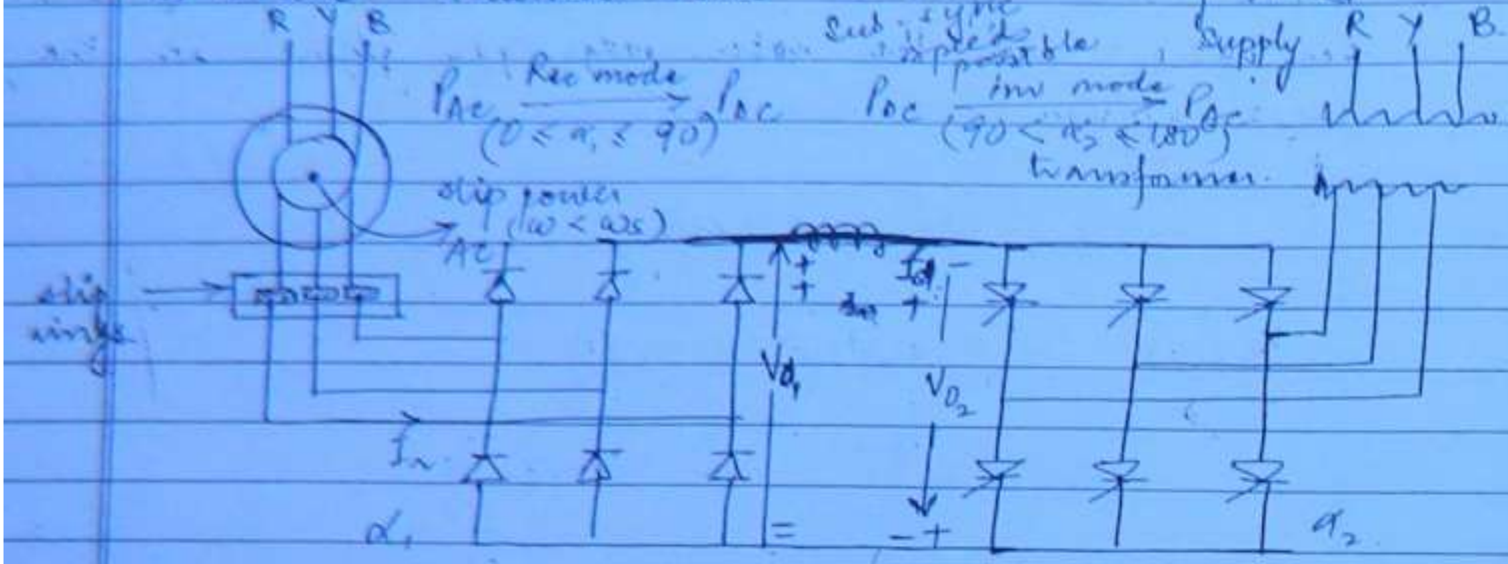
eff resistance in series with rotor wdgs per phase

This is not an efficient control cuz the slip power is dissipated in external resistance

4. Slip Power Recovery

This is an efficient speed control method because the slip power can be utilised or given back to supply line.

a) Static Kramer Drive



$\pm$ } Ref polarities of converter (201)

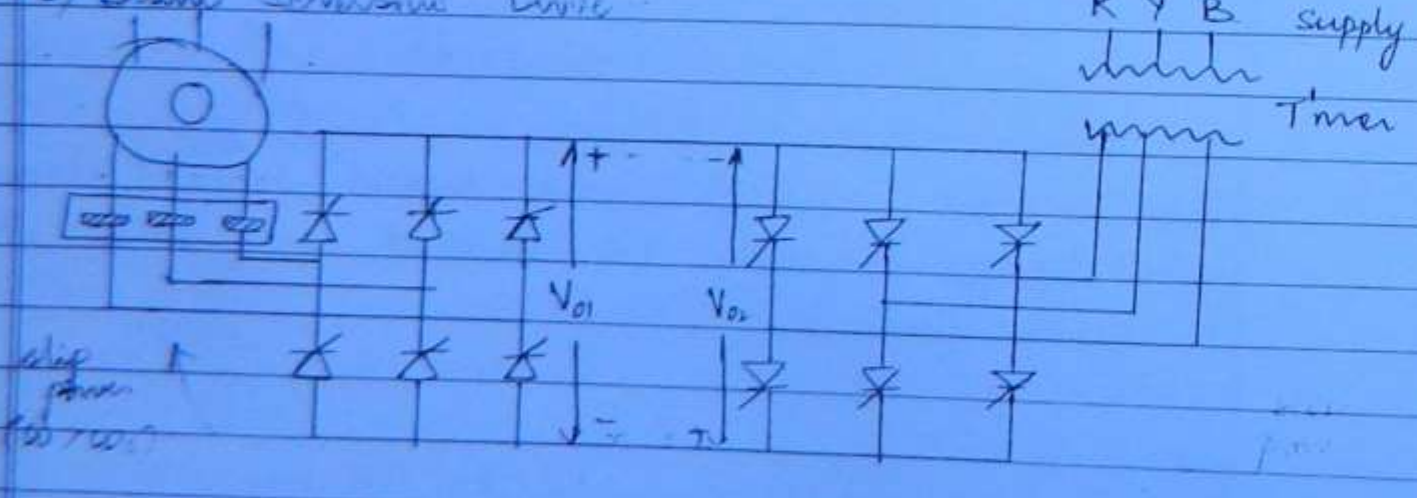
$\pm$ } Actual polarities of converter.

$$V_0 = \frac{3 V_{ML}}{\pi} \cos \alpha$$

$\alpha < 90^\circ$ V_0 + (Rec mode)

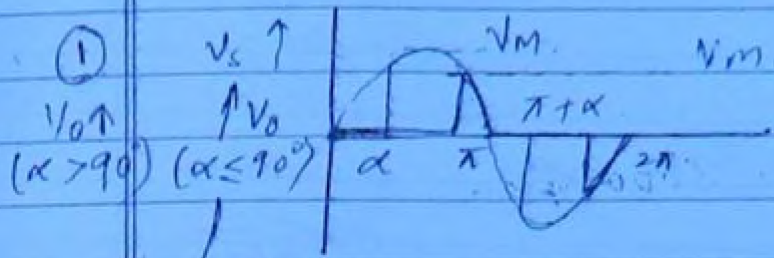
$\alpha > 90^\circ$ V_0 − (Inv mode)

b) Static Scherbius Drive



CWB chapter 5.

(1)



$V_m \sin \alpha$

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$$\text{peak power} = \frac{V_m^2}{R} = \frac{(230\sqrt{2})^2}{R} = \frac{(230\sqrt{2})^2}{10}$$

$$= 10580\ W \quad (d)$$

$$\text{peak power} \Rightarrow \frac{(V_m \sin \alpha)^2}{R}$$

(2) V_0 is uncontrolled at $\alpha \le \phi$

$$\phi = \tan^{-1}\frac{\omega L}{R} \Rightarrow \phi = \tan^{-1}\frac{50}{50}$$

$$= 45°$$

$$0 \le \alpha \le 45° \quad (a)$$

(6) ~~used~~ Per unit power $= \dfrac{P_0}{P_{max}}$

$$= \frac{V_{or}^2 / R}{V_{sr}^2 / R}$$

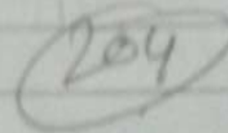
$$= \left(\frac{V_{or}}{V_{sr}}\right)^2 = PF^2$$

$$PF = \sqrt{\text{per unit power}} \quad (b)$$

DNR chapter 6

1) $\frac{V}{f}$ control of I.M –

$$N_s = \frac{120 f}{P} = \frac{120 \times 50}{2} = 3000 \text{ rpm}$$



$$s = \frac{N_s - N_r}{N_s} = \frac{3000 - 2850}{3000}$$

$s = 0.05$

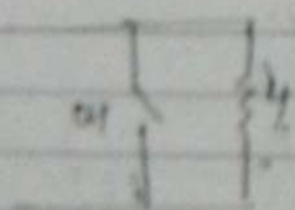
$N = ?$ at $1/2$ S_n $\quad f = 40\ Hz$

$$N_s = \frac{120 \times 40}{2} = 2400$$

$$0.05 = \frac{2400 - N_r'}{2400}$$

$N_r' = 2280$ rpm (c)

2) 2Ω – in series with rotor



$$R_{eff} = 2 + 0.5 R_o (1 - \alpha)$$

$$= 2 + 0.5 \times 4 \left(1 - \frac{T_{off}}{T}\right)$$

$$= 2 + 0.5 \times 4 \left(1 - 4 \times 10^{-3} \times 200\right)$$

$$= \frac{12}{5} \quad (a)$$

(14) PF at $\frac{1}{2} N_x$?

$$PF = \frac{3}{\pi} \cos\alpha$$

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$$\cancel{S = \frac{N_s - N_x}{N_s}}$$

$$V_0 = E_b + I_a R_a^{\,\approx 0}$$

$$\cancel{PF = \frac{V_0}{V_{max}}} = \cancel{g}$$

$$V_0 = \frac{3 V_{mL}}{\pi} \cos\alpha = E_b$$

At $\frac{1}{2} N_x$

$$\frac{3 V_{mL}}{\pi} \cos\alpha = \frac{440}{2}$$

$$\frac{3 \cdot 440\sqrt{2}}{\pi} \cos\alpha = \frac{440}{2}$$

$$\cos\alpha = \frac{\pi}{2\sqrt{2}} \times \frac{1}{3}$$

$$PF = \frac{3}{\pi} \cos\alpha = \frac{1}{2\sqrt{2}} = 0.354 \text{ (A)}$$

(15) As $\alpha \uparrow$ ripple $\uparrow$ smoothness $\downarrow$

At $\alpha \downarrow$ (smoothness of V_0 waveform) $\uparrow$

$\alpha \downarrow$ $V_0 \uparrow$ $\therefore \omega \uparrow$

higher speed.

(16) Regenerated power $= V_0 I_0$

$= V_0 (1-\alpha) I_0$

$= 600(1-0.7)100$

$= 18$ kW (C)

TRENDS IN TRANSMISSION OF POWER

1. EHVAC

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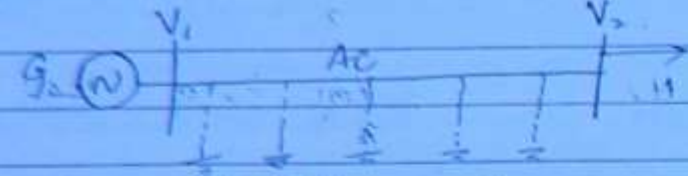
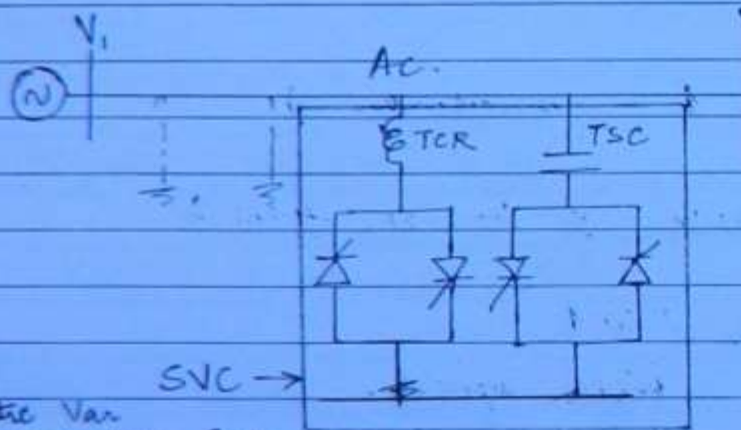
Features -



lightly Loaded $V_2 \uparrow > V_1$

$$P = \frac{V_1 V_2}{X} \sin \delta$$

* → We cannot control the power flow mag^n & dir^n quickly & easily



→ lightly loaded $\left(V_2 \uparrow > V_1\right)$ TCR

→ Heavily loaded $\left(V_2 \downarrow < V_1\right)$ TSC

Static Var Compensation System

Thyristorised Controlled Reactor → TCR

Thyristorised Switched Capacitance → TSC

* → There is continuous variation of reactive power flow in the line & hence responsible for voltage fluctuations & additional power loss

* → Intermediate substations are installed in the AC line for every 200 - 300 kms to compensate the

reactive power as per the requirement of reactive power in the line.

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* → Other problems in the AC line is
  - → Skin Effect
  - → Corona loss.

* → Area (1) —— AC —— Area (2)

Synchronous Tie

(frequencies must be equal)

Area (1) —— HVDC —— Area (2)

Asynchronous Tie

(frequencies in both areas must not be equal)

* → System disturbance in one of the area leads to power swings. If the power swings are unstable that may lead to cascaded tripping of alternators (if protection system fails)

* → With AC interconnection frequency disturbance is carried forward to other areas.

* → If multiple number of independent areas is interconnected by AC lines then the fault level of the system increases.

2. HVDC

* HVDC is economical to transmit large amount of power over long distance.

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* Advantages -

1. The power flow magnitude & direction can be quickly and easily controlled

2. The transient stability limit is improved.

3. We can fast clear the fault in HVDC line

4. There is no Skin Effect problem & Corona loss is reduced.

5. HVDC can utilise Earth for its return path.

6. The phase to phase clearance, phase to ground clearance & tower height requirement is lesser in HVDC line.

7. Power handling capacity of a Bipolar HVDC line is almost twice that of 3 ϕ single circuit AC lin

8. We can interconnect independent areas at different frequencies because it is an asynchronous tie.

9. Frequency disturbance is not transferred to other independent areas, with HVDC interconnection

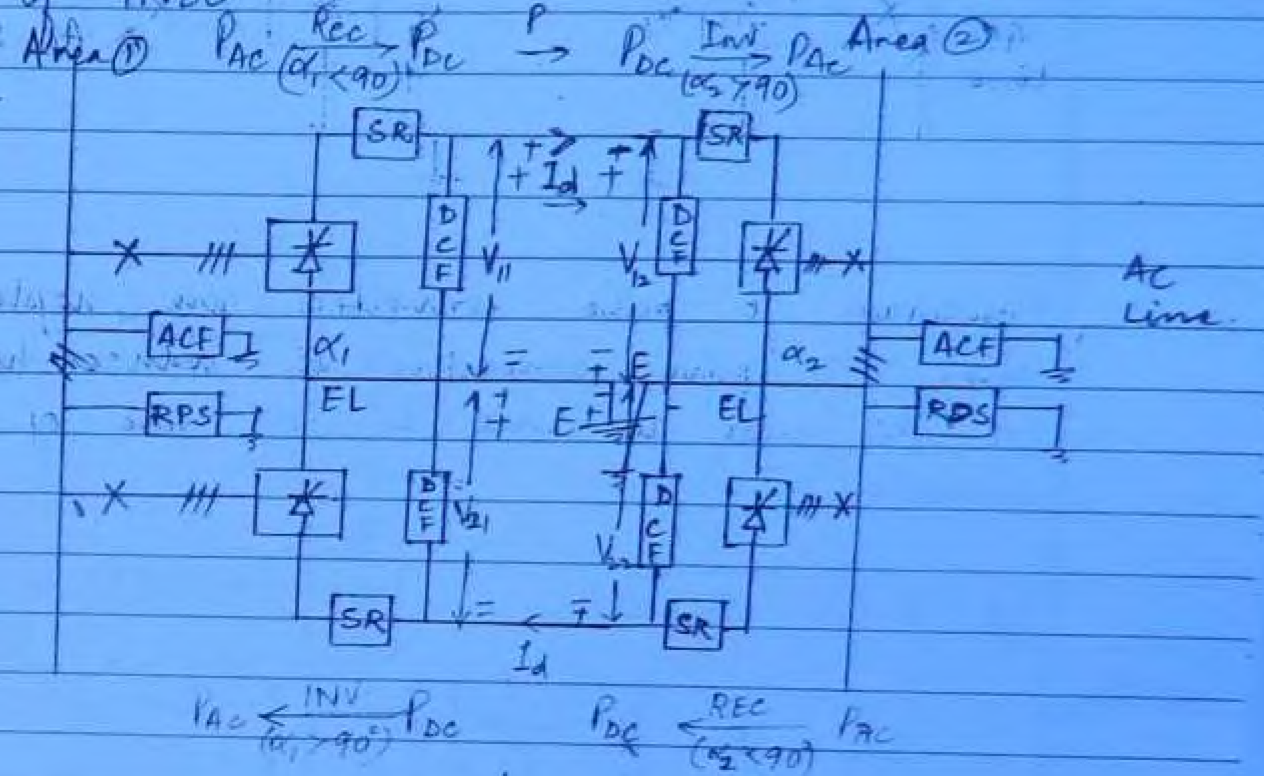
10. If multiple no. of independent areas is

interconnected by HVDC line the fault level of the system will not substantially increase.

11. HVDC is used for underground or submarine cables even for short distance because there is no continuous charging of DC cables.

* Types of HVDC

BIPOLAR



Smoothing Reactor (SR)
Used for smoothing

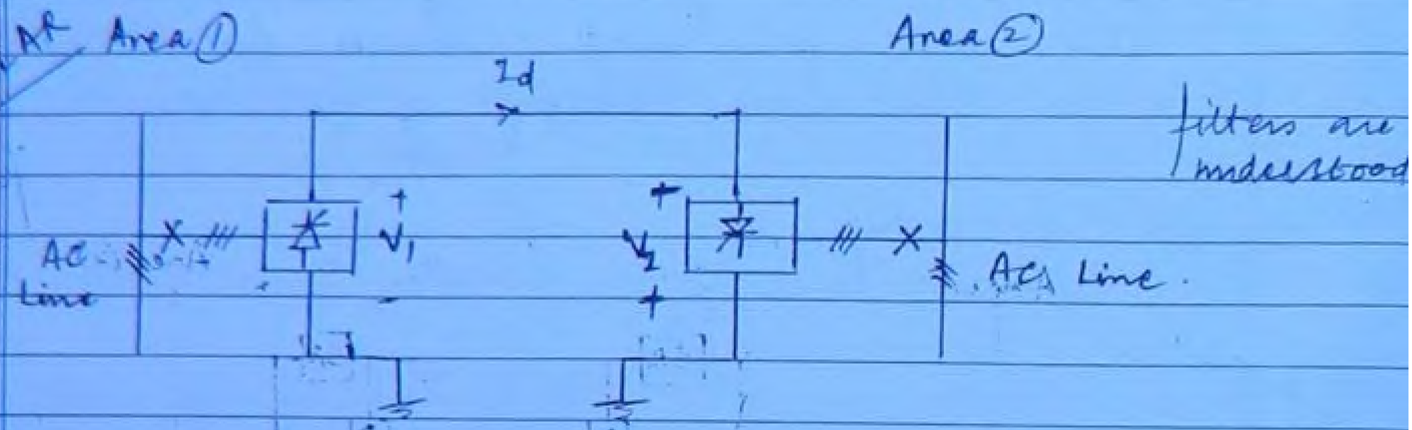
DC Filter (DCF)
Reduces harmonics on DC side of the converter (output side)

AC Filter (ACF)
Reduces harmonics on AC side of the converter

Reactive Power Source (RPS)

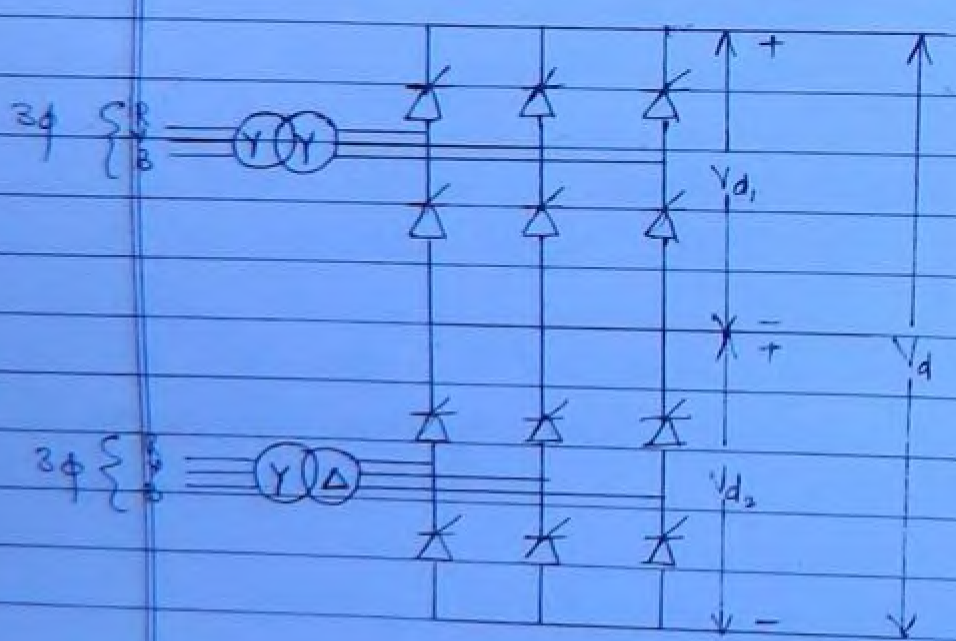
To compensate the reactive power required for converter

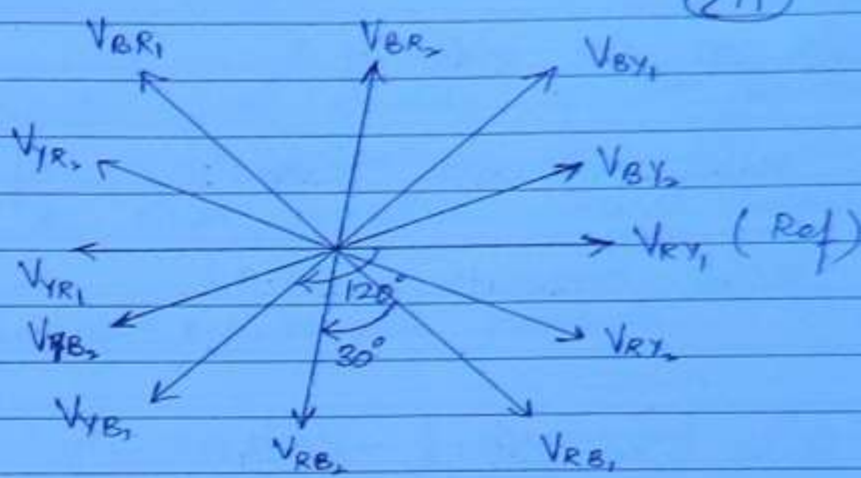
MONOPOLAR

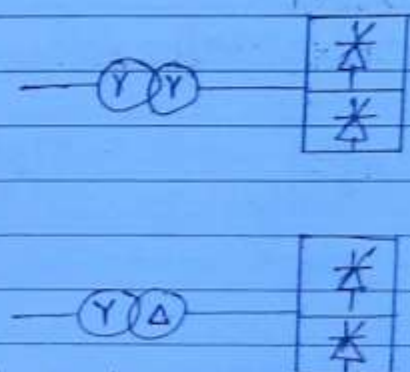


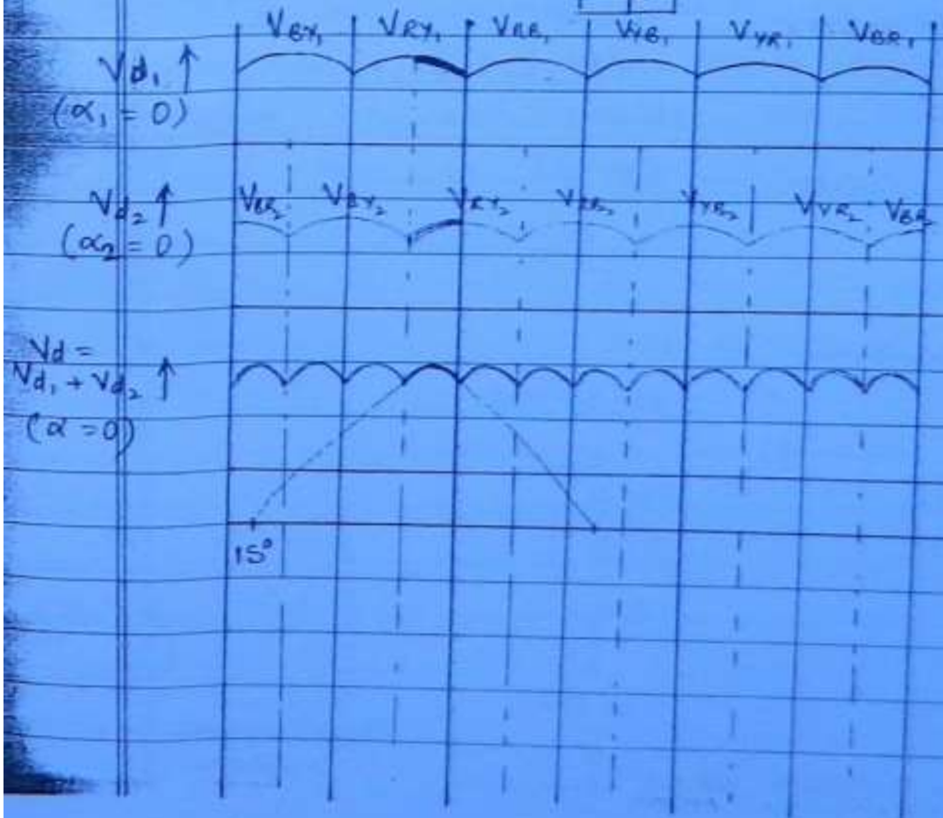
Nowadays 6 pulse converters are replaced by using 12 pulse converters to reduce harmonics on AC side as well as DC side of the converter.

12 Pulse Converters –









$V_{RY_1} = V_{ML} \sin\omega t$

$V_{RY_2} = V_{ML} \sin(\omega t - 30^\circ)$

$V_{RY} = V_{RY_1} + V_{RY_2}$

$= V_{ML}[\sin\omega t + \sin(\omega t - 30^\circ)]$

$= V_{ML}[\sin(\omega t - 15^\circ)]$

* Harmonics on AC side of :

m pulse converter → $mk \pm 1$

2 pulse converter → $2k \pm 1$
$= 3, 5, 7, 9, 11 \ldots$

6 pulse converter → $6k \pm 1$
$= 5, 7, 11, 13, 17, 19 \ldots$

12 pulse converter → $12k \pm 1$
$= 11, 13, 23, 25, \ldots$

* Harmonics on DC side of :

m pulse converter → mk

2 pulse converter → $2k$
$= 2, 4, 6, 8, 10, 12 \ldots$

6 pulse converter → $6k$
$= 6, 12, 18 \ldots$

12 pulse converter → $12k$
$= 12, 24, 36, \ldots$

SMPS

* SMPS provides good quality of DC power supply required for some applications like ICs, digital circuits & other sensitive circuit boards.

* SMPS operates on chopper principle.

* At very high switching frequency, the ripple is almost reduced.

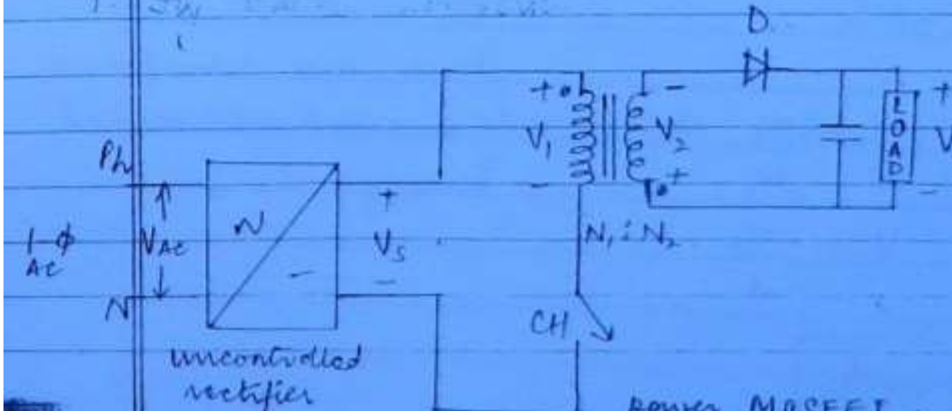
* At such a high frequency we can reduce the size of the filter as well as transformer.

* In SMPS the transistor operates in the switched mode (Cut-off region is used for OFF state & Saturation region for ON state).

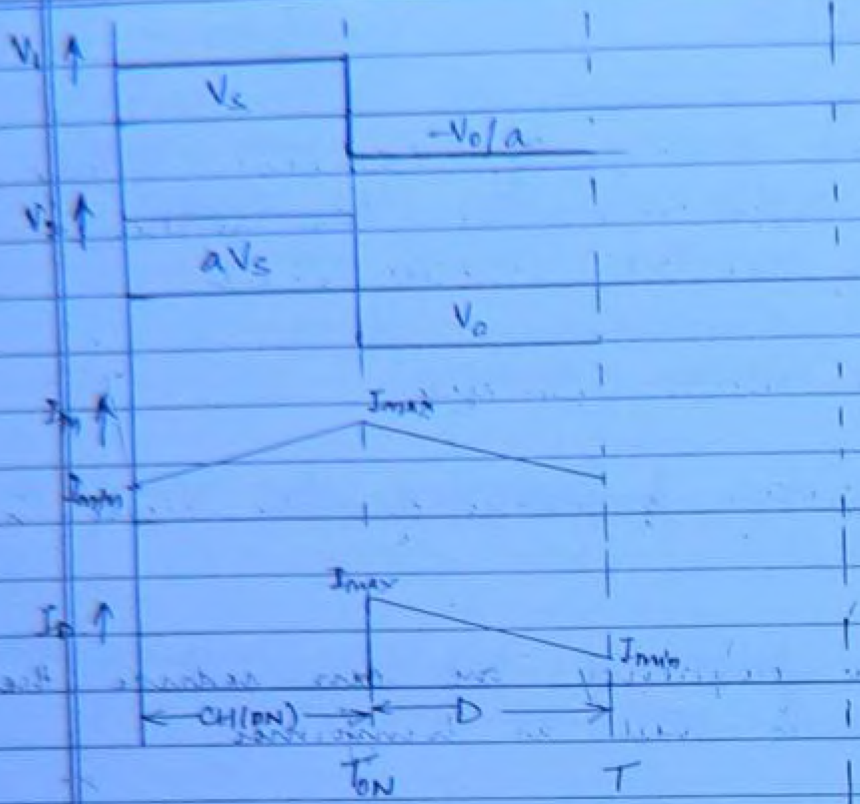
* SMPS is more efficient & compact in size compared to linear power supplies. In linear power supplies transistor operates in the active region & hence Power loss is higher.

Types of SMPS –

1. [illegible]



$\therefore a = \frac{N_2}{N_1}$



$\alpha = \frac{T_{ON}}{T}$

$$\underset{50\,Hz}{AC} \xrightarrow{1^{st}} DC \xrightarrow{2^{nd}} AC(f\uparrow) \xrightarrow{3^{rd}} DC$$

① $0 \leq t \leq T_{ON}$

CH → ON $\qquad V_1 = V_s$

D → OFF $\qquad V_2 = \frac{N_2}{N_1} V_1$

$$V_2 = aV_1$$

transformer stores energy ∴ $I_m \uparrow$

CH is in ON state

D is RD ∴ $I_D = 0$

② $T_{ON} \leq t \leq T$

CH → OFF $\qquad V_2 = -V_0$

D → ON $\qquad V_1 = \frac{N_1}{N_2} V_2$

$$V_1 = \frac{-V_0}{a}$$

Here the transformer releases the stored energy. ∴ $I_m \downarrow$

Avg vtg ⇒ $\boxed{V_0 = a \frac{\alpha V_s}{1-\alpha}}$

CWB chapter 7.

(4) $V_0 = 2\, a \cdot \dfrac{\alpha V_s}{1-\alpha}$ 215

$= 2 \dfrac{N_2}{N_1} \dfrac{T_{ON}}{T_{OFF}}$ (c)

Peak Forward Blocking Voltage of chopper switc

$= V_s + \dfrac{V_0}{a}$

(3) $PFB\ v/g = V_s + \dfrac{V_0}{a}$

$= V_s + V_s\left(\dfrac{\alpha}{1-\alpha}\right)$

$= V_s\left[1 + \dfrac{\alpha}{1-\alpha}\right] = 115\sqrt{2}\left[1 + \dfrac{0.3}{1-0.3}\right]$

$= 232.34\ V$ (a)

Output dc v/g of front end rectifier

→ without v/g doubling = V_m → peak Ac i/p v/g

→ with v/g doubling = $2V_m$

(1) mark the highest freq given

2. Push-pull converter

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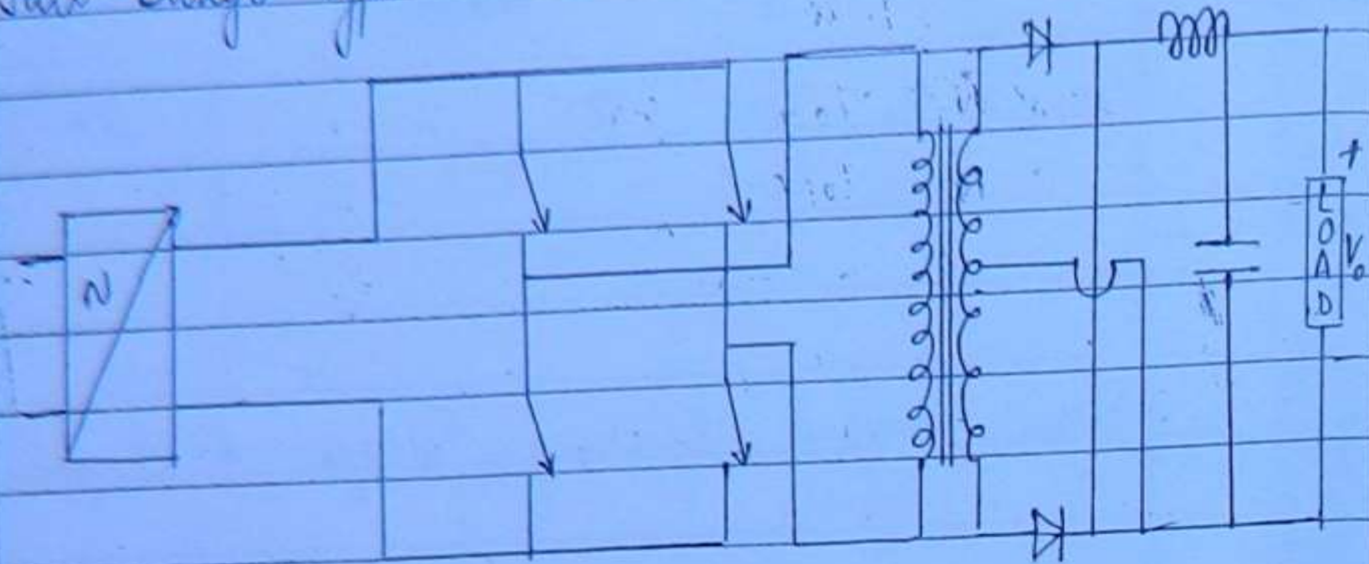
3. Full Bridge Type -



Remembering Formulas -

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R-Load -

$V_0 = \frac{V_M}{2\pi}(1+\cos\alpha) \rightarrow$ 1 pulse

cont. $V_0 \propto \cos\alpha$.

discont $V_0 \propto (1+\cos\alpha)$

$V_0 = \frac{V_m}{\pi}(1+\cos\alpha) \rightarrow$ 2 pulse

$V_0 = \frac{V_{mph}}{2\pi/3}\left[1+\cos(\alpha+30^\circ)\right] \rightarrow (\alpha > 30^\circ)$

3 pulse.

In 3 pulse

$\rightarrow \alpha < 30$ cont.

$\rightarrow \alpha > 30$ discont.

~~$V_0 = \frac{V_{mph}}{2\pi/6}\left[1+\cos(\alpha+60^\circ)\right] \rightarrow (\alpha > 60^\circ)$~~

In 6 pulse

$\rightarrow \alpha < 60^\circ$ cont

$\rightarrow \alpha > 60^\circ$ discont.

$V_0 = \frac{V_{ML}}{\pi/6}\left[1+\cos(\alpha+60^\circ)\right] \rightarrow (\alpha > 60^\circ)$