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PART - A

2 Marks Questions

1.
Sol. The triac is a three terminal, four layer, bidirectional, semi-conductor device that controls ac power whereas an SCR controls dc power or forward biased half cycles of ac in a load. Because of its bidirectional conduction property, the triac is widely used in the field of power electronics for control purposes.
2.
Sol. A diac is an important member of the thyristor family and is usually employed for triggering triacs. A diac is a two electrode bidirectional avalanche diode which can be switched from offstate to the on-state for either polarity of the applied voltage.
3.
Sol. **SCS (Silicon controlled switch)**, like an SCR, is a unilateral, four-layer, three junction P-N-P-N silicon device, with four electrodes, namely, cathode, cathode gate, anode gate and anode. It is a low power device compared with the SCR. A few of the more common areas of application of SCS include few of the more common areas of application of SCS include a variety of computer circuits (such as counters, registers, and timing circuits) voltage sensors, pulse generators, oscillators.
4.
Sol. GTO is, like an SCR, a four-layer, three junction semiconductor device with three external terminals, GTO, like an SCR, can be turned on by applying a positive gate signal. However, a GTO can be turned off by a negative pulse applied to the gate. It has improved switching characteristics. GTO devices are now used in (i) high performance drive systems (ii) traction purposes and (iii) adjustable frequency inverter drives.
5.
Sol. Light activated SCR (LASCR) is just an ordinary SCR except that it can also be light triggered. It finds applications including optical light controls, relays, phase control, motor control and a large number of computer applications.
6.
Sol. Power transistors are generally of three types: BJTs, MOSFETs and IGBTs. BJTs are current-controlled devices and their parameters are sensitive to junction temperature. BJTs suffer from second breakdown and need reverse base current during turn-off to reduce the storage time, but they have low on-state or saturation voltage.
7.
Sol. It is resistance between B_1 and B_2 and it is of the order of 4-10 $k\Omega$. From the equivalent circuit it is obvious that $R_{BB} = R_{B2} + R_{B1}$.
8.
Sol. The voltage division factor is denoted by η and is known as intrinsic stand-off ratio. It is the property of UJT and lies between 0.5 and 0.8.
9.
Sol. The advantages of Power Electronics Devices are
 - High efficiency due to low loss in power semiconductor device.
 - Less heating and lesser cooling requirement.
 - Long life and lesser maintenance due to the absence of any moving parts.
 - Smaller size and less weight.
 - Faster response.



10.

Sol.

- Sinusoidal terms having frequency multiple of fundamental frequency.
- It downgrade the equality of power supply.
- Increases losses.
 $\uparrow P = I_r \uparrow m s^2 R$
- Due to triple harmonics current flowing through neutral conductor on account of neutral conductor heated.
- Overload, vibration & premature aging of machine.
- Disturbance with communication lines.

11.

Sol.

- Switch does not require any power to change its state.
- Completely controllable.
- Switch does not required an external signal.
- Low thermal impedance.

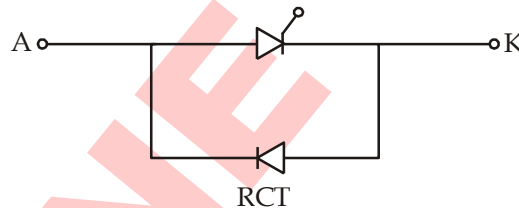
12.

Sol.

An RCT (Reverse Conducting Thyristor) can be employed in the place of antiparallel connection of SCR and diode to realise halfwave AC voltage controller. It may be considered as a thyristor with a built-in antiparallel diode.

13.

Sol.



14.

Sol.

Requirements for a SCR to be triggered by a gate pulse.

1. SCR should be forward bias mode.
2. A positive gate voltage between gate and cathode.
3. The gate pulse width should be chosen to ensure that anode current rises above the latching current.

15.

Sol.

If the forward voltage is increase upto V_{BO} breakdown will occurs junction (J_2) large number charge carriers are generated the anode current increases and SCR start conducting and SCR - start conducting forward current i.e. forward conduction mode.

16.

Sol.

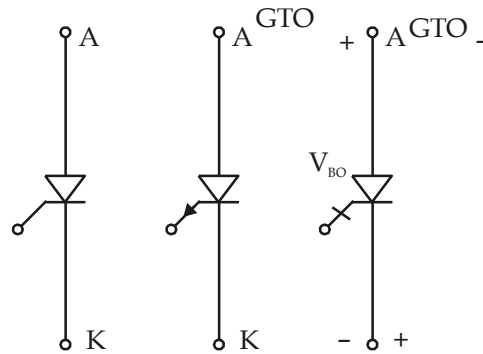
If +ve voltage applied between gate and cathode the +ve gate current flows from gate to cathode i.e. holes from G to K and e^- from K to G. ($K \rightarrow G$). As outer n-laye is highly dopped in compersion to inner P-layer the gate current flows mainly due to e^- .

When of the e^- enter the depletion region of P-layer, the width of depelction region is reduced so breakdown occurs at normal voltage.

Higher is the gate current more will be e^- entering the depletion region more reduction in the width at D.R the breakdown voltage V_{BO} is further reduced.

17.

Sol.

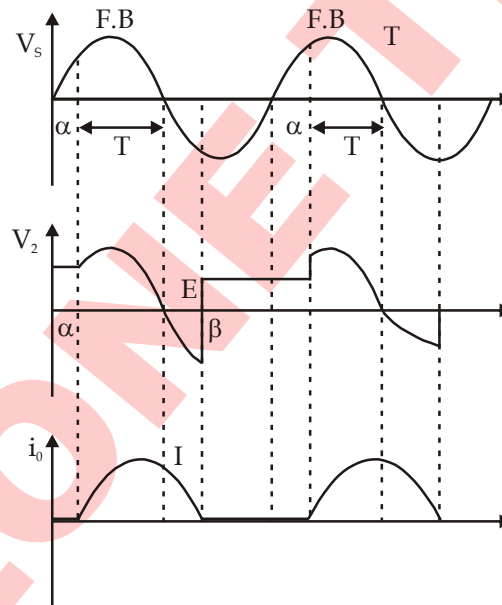


18.

Sol. The dc to ac power converters are known as inverters. In other words, an inverter is a circuit which converts a dc power into an ac power at desired output voltage and frequency.

19.

Sol.



20.

Sol.

- Input PF is improved
 - Output voltage wave is improved, $V_o \neq V_e$
 - i_0 wave form is improved i.e. closer to dc.
- 1- ϕ half wave rectifier $\Rightarrow$ No. of pulses in V_o in 2π is equal to 1.

21.

Sol.

A step-up chopper works on the principle of storing energy in an inductance to provide output voltage higher than source voltage. The output voltage of a step-up chopper is given by $V_a = \frac{V}{1-\alpha}$.



22.

Sol. In forced commutation, the commutating components L and C do not carry load current continuously. In this commutation method, forward current of the thyristor is forced to zero by external circuitry called commutation circuit.

Forced commutation methods are:

1. Resonant-pulse commutation (or class B commutation).
2. Complementary commutation (or class C commutation).
3. Impulse commutation (or class D commutation).

23.

Sol. This load voltage can be controlled by varying duty cycle α .

$$V_0 = f \cdot T_{on} \cdot V_s$$

24.

Sol. Step down cyclo converter operation is based on Line/Natural commutation.

25.

Sol. An ac regulator converts a constant ac voltage into a variable ac voltage of the same frequency. AC regulators using thyristors and triacs have the advantages of high efficiency, compact size and fast control.

26.

Sol. AC to AC converter converts either a constant voltage into a variable ac voltage of the same frequency or converts ac at one frequency to ac of another frequency. The former are known as ac voltage controllers or ac voltage regulators while the latter are known as cycloconverters.

27.

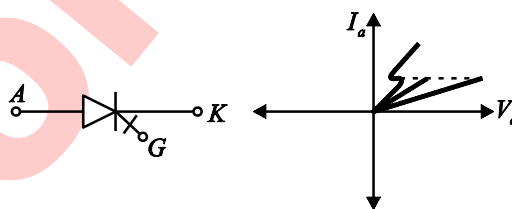
Sol. 6-pulse 3- ϕ to 3- ϕ cyclo converters require 36 thyristors.

PART - B

5 Marks Questions

28.

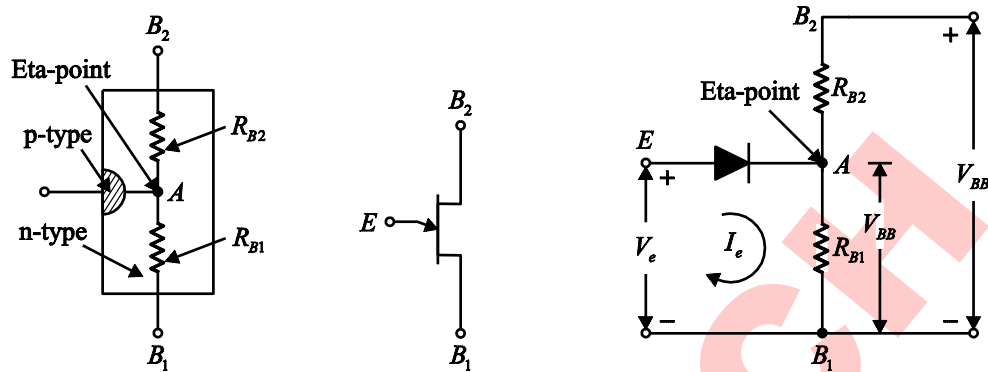
Sol. GTO is pnpn device, can be turned on like an ordinary thyristor by a pulse of positive gate current, but it can be turned off easily by a negative gate pulse of appropriate amplitude.



1. GTO has faster switching speed.
2. Its surge current capability is comparable with an SCR.
3. It has more di/dt rating at turn-on.
4. GTO unit has higher efficiency due to the elimination of forced commutation losses.
5. GTO has reduced acoustical and electromagnetic noise due to elimination of commutation chokes.
6. GTO circuit configuration has lower size and weight as compared to thyristor circuit unit.
 - In BJT switching, lateral current flow is the basic limiting factor in BJT performance. It causes lateral voltage drop which leads to emitter current crowding, which causes decrease in current gain. Excessive current crowding causes second breakdown and destruction of the device.
 - Turning off of BJT should be done in a controlled manner of negative base current to avoid excessive stored charge which may cause long turn off time and large power dissipation. These problems are avoided in a GTO in low power applications.

29.

Sol. An UJT is made up of an n-type silicon (si) base to which p-type emitter is embedded. It has three terminals the emitter (E), base one B₁, and base two B₂ between bases B₁ and B₂ the Unijunction behave like ordinary resistance.



Basic structure of UJT Symbolic representation Equivalent circuit

Figure (a), when source voltage V_{BB} is applied, capacitor C begins to charge through R exponentially towards V_{BB} . During this charging, emitter circuit of UJT is an open circuit.

The capacitor voltage v_c , equal to emitter voltage v_e , is given by

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

The time constant of the charge circuit is $\tau_1 = RC$.

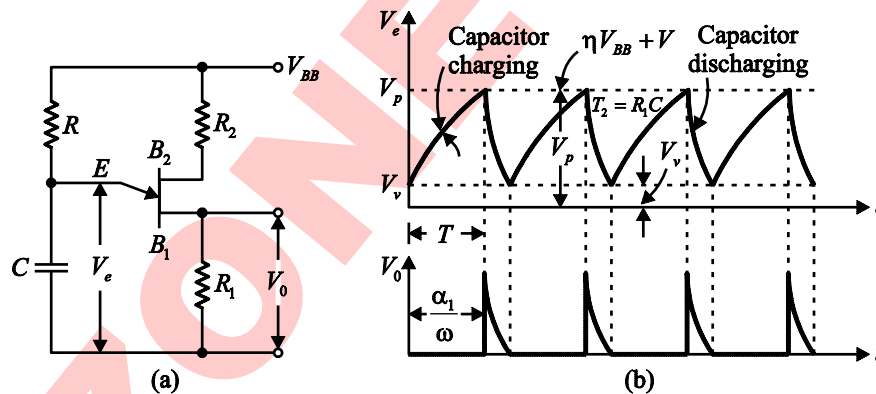


Fig. (2) UJT oscillator (a) Connection diagram and (b) Voltage waveforms

When this emitter voltage v_e (or v_c) reaches the peak-point voltage $V_p (= \eta V_{BB} + V_D)$, the Unijunction between E - B₁, breaks down. As a result, UJT turns on and capacitor C rapidly discharge through low resistance R_1 with a time constant $\tau_2 = R_1 C$. Here τ_2 is much smaller than τ_1 . When the emitter voltage decays to the valley-point voltage V_v , UJT turns off. The time T required for capacitor C to charge from initial voltage V_v to peak-point voltage V_p , through large resistance R, can be obtained as under:

$$V_p = \eta V_{BB} + V_D = V_v + V_{BB}(1 - e^{-T/RC})$$

Assuming $V_D = V_v$, $\eta = (1 - e^{-T/RC})$

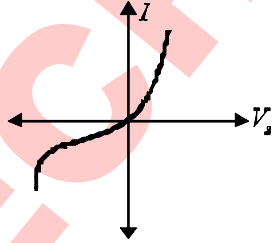
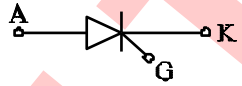
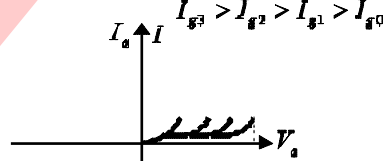
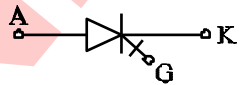
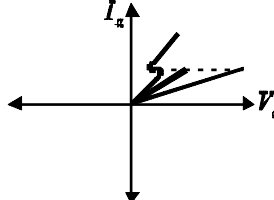
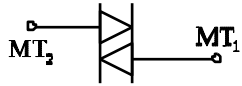
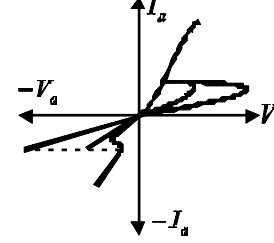
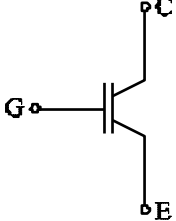
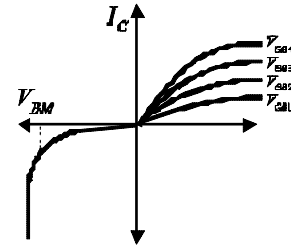
$$T = \frac{1}{f} = RC \ln \left(\frac{1}{1-\eta} \right)$$

In case T is taken as the time period of output pulse duration (neglecting small discharge time), then the value of firing angle α_1 is given by

$$\alpha_1 = \omega T = \omega RC \ln \frac{1}{1-\eta}$$

where ' ω ' is the angular frequency of UJT oscillator.

30.
Sol.

Device	Symbol	Characteristics
(i) Diode		
(ii) Thyristor		
(iii) GTO		
(iv) Triac		
(v) IGBT		

31.

Sol. Disadvantage of GTO compared to conventional thyristor :

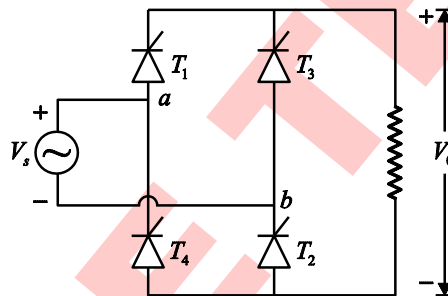
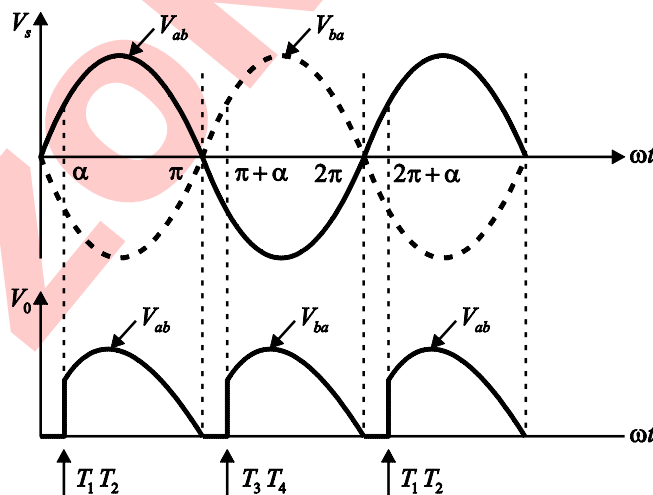
1. Magnitude of latching and holding currents is more in a GTO.
2. On state voltage drop and the associated loss is more in a GTO.
3. Due to multi cathode structure of GTO, Gate current is higher.
4. GTO has its reverse blocking capability less than its forward voltage blocking capability.
5. Gate drive circuit losses are more.

Advantage of GTO over conventional thyristor :

1. GTO has faster switching speed.
2. Its surge current capability is comparable with an SCR
3. It has more di/dt rating at turn-on.
4. Its circuit configuration has lower size and weight.
5. It has higher efficiency due to the elimination of forced commutation losses.
6. GTO unit has reduced electromagnetic noise due to elimination of commutation chokes.

Due to its faster switching speed and its elimination of forced commutation losses, mainly we use it in inverter and chopper circuits.

32.

Sol.**Single-phase full converter with resistive load**

Let, V_{or} = rms output voltage and $V_s = V_m \sin \omega t$

$$V_{or}^2 = \frac{1}{\pi} \int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \cdot d(\omega t) = \frac{V_m^2}{2\pi} \left[\omega t - \frac{\sin 2\omega t}{2} \right]_{\alpha}^{\pi}$$



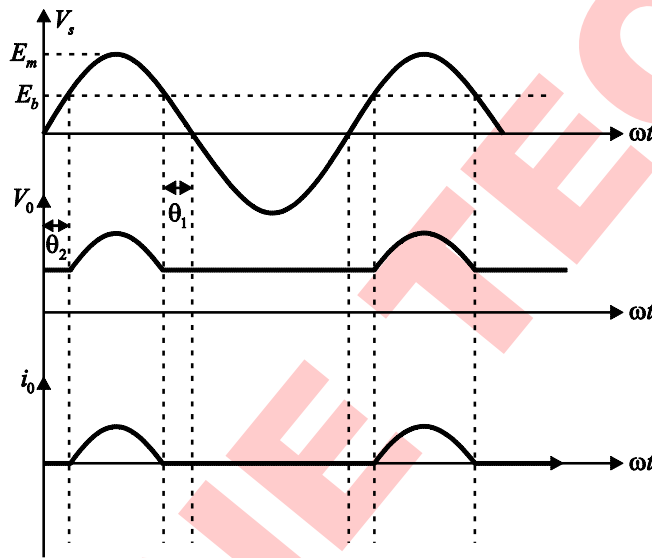
$$V_{or} = \frac{V_m}{\sqrt{2}} \left[\frac{1}{\pi} \left\{ (\pi - \alpha) + \frac{\sin 2\alpha}{2} \right\} \right]^{1/2}$$

For $V_m = 230 \times \sqrt{2}V$ and $\alpha = 30^\circ = \frac{\pi}{6}$

$$V_{or} = \frac{230 \times \sqrt{2}}{\sqrt{2}} \left[\frac{1}{\pi} \left\{ \left(\pi - \frac{\pi}{6} \right) + \frac{\sin 60}{2} \right\} \right]^{1/2}$$

$$= 226.66V \quad (\text{rms value of output voltage})$$

33.
Sol.



The voltage equation is,

$$E_m \sin \omega t = E_b + i_0 R$$

$$i_0 = \frac{E_m \sin \omega t - E_b}{R}$$

SCR is turned on when $E_m \sin \theta_1 = E_b$ and is turned off when $E_m \sin \theta_2 = E_b$

Where $\theta_2 = \pi - \theta_1$

The charging current of the battery is equal to average value of the output current.

$$\text{Where } i_0 = \begin{cases} \frac{E_m \sin \omega t - E_b}{R}, & \theta_1 < \theta < \pi - \theta_1 \\ 0, & \text{else} \end{cases} \quad (\because E_m = \sqrt{2}E)$$

$$\therefore \text{average current} = I_{0(av)} = \frac{1}{2\pi} \int_0^{2\pi} i_0 d\theta$$

$$= \frac{1}{2\pi R} \int_{\theta_1}^{\pi - \theta_1} (E_m \sin \omega t - E_b) d\theta$$

$$= \frac{1}{2\pi R} [2E_m \cos \theta_1 - E_b(\pi - 2\theta_1)]$$

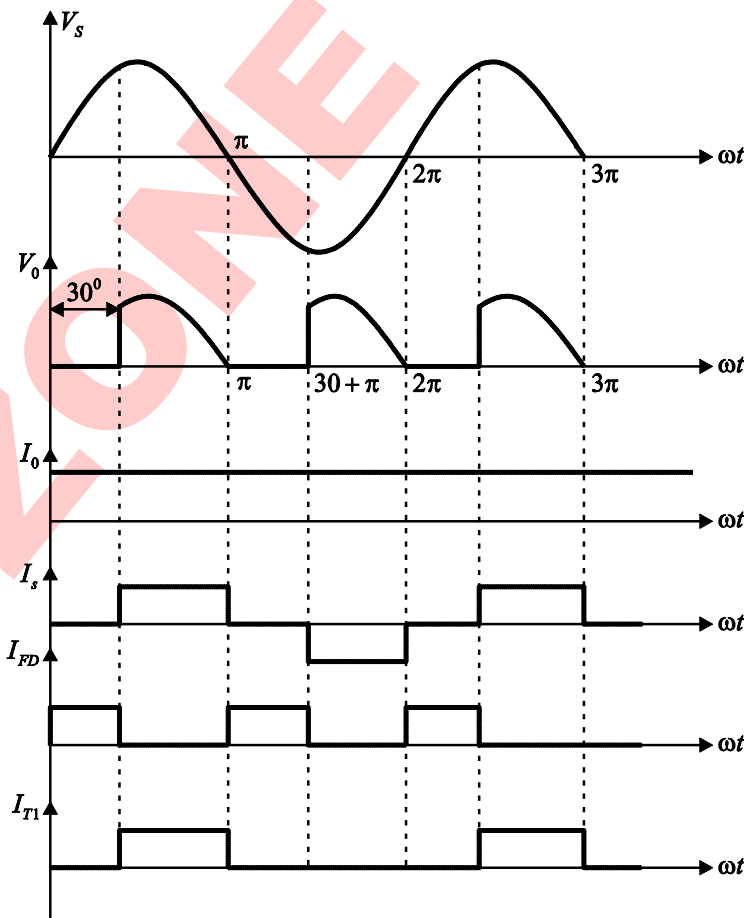
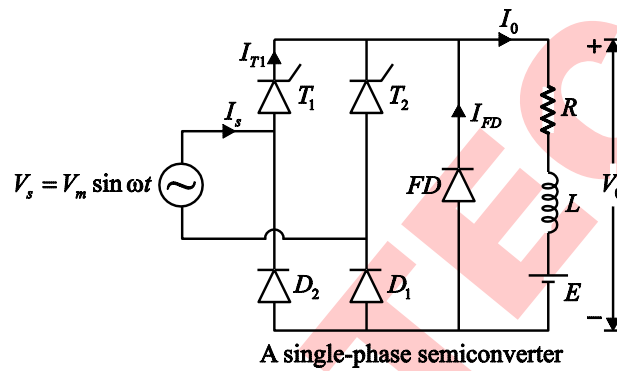
$$= \frac{1}{2\pi \times 10} [2 \times \sqrt{2} \times 220 \times \cos \theta_1 - 100(\pi - 2\theta_1)]$$

Where $\theta_1 = \sin^{-1} \left(\frac{100}{\sqrt{2} \times 220} \right) = 18.75^\circ = 0.327 \text{ rad}$

$\therefore I_{av} = 5.42$

The power supplied to battery $= E \times I_{av} = 100 \times 5.42 = 542 \text{ W}$

34.
Sol.



For single phase semiconverter, average output voltage

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

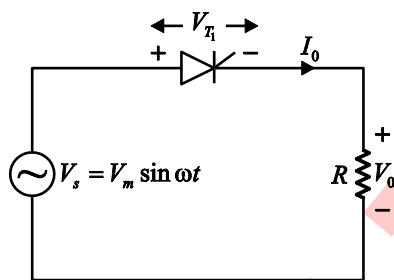
R.M.S. value of source current $I_{sr} = I_0 \sqrt{\frac{\pi - \alpha}{\pi}}$

and input power factor for ($\alpha = 90^\circ$)

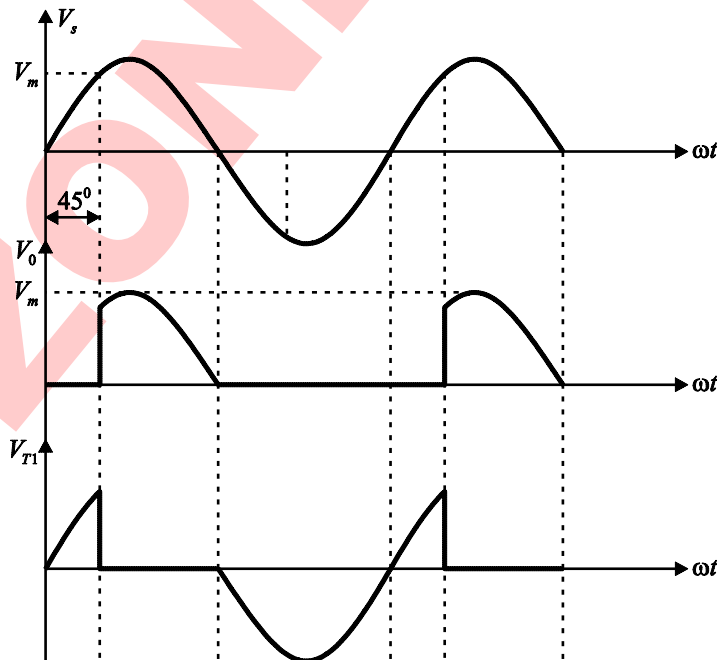
$$\text{Input power factor} = \frac{V_t \times I_0}{V_s \times I_{sr}} = \frac{\frac{V_m}{\pi} [1 + \cos 90] \times I_0}{\left(\frac{V_m}{\sqrt{2}}\right) \times \sqrt{\frac{\pi/2}{\pi}} \times I_0} = \frac{2}{\pi} = 0.63662$$

35.

Sol.



(i)



$$(ii) \quad V_{0(avg)} = \frac{1}{2\pi} \int_{\alpha}^{\pi} V_m \sin \omega t d(\omega t)$$

$$V_{0(avg)} = \frac{V_m}{2\pi} (1 + \cos \alpha) = \frac{\sqrt{2} \times 230}{2 \times \pi} (1 + \cos 45^\circ) = 88.37 \text{ volts}$$

$$I_{0(avg)} = \frac{V_{0(avg)}}{R} = \frac{88.37}{100} = 0.88 \text{ amp.}$$

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

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

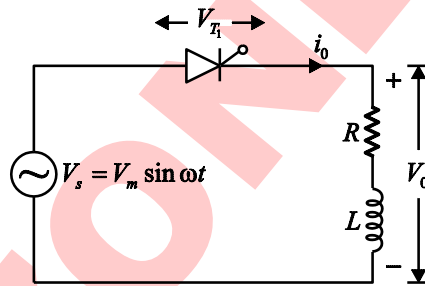
$$= \frac{230 \times \sqrt{3}}{2 \times \sqrt{\pi}} \left[\left(\pi - \frac{\pi}{4} \right) + \frac{1}{2} \times \sin 2 \times 45^\circ \right]^{1/2} = 155.1 \text{ Volts}$$

$$I_{0(rms)} = \frac{V_{0(rms)}}{R} = \frac{155.1}{100} = 1.55 \text{ amp.}$$

$$\text{Ripple factor} = \frac{\sqrt{V_{0(rms)}^2 - V_{0(av)}^2}}{V_{0(av)}} = 1.44$$

36.

Sol.



Half wave rectifier with (RL) load,

Circuit turn-off time

$$t_c = \frac{2\pi - \beta}{\omega} = \frac{(360 - 120)\pi}{180 \times 2\pi \times 50} = 13.33 \text{ m-sec}$$

Average output voltage

$$V_0 = \frac{V_m}{2\pi} [\cos \alpha - \cos \beta] = \frac{\sqrt{2} \times 230}{2\pi} [\cos 45^\circ - \cos 120^\circ] = 65.54 \text{ volts}$$

and Average load current

$$I_0 = \frac{V_0}{R} = \frac{65.54}{5} = 13.10 \text{ Amps}$$

37.

Sol. Average output voltage of 3- ϕ full converter,

$$V_0 = \frac{3V_{ml}}{\pi} \cos \alpha \quad \text{for} \quad 0 < \alpha < \pi/3$$

For $\alpha = 30^\circ$, $V_0 = 300$

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

$$300 = \frac{3V_{ml}}{\pi} \quad \dots\dots (i)$$

Average output voltage of 3- ϕ full converter, for $\frac{\pi}{3} < \alpha < \frac{2\pi}{3}$

$$V_0 = \frac{3V_{ml}}{\pi} \left[1 + \cos \left(\alpha + \frac{\pi}{3} \right) \right] = 300 \times \left[1 + \cos \left(\frac{\pi}{2} + \frac{\pi}{3} \right) \right] = 40.19 \text{ volts}$$

Hence, option (B) 40 volts is correct.

38.

Sol. Capacitor C will charge upto maximum voltage of the ac source through load resistor R_L and discharge through resistor R.

$$\text{Discharge current} = I_d = \frac{V_m}{R} = \frac{\sqrt{2}V_s}{R}$$

$$I_d = 2 \text{ A}$$

$$R = \frac{\sqrt{2}V_s}{I_d} = \frac{\sqrt{2} \times 100}{2} = 70.71 \Omega$$

$$\left(\frac{dV}{dt} \right)_{\max} = \frac{0.632 V_m}{R_L C}$$

$$\Rightarrow C = \frac{0.632 \times \sqrt{2} \times 100}{10 \times 50} = 0.178 \mu\text{F}$$

39.

Sol. For a step up chopper,

$$V_0 = \frac{V_s}{1-\delta}$$

$$\therefore 600 = \frac{200}{1-\delta}$$

$$\therefore \delta = \frac{2}{3}$$

$$\therefore \delta = \frac{T_{ON}}{T} = \frac{2}{3} = \frac{T - T_{OFF}}{T}$$

$$\therefore T_{OFF} = \frac{T}{3}$$

$$\therefore T = 3 \times 50 \mu\text{sec} = 150 \mu\text{sec}$$

$$\text{and } T_{ON} = T - T_{OFF} = 100 \mu\text{sec}$$



Now if desired output voltage is (300 V), then

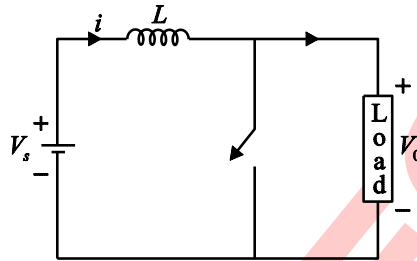
$$V_0 = 300 = \frac{V_s}{1-\delta} = \frac{200}{1-\delta}$$

$$\therefore \delta = \frac{1}{3}$$

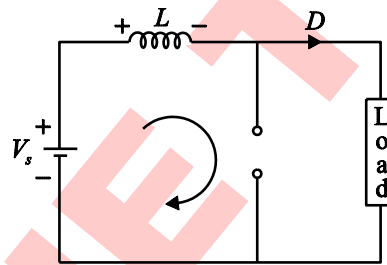
$$\therefore T_{ON} = \frac{T}{3} = \frac{150 \mu\text{sec}}{3} = 50 \mu\text{sec}$$

40.

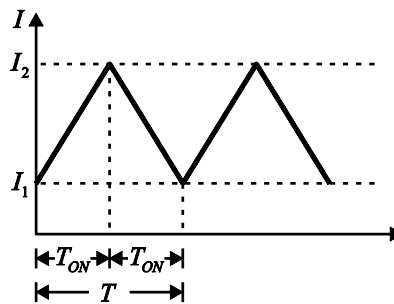
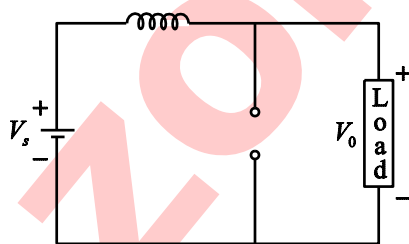
Sol. Basic circuit of step-up chopper



On the time of (T_{ON})



On the time of (T_{off})



During period T_{ON} , the energy given to inductor from source is,

$$W_{ON} = V_s \left(\frac{I_1 + I_2}{2} \right) T_{ON}$$

and during period ' T_{OFF} ' the energy release by inductor to the load is

$$W_{OFF} = (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) T_{OFF}$$

Considering the system to be lossless,



$$V_s \left(\frac{I_1 + I_2}{2} \right) T_{ON} = (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) T_{OFF}$$

$$V_0 = V_s \cdot \frac{T}{T - T_{ON}} = \frac{V_s}{1 - \delta}$$

If $V_s = V$, $V_0 = 2\text{ V}$ (given)

Thus, $(1 - \delta) = \frac{V}{2V}$

$$\therefore \delta = \frac{1}{2}$$

Thus for a duty cycle by having $T_{ON} = T_{OFF}$ the voltage can be stepped from $\frac{1}{2}V$ to $2V$.

41.

Sol. Given that $I_f = 26\text{ mA} = 26 \times 10^{-3}\text{ A}$

Mean life time of holes $\tau = 20\text{ }\mu\text{sec} = 20 \times 10^{-6}\text{ sec}$

$$T = 300^\circ\text{K}$$

We know that C_D (diffusion capacitance) $= \tau \cdot g$

where τ = is mean carrier life time

g = dynamic conduction

and equal to $g = \frac{I_f}{\eta V_T}$

where η utility factor and for G_e diode $\eta = 1$

$$V_T = 26\text{ mV at } 300^\circ\text{K}$$

$$V_T = 26 \times 10^{-3}\text{ V}$$

$$\therefore C_D = \frac{\tau \cdot I_f}{\eta V_T}$$

Putting the corresponding value we get

$$C_D = \frac{20 \times 10^{-6} \times 26 \times 10^{-3}}{1 \times 26 \times 10^{-3}} = 20 \times 10^{-6}\text{ F}$$

$$C_D = 20\text{ }\mu\text{F}$$

42.

Sol. When chopper is on, output voltage is $(V_s - 1.5)$ volts and during the time chopper is off, output is zero

$$\therefore \text{Average output voltage} = \left(\frac{(V_s - 1.5) T_{on}}{T} \right) = (200 - 1.5) \times 0.5 = 99.25\text{ V}$$

R.M.S. value of output voltage

$$V_{OR} = \left[(V_s - 1.5)^2 \cdot \frac{T_{on}}{T} \right]^{1/2} = \left[(200 - 1.5)^2 \times 0.5 \right]^{1/2} = 140.36\text{ Volt}$$

Power output or power delivered to load

$$P_0 = \frac{V_{OR}^2}{R} = \frac{140.36^2}{20} = 985\text{ Watts}$$

Power input to chopper

$$P_i = V_s \times I_0 = 200 \times \frac{99.25}{20} = 992.5 \text{ W}$$

$$\therefore \text{Chopper efficiency} = \frac{P_o}{P_i} \times 100 = \frac{985}{992.5} \times 100 = 99.24\%$$

43.

Sol. For RL load, $\sin(\beta - \phi) = \sin(\alpha - \phi) e^{\left[\frac{R}{L} \frac{(\alpha - \beta)}{\omega} \right]}$

Where, $\beta \rightarrow$ extinction angle, $\alpha \rightarrow$ firing angle and conductance angle, $r = \beta - \alpha$, ϕ is pf angle

For $(R = 0)$, (purely inductive load)

$$\sin(\beta - \phi) = \sin(\alpha - \phi)$$

$$\sin \beta = \sin \alpha$$

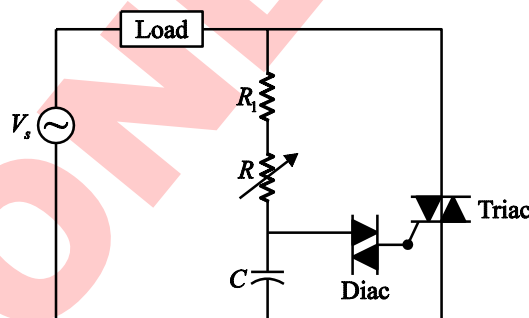
$$\therefore \beta = \alpha$$

Then, $r = \text{conduction angle} = \beta - \alpha = 0$

For another SCR, $r' = \pi - \alpha = 180 - 75 = 105^\circ$

44.

Sol. This is the circuit diagram of the simplest fan regulator. The circuit is based on the principle of power control using a Triac. The circuit works by varying the firing angle of the Triac. The firing angle can be varied by varying the resistance R. By varying the value of R, the firing angle of Triac changes. This directly varies the load power, since load is driven by Triac. The firing pulses are given to the gate of Triac using Diac.



45.

Sol. The average output voltage of 1- ϕ semiconductor

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

Maximum possible value of $V_o = \frac{2V_m}{\pi}$ for $\alpha = 0$

$$0.25 \times \left(\frac{2V_m}{\pi} \right) = \frac{V_m}{\pi} (1 + \cos \alpha)$$

$$1 + \cos \alpha = 0.5$$



⇒

$$\cos \alpha = -0.5$$

$$\alpha = 120^\circ$$

$$\text{Average load current } I_0 = \frac{V_0}{R} = \frac{V_m}{\pi R} (1 + \cos \alpha) = \frac{\sqrt{2} V_s}{\pi R} (1 + \cos \alpha)$$

$$= \frac{\sqrt{2} \times 240}{\pi \times 100} \times (1 + \cos 120^\circ) = 0.54 \text{ Amp.}$$

$$\text{Rms load current} = I_{or} = \frac{V_{or}}{R} = \frac{V_s}{R} \left[\frac{1}{\pi} \left\{ (\pi - \alpha) + \frac{\sin 2\alpha}{2} \right\} \right]^{1/2}$$

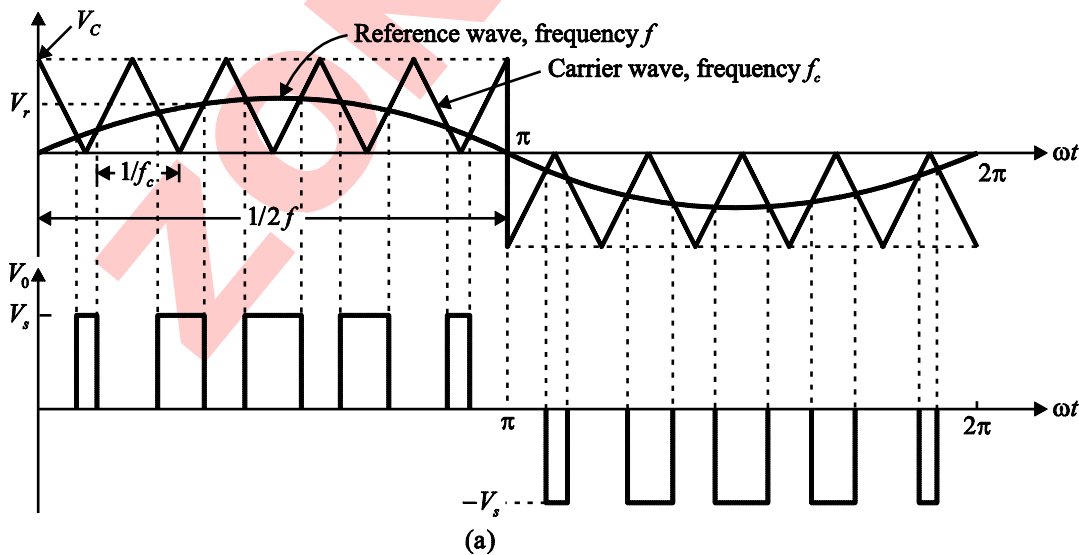
$$= \frac{240}{100} \times \left[\frac{1}{\pi} \times \left\{ \left(\pi - \frac{2\pi}{3} \right) + \frac{\sin 240}{2} \right\} \right]^{1/2} = 1.06 \text{ Amp.}$$

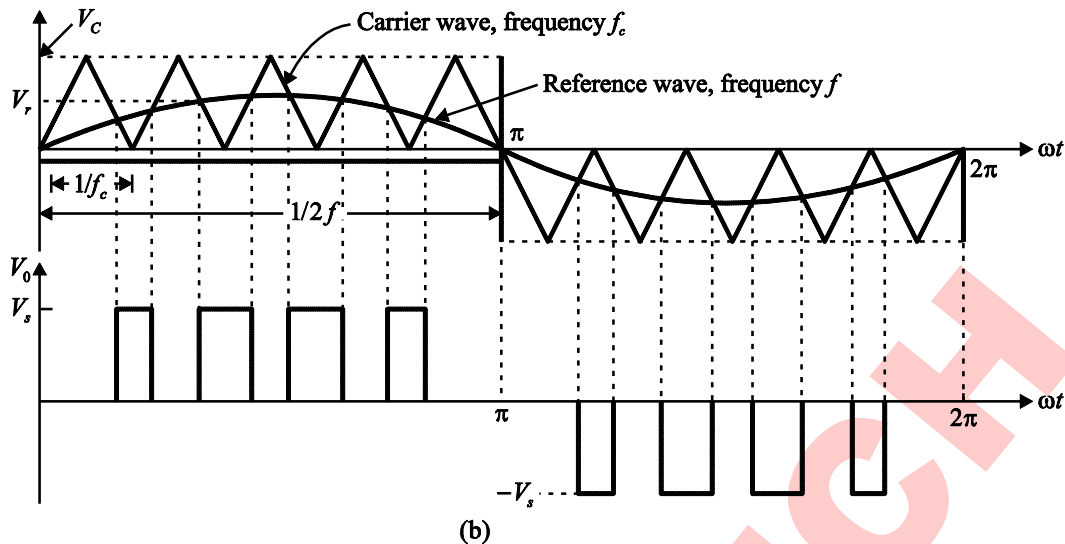
46.

Sol. PWM (pulse width modulation) is a internal control method of inverter. In this method, a fixed dc input voltage is given to the inverter and a controlled output voltage is obtained by, adjusting the on and off periods of the inverter components. In Sine modulation [Sin M] the pulse width is a sinusoidal function of the angular position of the pulse in a cycle as shown in figure.

For realizing sin M, a high-frequency triangular carrier wave V_c is compared with a sinusoidal reference wave V_r of the desired frequency. The intersection of V_c and V_r waves determines the switching instants and commutation of the modulated pulse. In Fig. V_c is the peak value of triangular carrier wave and V_r that of the reference, or modulating signal.

The carrier and reference waves are mixed in a comparator. When sinusoidal wave has magnitude higher than the triangular wave, the comparator output is high, otherwise it is low. The comparator output is processed in a trigger pulse generator in such a manner that the output voltage wave of the inverter has a pulse width in agreement with the comparator output pulse width.





Output voltage waveforms with sinusoidal pulse modulation

When triangular carrier wave has its peak coincident with zero of the reference sinusoid, there are

$N = f_c / 2f = \frac{f_c}{2f}$ pulses per half cycle ; Fig. (a) has five pulses. In case zero of the triangular wave coincides with zero of the reference sinusoid, there are $(N-1)$ pulses per half cycle; Fig. (b) has $(f_c/2f - 1)$, i.e. four, pulses per cycle.

The ratio of V_r / V_c is called the modulation index (MI) and it controls the harmonic content of the output voltage waveform. The magnitude of fundamental component of output voltage is proportional to MI , but MI can never be more than unity. Thus the output voltage is controlled by varying MI . Harmonic analysis of the output modulated voltage wave reveals that $\sin M$ has the following important features:

(i) For MI less than one, largest harmonic amplitudes in the output voltage are associated with harmonics of order $f_c/f \pm 1$ or $2N \pm 1$, where N is the number of pulses per half cycle. Thus, by increasing the number of pulses per half cycle, the order of dominant harmonic frequency can be raised, which can then be filtered out easily.

It is observed from above that as N is increased, the order of significant harmonic increases and the filtering requirements are accordingly minimised. But higher value of N entails higher switching frequency of thyristors. This amounts to more switching losses and therefore an impaired inverter efficiency. Thus a compromise between the filtering requirements and inverter efficiency should be made.

(ii) For MI greater than one, lower order harmonics appear, since for $MI > 1$, pulse width is no longer a sinusoidal function of the angular position of the pulse.

47.

Sol. For single pulse modulation, output voltage of single phase DC to AC inverter,

$$V_o = \sum_{n=1,3}^{\infty} \frac{4V_s}{n\pi} \sin \frac{n\pi}{2} \sin(nd) \sin(n\omega t)$$

$$\therefore nd = \pi \text{ and } \beta = 2d$$

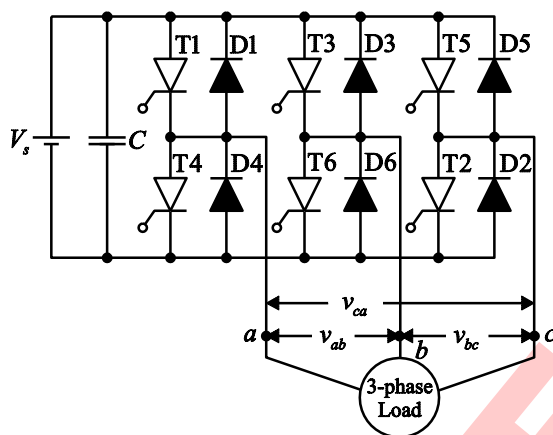
or if pulse width made equal to $\left(2d = \frac{2\pi}{n}\right)$ then n^{th} harmonic will be absent in the output.

Hence, option (b) is correct.



48.

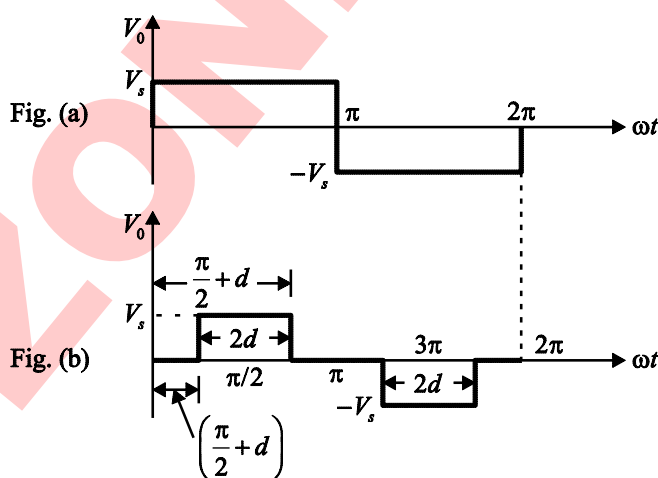
Sol. A basic three-phase inverter is a six-step bridge inverter it uses a minimum of six thyristors. In inverter terminology a step is defined as a change in the firing from one thyristor to the next thyristor in proper sequence. For one cycle of 360° , each step would be of 60° interval for a six-step inverter. This means that thyristors would be gated at regular intervals of 60° in proper sequence so that 3-phase ac voltage is synthesized at the output terminals of a six-step inverter.



Three-phase bridge inverter using thyristors

49.

Sol. The output voltage from single-phase full-bridge inverter is shown in figure (a). When this waveform is modulated, the output voltage is of the form shown in figure (b). It consists of a pulse of width $2d$ located symmetrically about $\pi/2$ and another pulse located symmetrically about $3\pi/2$. The range of pulse width $2d$ varies from 0 to π ; i.e. $0 < 2d < \pi$. The output voltage is controlled by varying the pulse-width $2d$.



Fourier analysis of figure (b) is as under

$$v_0 = \sum_{n=1,3,5}^{\infty} \frac{4V_s}{n\pi} \sin \frac{n\pi}{2} \sin nd \sin n\omega t$$

Or
$$v_0 = \frac{4V_s}{\pi} \left[\sin d \sin \omega t - \frac{1}{3} \sin 3d \sin 3\omega t + \frac{1}{5} \sin 5d \sin 5\omega t + \dots \right]$$

1. When pulse width $2d$ is equal to its maximum value of it radians, then the fundamental component of output voltage has a peak value of

$$v_{01m} = \frac{4V_s}{\pi}$$

2. If pulse width is equal to $2d = (2\pi/n)$, the n th harmonic is eliminated from the inverter output voltage.
3. The peak value of n th harmonic

$$v_{onm} = \frac{4V_s}{n\pi} \sin nd$$

4. The rms value of output voltage

$$V_{or} = \left[\frac{V_s^2 \cdot 2d}{\pi} \right]^{1/2} = V_s \left[\frac{2d}{\pi} \right]^{1/2}$$

PART - C

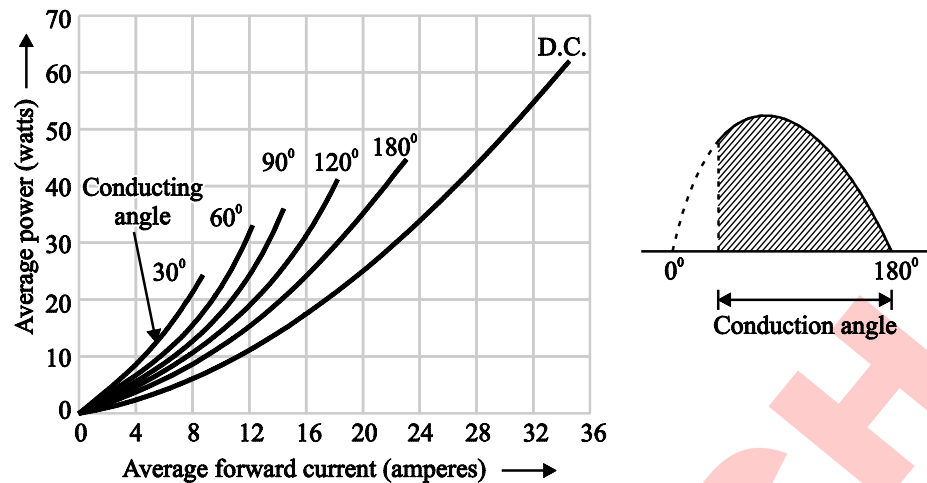
20 Marks Questions

50.

Sol. The power-generated in the junction region of a thyristor in a normal operation consists of the following components of dissipation.

1. Forward-conduction.
3. Turn-off or commutation.
2. Turn-on switching
4. Forward and Reverse blocking.
5. Gate pulse triggering.

1. **Forward Conduction Loss:** The average anode current multiplied by the forward voltage drop across the SCR is the average power dissipated in the thyristor. ON state conduction losses are the major source of junction heating for normal duty cycle and power-frequencies, figure shown below illustrates the variation of the ON state average conduction loss in watts with the average current in amperes for various conduction angles for operation on from 50 Hz to 400 Hz. This type of information is generally supplied by the manufacturer. The curve marked d.c. is applicable for continuous direct-current. These curves are based on the current waveform which is the remainder of the half-sine wave which results when delayed angle triggering is used in a single phase resistive load circuit. These power curves are integrated product of the instantaneous anode current and ON state voltage drop. This integration can be performed graphically or analytically for conduction angles other than those listed, using the ON state voltage-current characteristic curves for the specific device. In the line commutated converters) and a.c. regulator circuits, the forward conduction loss is the major source of junction heating.



Average forward power dissipation for sinusoidal current waveform

2. **Turn-on Losses:** Since the switching process takes a finite time, there is a relatively high voltage across the thyristor while a current flows. Therefore, this loss is rather higher than the turn-off loss. For example, by the time the current has reached 90 per cent of its final value, there may still be 10 per cent of the supply voltage across the device. Accordingly, appreciable power may be dissipated during this turn-on interval. Above 400 Hz switching, additional circuitry is used to reduce the switching losses or else some derating of the normal forward current is made to allow for the extra dissipation.
3. **Turn-off Losses:** The turn-off power losses arises during the time of decay of reverse current, according to the product of the instantaneous values of reverse current and reverse voltage, may reach high peak values up to several kilowatts. It is possible during rapid turn-off for the reverse current to rise to a value comparable to the forward current. When the thyristor impedance starts to increase, dissipation occurs as the current falls and the reverse voltage builds up. To limit the rate of change of current at turn-off and hence the energy to be dissipated, circuit inductance is used. This also limits the rate of rise of forward current which is an advantage but the inductance can give rise to high reverse voltage transients during turn-off. In high frequency inverters, where the thyristors are switched ON and OFF several times in each supply cycle, the turn-on and turn-off losses may also have to be taken into consideration while selecting the device ratings since the switching loss may constitute a significant portion of the total loss.
4. **Forward and Reverse Blocking Losses:** As mentioned in the previous section, a thyristor has different regions of operation. In the forward blocking region, anode is made positive with respect to cathode and the anode current is the small forward leakage current. Therefore, the forward blocking power loss is the integration of product of the forward blocking voltage and forward leakage current. Similarly, reverse power loss occurs in reverse blocking region. The forward power loss is generally small compared to the conduction-loss.
5. **Gate Power Loss:** The gate power loss is the mean power loss due to gate current between the gate and main terminals. Gate losses are negligible for pulse types of triggering signals. Losses may become more significant for gate signals with a high duty cycle.

51.

Sol. Latching current: The minimum value of anode current which the thyristor must attain during turn-on process to maintain conduction even after the removal of gate pulse is called latching current.

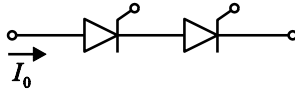
(i) **Holding current:** It is defined as the minimum value of anode current below which it must fall for turning off the thyristor.

In general latching current > holding current.

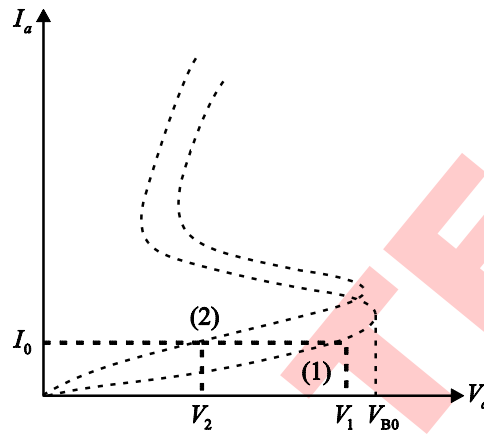
String efficiency is a term that is used for measuring the degree of utilization of SCR in a string.

- (ii) String efficiency = $\frac{\text{Actual voltage (series) or current (parallel) rating of the whole string}}{\text{No. of SCR's in the string} \times \text{voltage or current rating of each SCR}}$

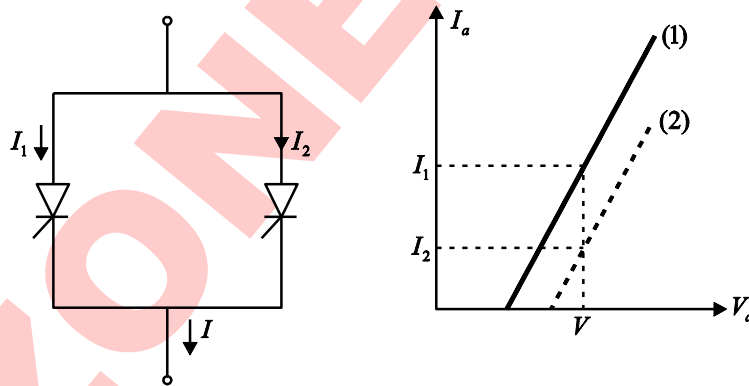
Series operation of SCR's



$$\eta_s = \frac{V_1 + V_2}{2V_1}$$

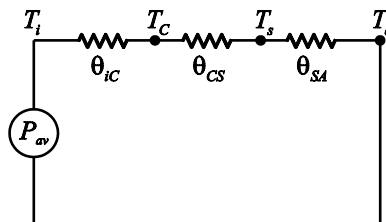


Parallel operation of SCR's



$$\text{String efficiency} = \frac{I_1 + I_2}{2I_1}$$

- (iii)



From the equivalent circuit,

$$T_j = T_s + P_{av}(\theta_{jC} + \theta_{CS})$$

$$\therefore P_{av1} = \frac{125 - 70}{0.16 + 0.08} = 229.17 \text{ W}$$

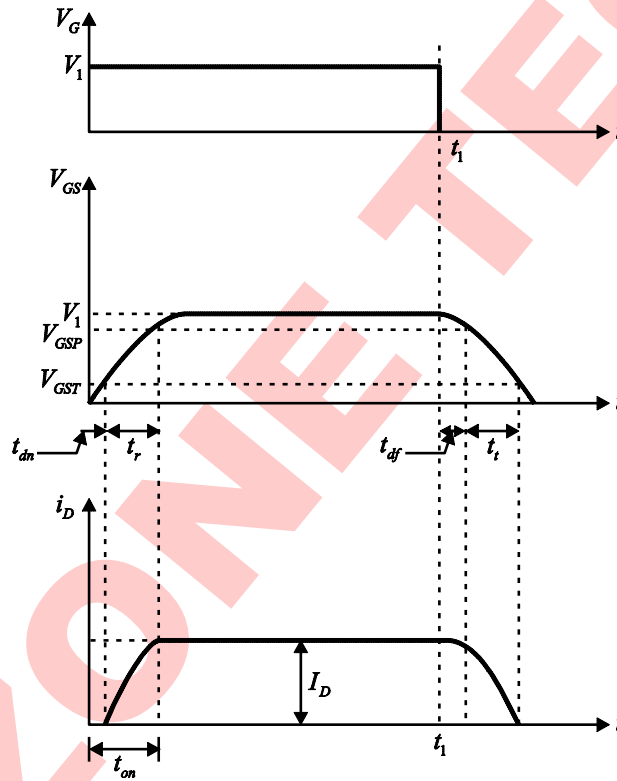
Thus total power loss in the thyristor-sink combination is 229.17 W with improved cooling

$$P_{av2} = \frac{125 - 60}{0.24} = 270.83 \text{ W}$$

Thyristor rating is proportional to the square root of average power loss.

$$\% \text{ increase in thyristor rating} = \frac{\sqrt{270.83} - \sqrt{229.17}}{\sqrt{229.17}} \times 100 = 8.71$$

52.
Sol.

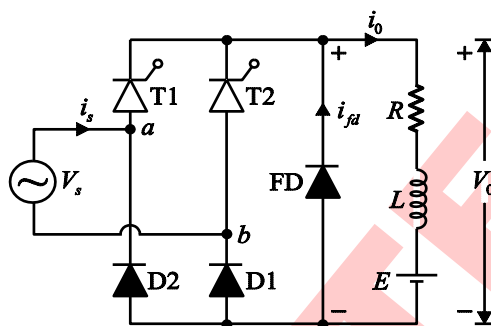


1. The switching characteristics of a power MOSFET are influenced to a large extent by the internal capacitance of the device and the internal impedance of the gate drive circuit. At turn-on, there is an initial delay t_{dn} during which input capacitance charges to gate threshold voltage V_{GST} . Here t_{dn} is called turn-on delay time.
2. There is further delay t_r , called rise time, during which gate voltage rises to V_{GSP} , a voltage sufficient to drive the MOSFET into on state. During t_r , drain current rises from zero to full-on current I_D . Thus, the total turn on time is $t_{on} = t_{dn} + t_r$. The turn-on time can be reduced by using low-impedance gate-drive source.

3. As MOSFET is a majority carrier device, turn-off process is initiated soon after removal of gate voltage at time t_1 . Time turn-off delay time t_{df} , is the time during which input capacitance discharges from overdrive gate voltage V_1 to V_{GSP} . The full time t_f , is the time during which, input capacitance discharges from V_{GSP} to threshold voltage. During t_f , drain current falls from I_D to zero. So when $V_{GS} \leq V_{GST}$, PMOSFET turn-off is complete.

53.

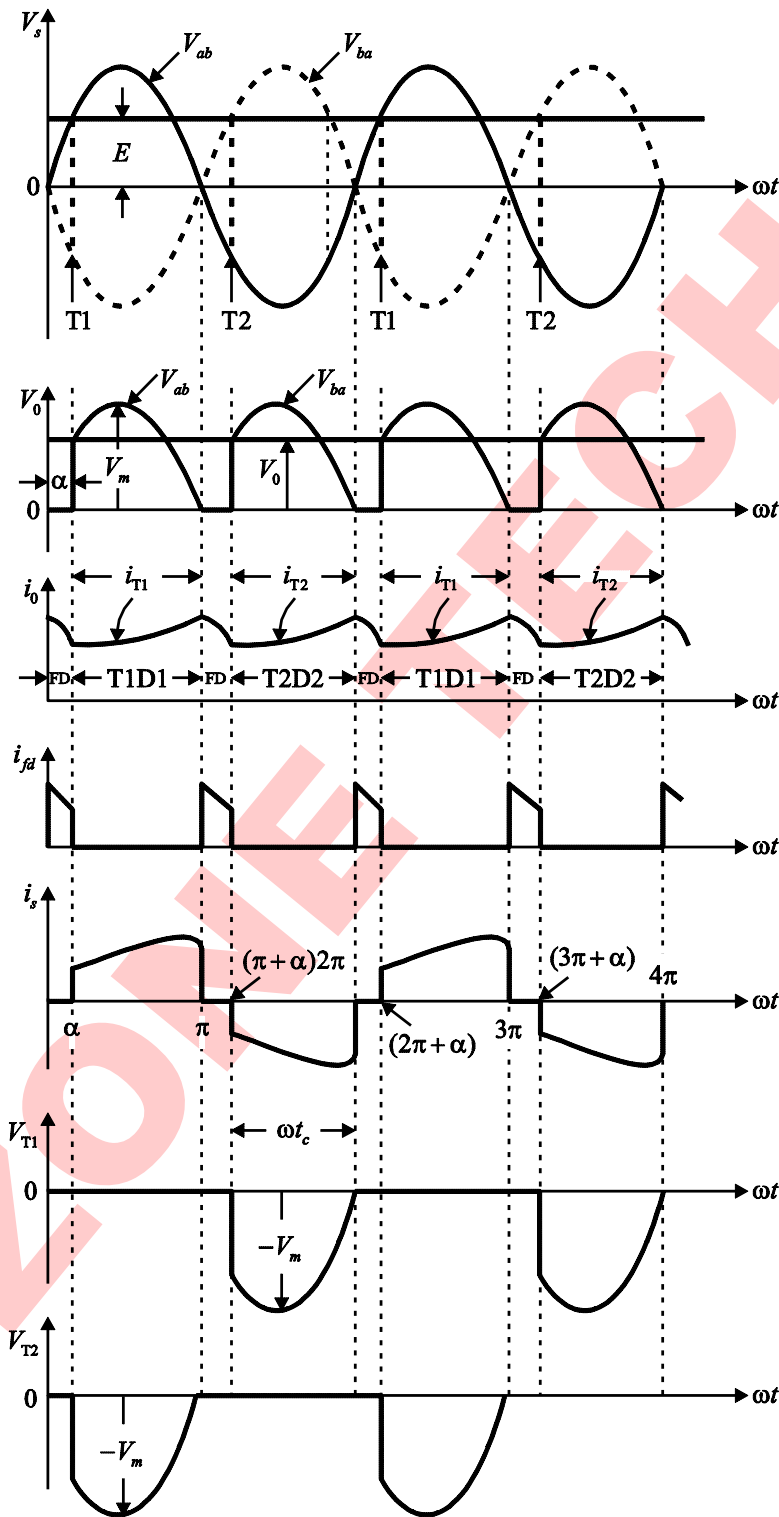
- Sol.** 1. A single phase semiconverter bridge with two thyristors and three diodes is shown in figure. The two thyristors are T1, T2; the two diodes are D1, D2; the third diode connected across load is freewheeling diode FD. The load is of RLE type as for the full converter bridge. Various voltage and current waveforms for this converter are shown in (b), where load current is assumed continuous over the working range.



(a) Single-phase semiconverter

2. After $\omega t = 0$, thyristor T1 is forward biased only when source voltage $V_m \sin \omega t$ exceeds E . Thus, T1 is triggered at a firing angle delay α such that $V_m \sin \alpha > E$. With T1 on, load gets connected to source through T1 and D1 for the period $\omega t = \alpha$ to π , load current i_0 flows through RLE, D1, source and T1 and the load terminal voltage v_0 is of the same waveshape as the ac source voltage v_s . Soon after $\omega t = \pi$, load voltage v_0 tends to reverse as the ac source voltage changes polarity. The load or output, current i_0 is transferred from T1, D1 to FD. As SCR T1 is reverse biased at $\omega t = \pi +$ through FD, T1 is turned off at $\omega t = \pi +$. The load terminals are short circuited through FD, therefore load, or output, voltage v_0 is zero during $\pi < \omega t < (\pi + \alpha)$. After $\omega t = \pi$, during the negative half cycle; T2 will be forward biased only when source voltage is more than E . At $\omega t = (\pi + \alpha)$, source voltage exceeds E , T2 is therefore triggered. Soon after $(\pi + \alpha)$, FD is reverse biased and is therefore turned off; load current now shifts from FD to T2, D2. At $\omega t = 2\pi$, FD is again forward biased and output current i_0 is transferred from T2, D2 to FD as explained before. The source current i_s is positive from α to π when T1, D1 conduct and is negative from $(\pi + \alpha)$ to 2π when T2, D2 conduct, see figure.
3. During the interval α to π , T1 and D1 conduct and ac source delivers energy to the load circuit. This energy is partially stored in inductance L , partially stored as electric energy in load-circuit emf E and partially dissipated at heat in R . During the freewheeling period π to $(\pi + \alpha)$, energy stored in inductance is recovered and is partially dissipated in R and partially added to the energy stored in load emf E . No energy is fed back to the source during freewheeling period. For semiconverter, the average output voltage V_0 , from figure (b), is given by

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



(b) Voltage and current waveforms for continuous load current

54.

- Sol.** A d.c. chopper is a static device that converts fixed d.c. input voltage to a variable d.c. output voltage directly. Actually a step down chopper is a high speed on/off semiconductor switch. It connects source to load and disconnects the load from source at a fast speed. In this manner, a chopped load voltage as shown in Fig. (b) is obtained from a constant dc supply of magnitude V_s . In Fig. (a) chopper is represented by a switch SW inside a dotted rectangle, which may be turned-on or turned-off as desired.

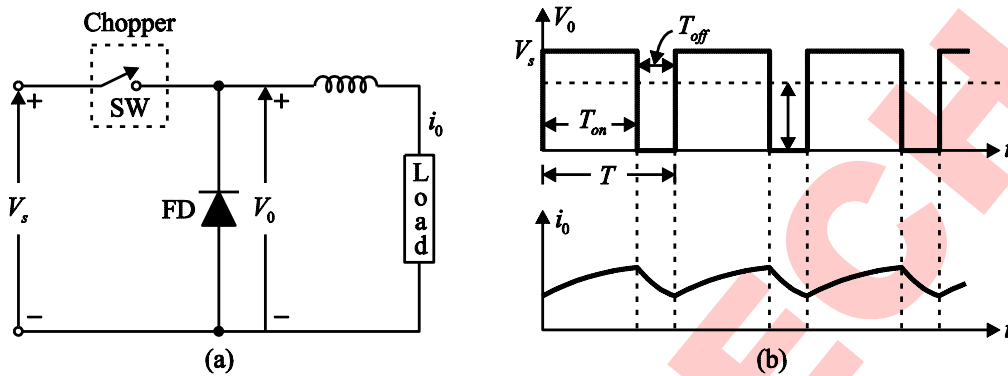


Fig. (a) Elementary chopper circuit and (b) output voltage current waveforms

During the period T_{on} , chopper is on and load voltage is equal to source voltage V_s . During the interval T_{off} , chopper is off, load current flows through the freewheeling diode FD. As a result, load terminals are short circuited by FD and load voltage is therefore zero during T_{off} . In this manner a chopped dc voltage is produced at the load terminals. The load current as shown in Fig. (b) is continuous. From Fig. (b), average load voltage V_0 is given by

$$V_0 = \frac{T_{on}}{T_{on} + T_{off}} V_s = \frac{T_{on}}{T} V_s \quad \alpha = V_s \quad \dots\dots (i)$$

Where T_{on} = on-time; T_{off} = off-time

$$T = T_{on} + T_{off} = \text{chopping period}$$

$$\alpha = \frac{T_{on}}{T} = \text{duty cycle}$$

Thus load voltage can be controlled by varying duty cycle α . Equation (i) shows that load voltage is independent of load current. Equation (i) can also be written as

$$V_0 = f \times T_{on} \cdot V_s$$

where, $f = \frac{1}{T}$ = chopping frequency

Step-up Choppers :

For the chopper configuration of Fig. (a). Average output voltage V_0 greater than input voltage V_s can be obtained by a chopper called step-up chopper. Fig. illustrates an elementary form of a step-up chopper.

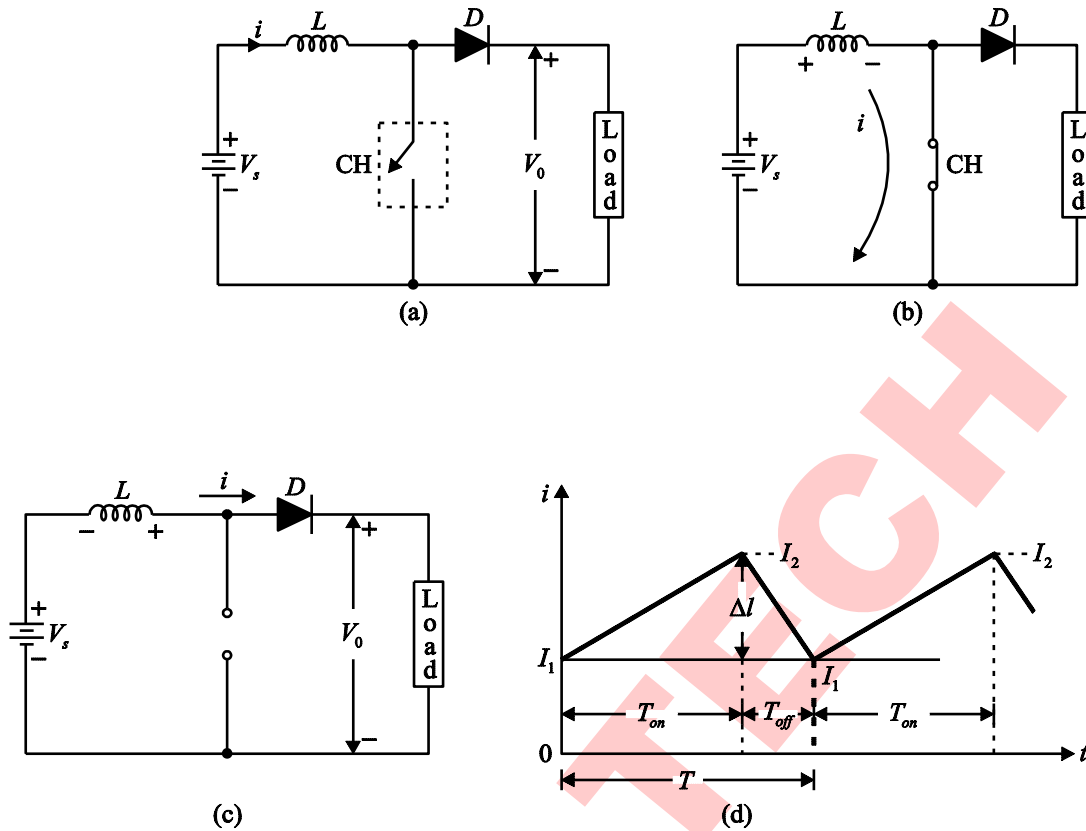


Fig. (a) Step-up chopper (b) L stores energy (c) L. di/dt is added to V_s (d) current waveform

In this chopper, a large inductor L in series with source voltage V_s is essential as shown in Fig. (a). When the chopper CH is on, the closed current path is as shown in Fig. (b) and inductor stores energy during T_{on} period. When the chopper CH is off, as the inductor current cannot die down instantaneously, this current is forced to flow through the diode and load for a time T_{off} , Fig. (c). As a result, voltage across the load, given by $V_0 = V_s + L(di/dt)$, exceeds the source voltage V_s . In this manner, the circuit of Fig.(a) acts as a step-up chopper and the energy stored in L is released to the load.

When CH is on, current through the load would increase from I_1 to I_2 as shown in Fig. (d). When CH is off, current would fall from I_2 to I_1 . With CH on, source voltage is applied to L i.e. $v_L = V_s$. When CH is off, KCL for Fig. (c) gives $v_L - V_0 + V_s$, or $v_L = V_0 - V_s$. Here v_L = voltage across L . Assuming linear variation of output current, the energy input to inductor from the source, during the period T_{on} , is

$$\begin{aligned} W_{in} &= (\text{voltage across } L) (\text{average current through } L) T_{on} \\ &= V_s \cdot \left(\frac{I_1 + I_2}{2} \right) T_{on} \end{aligned} \quad \text{..... (i)}$$

During the time T_{off} , when chopper is off, the energy released by inductor to the load is

$$\begin{aligned} W_{off} &= (\text{voltage across } L) (\text{average current through } L) T_{off} \\ &= (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) T_{off} \end{aligned} \quad \text{..... (ii)}$$

Considering the system to be lossless, these two energies given by Eq.. (i) and (ii) will be equal

$$\therefore V_s \left(\frac{I_1 + I_2}{2} \right) T_{on} = (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) \cdot T_{off}$$

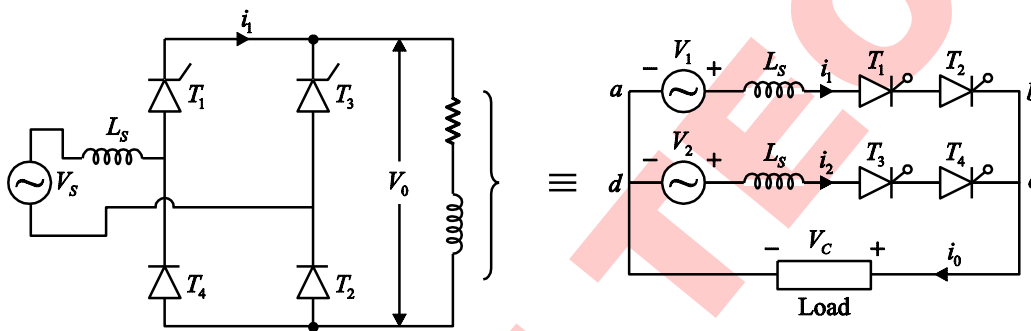
$$V_0 = V_s \frac{T}{T - T_{on}} = V_s \frac{1}{1 - \alpha}$$

Output voltage depends upon α that can be controlled by varying frequency There are two schemes:

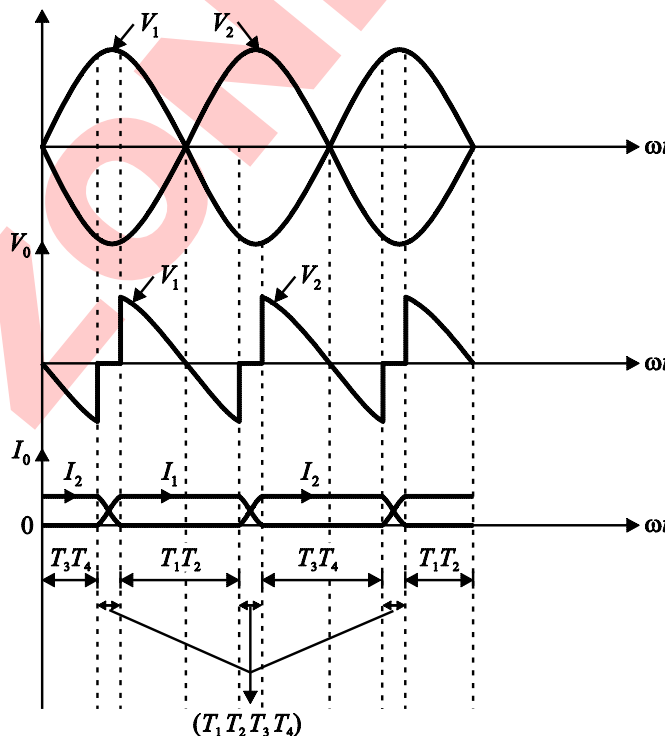
- (i) Constant frequency and
- (ii) Variable frequency for having such control.

Constant frequency scheme is better than variable frequency scheme. In constant frequency scheme, T_{on} , cannot be reduced to zero for most of the commutation circuits used in choppers. As such, low range of α control is not possible in this scheme.

55.
Sol.



1 ϕ full converter with source inductance L.



From the equivalent circuit, applying KVL in loop 'abcd' gives

$$V_1 - L_s \frac{di_1}{dt} = V_2 - L_s \frac{di_2}{dt}$$

$$V_1 - V_2 = L_s \left(\frac{di_1}{dt} - \frac{di_2}{dt} \right)$$

If $V_1 = V_m \sin \omega t$ then,

$$V_2 = -V_m \sin \omega t$$

$$\therefore L_s \left(\frac{di_1}{dt} - \frac{di_2}{dt} \right) = 2V_m \sin \omega t$$

As the load assumed constant throughout, $i_1 + i_2 = I_0$

$$\frac{di_1}{dt} + \frac{di_2}{dt} = 0$$

$$\frac{di_1}{dt} - \frac{di_2}{dt} = \frac{2V_m}{L_s} \sin \omega t$$

Addition of equation gives,

$$\frac{di_1}{dt} = \frac{V_m}{L_s} \sin \omega t$$

Load current ' i_1 ' through thyristor pair T_1, T_2 build up from zero to (I_0) during the overlap angle μ at $(\omega t = \alpha)$

At $\omega t = \alpha$, $i_1 = 0$ and at $\omega t = (\alpha + \mu)$, $i_1 = I_0$

$$\int_0^{I_0} i_1 = \frac{V_m}{L_s} \int_{\alpha/\omega}^{(\alpha+\mu)/\omega} \sin \omega t \cdot dt \Rightarrow I_0 = \frac{-V_m}{\omega L_s} [\cos \alpha - \cos(\alpha + \mu)] \quad \dots (i)$$

And average voltage (output),

$$\begin{aligned} V_0 &= \frac{V_m}{\pi} \int_{\alpha+\mu}^{(\alpha+\pi)} \sin \omega t d(\omega t) \\ &= \frac{V_m}{\pi} [\cos(\alpha + \mu) + \cos \alpha] \quad \dots (ii) \end{aligned}$$

$$\therefore V_0 = \frac{2V_m}{\pi} \cos \alpha - \frac{\omega L_s}{\pi} I_0$$

$$\text{Or } V_0 = \frac{2V_m}{\pi} \cos(\alpha + \mu) + \frac{\omega L_s}{\pi} I_0$$

Here $V_m = 330V$, $\omega = 314$, $\alpha = \frac{\pi}{4}$

$$I_0 = 5A, V_{0(av)} = 140V$$

$$(i) \quad V_0 = \frac{1}{\pi} \left[2 \times 330 \times \cos\left(\frac{\pi}{4}\right) - 314 \times L_s \times 5 \right] = 140 \quad \therefore L_s = 17.113 \text{ mH}$$

$$(ii) \quad I_0 = \frac{V_m}{\omega L_s} [\cos \alpha - \cos(\alpha + \mu)]$$

$$5 = \frac{330}{314 \times 0.017113} \left[\cos \frac{\pi}{4} - \cos \left(\frac{\pi}{4} + \mu \right) \right]$$

$$\therefore \mu = 6.267^\circ$$

(iii) The average voltage across the output resistance.

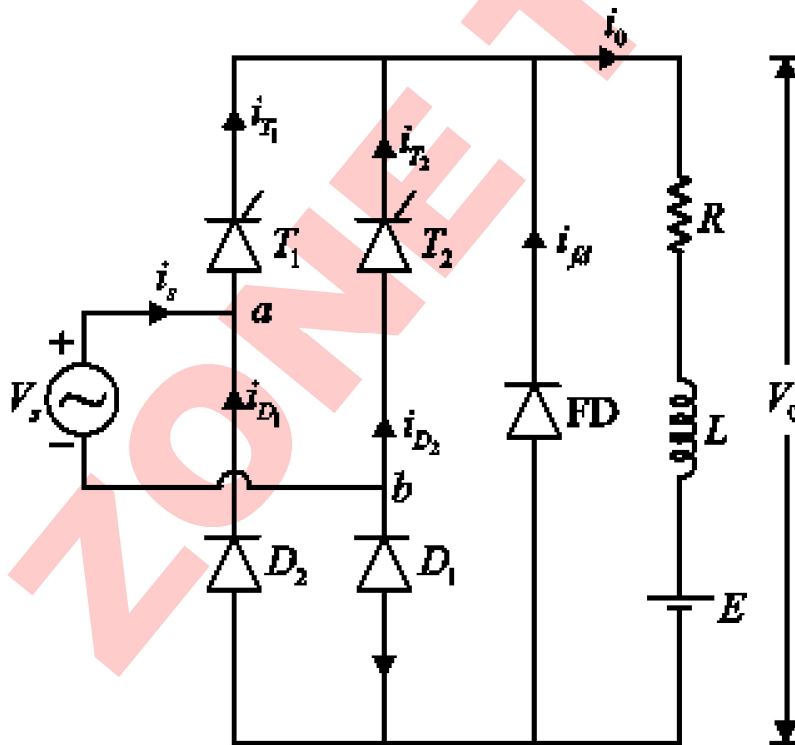
$$V_0 = I_0 R$$

$$\Rightarrow R = \frac{V_0}{I_0} = \frac{140}{5} = 28 \Omega$$

56.

Sol. Regions of Converter Operation :

1. $0 < \omega t < \alpha$, FD conducts.
2. $\alpha < \omega t < \pi$, $T_1 D_1$ conducts.
3. $\pi < \omega t < \pi + \alpha$, FD conducts.
4. $\pi + \alpha < \omega t < 2\pi$, $T_2 D_2$ conducts.
5. When $T_1 D_1$ and $T_2 D_2$ conducts output voltage follows input voltage.



6. For semiconverter, the average output voltage,

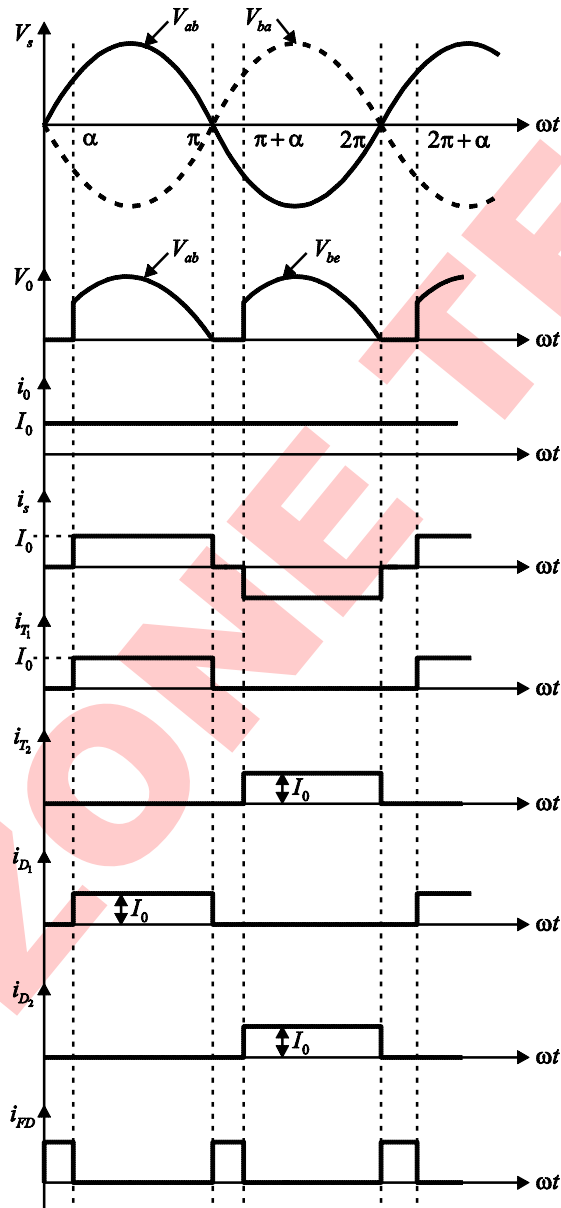
$$V_0 = \frac{1}{\pi} \int_{\alpha}^{\pi} V_m \sin \omega t \cdot d(\omega t) = \frac{V_m}{\pi} (1 + \cos \alpha)$$

7. Rms value of output voltage,

$$V_{or}^2 = \frac{1}{\pi} \int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \cdot d(\omega t) = \frac{V_m^2}{2\pi} \left[\omega t - \frac{\sin 2\omega t}{2} \right]_{\alpha}^{\pi}$$

$$= \frac{V_s^2}{\pi} \left[(\pi - \alpha) + \frac{\sin 2\alpha}{2} \right]$$

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



57.
Sol.

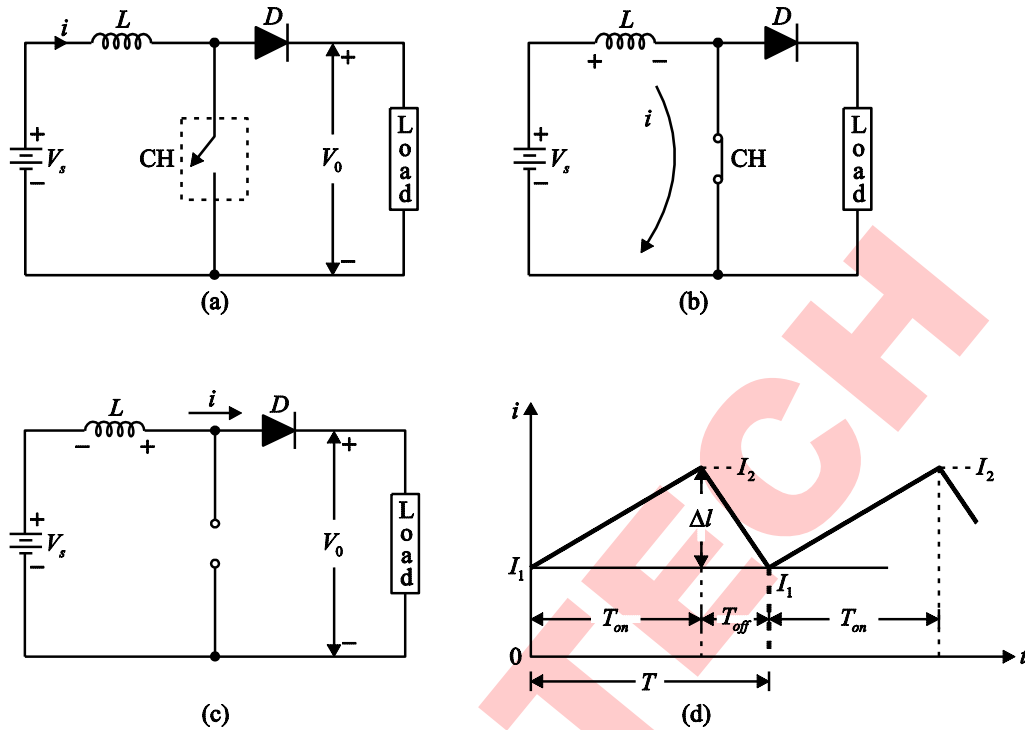


Fig. (a) Step-up chopper (b) L stores energy (c) L . di/dt is added to V_s
(d) current waveform

In this chopper, a large inductor L in series with source voltage V_s is essential as shown in Fig. (a). When the chopper CH is on, the closed current path is as shown in Fig. (b) and inductor stores energy during T_{on} period. When the chopper CH is off, as the inductor current cannot die down instantaneously, this current is forced to flow through the diode and load for a time T_{off} , Fig. (c). As a result, voltage across the load, given by $V_0 = V_s + L(di/dt)$, exceeds the source voltage V_s . In this manner, the circuit of Fig. (a) acts as a step-up chopper and the energy stored in L is released to the load.

When CH is on, current through the load would increase from I_1 to I_2 as shown in Fig. (d). When CH is off, current would fall from I_2 to I_1 . With CH on, source voltage is applied to L i.e. $v_L = V_s$. When CH is off, KCL for Fig. (c) gives $v_L - V_0 + V_s = 0$, or $v_L = V_0 - V_s$. Here v_L = voltage across L . Assuming linear variation of output current, the energy input to inductor from the source, during the period T_{on} , is

$$\begin{aligned} W_{in} &= (\text{voltage across } L) (\text{average current through } L) T_{on} \\ &= V_s \cdot \left(\frac{I_1 + I_2}{2} \right) T_{on} \end{aligned} \quad \dots (i)$$

During the time T_{off} , when chopper is off, the energy released by inductor to the load is

$$\begin{aligned} W_{off} &= (\text{voltage across } L) (\text{average current through } L) T_{off} \\ &= (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) T_{off} \end{aligned} \quad \dots (ii)$$

Considering the system to be lossless, these two energies given by Eqs. (i) and (ii) will be equal

$$\therefore V_s \left(\frac{I_1 + I_2}{2} \right) T_{on} = (V_0 - V_s) \left(\frac{I_1 + I_2}{2} \right) \cdot T_{off}$$

$$V_0 = V_s \frac{T}{T - T_{on}} = V_s \frac{1}{1 - \alpha} \quad \dots (iii)$$

It is seen from equation (iii) that average voltage across the load can be stepped up by varying the duty cycle. If chopper of Fig (a) is always off, $\alpha = 0$ and $V_0 = V_s$. If this chopper is always on, $\alpha = 1$ and $V_0 = \infty$ (infinity). In practice, chopper is turned on and off so that α is variable and the required step-up average output voltage, more than source voltage, is obtained.

(ii) Using, $V_0 = V_s \cdot \frac{1}{1 - \alpha}$

$$660 = 220 \frac{1}{1 - \alpha}$$

Or $\alpha = \frac{2}{3} = \frac{T_{on}}{T}$

$$\therefore T_{on} = \frac{2}{3}T \text{ and } T_{off} = T - T_{on} = \frac{1}{3}T$$

$$\frac{1}{3}T = 100 \mu\text{s (given)}$$

$$\therefore T = 300 \mu\text{s and } T_{on} = \frac{2}{3} \times 300 = 200 \mu\text{s (pulse width)}$$

When pulse width is halved, $T_{on} = \frac{1}{2} \times 200 = 100 \mu\text{s}$

For constant frequency operation,

$$T = 300 \mu\text{s}; T_{off} = T - T_{on} = 200 \mu\text{s} \quad \alpha = \frac{T_{on}}{T} = \frac{100}{300} = \frac{1}{3}$$

$$\therefore \text{New output, } V_0 = 220 \frac{1}{1 - \frac{1}{3}} = 330 \text{ V}$$

58.

Sol. A basic three-phase inverter is a six-step bridge inverter it uses a minimum of six thyristors.

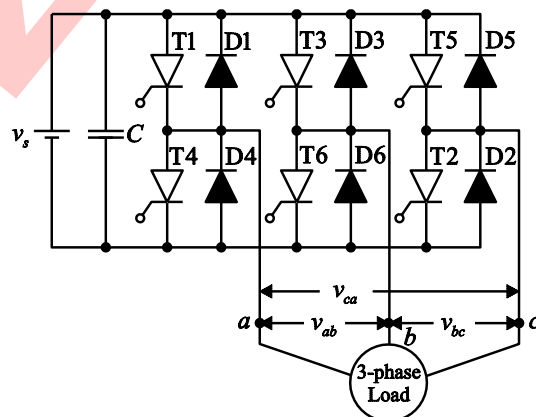


Fig. 1 : Three-phase bridge inverter using thyristors

In inverter terminology a step is defined as a change in the firing from one thyristor to the next thyristor in proper sequence. For one cycle of 360° , each step would be of 60° interval for a six-step inverter. This means that thyristors would be gated at regular intervals of 60° in proper sequence so that 3-phase ac voltage is synthesized at the output terminals of a six-step inverter.

In the three-phase inverter of figure each SCR conducts for 180° of a cycle. Thyristor pair in each arm i.e. T1, T4 ; T3, T6 and T5, T2 are turned on with a time interval of 180° .

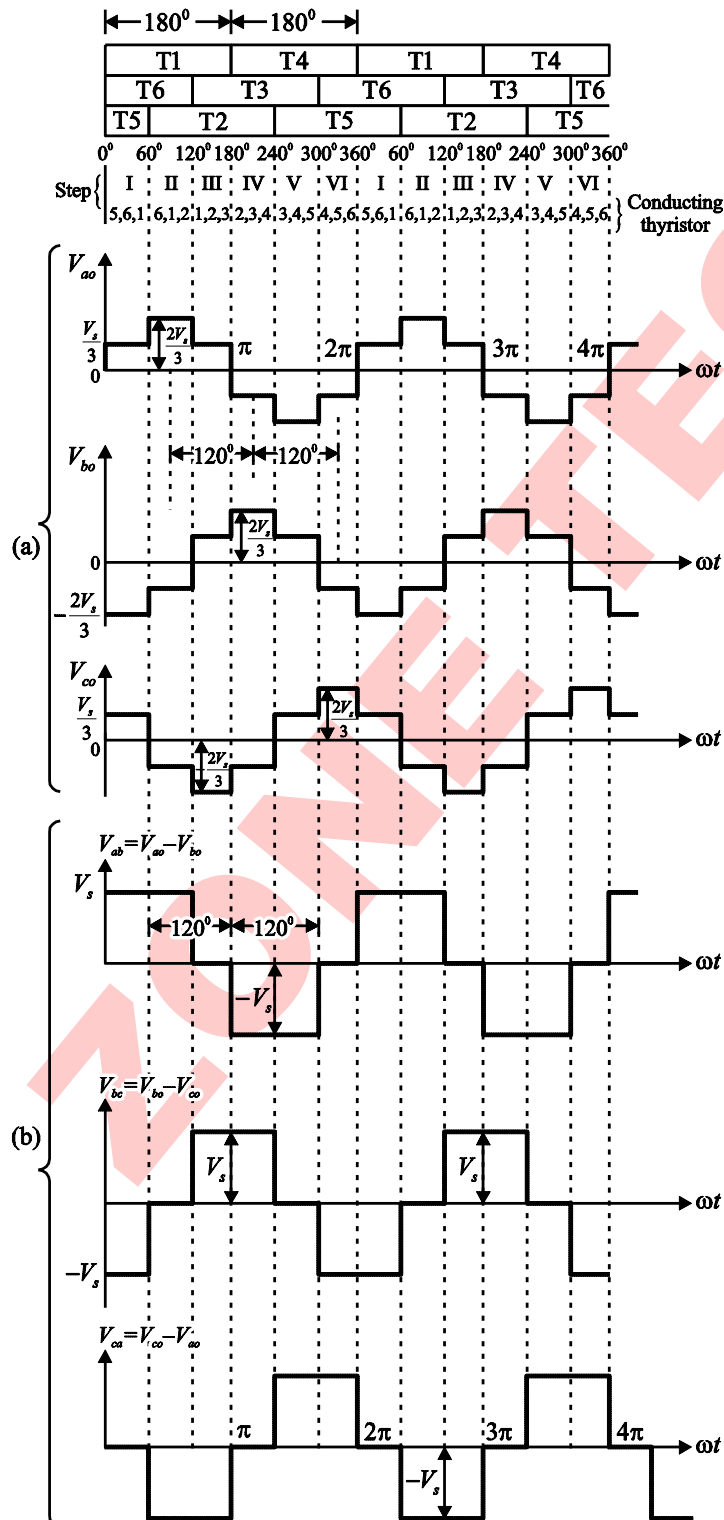
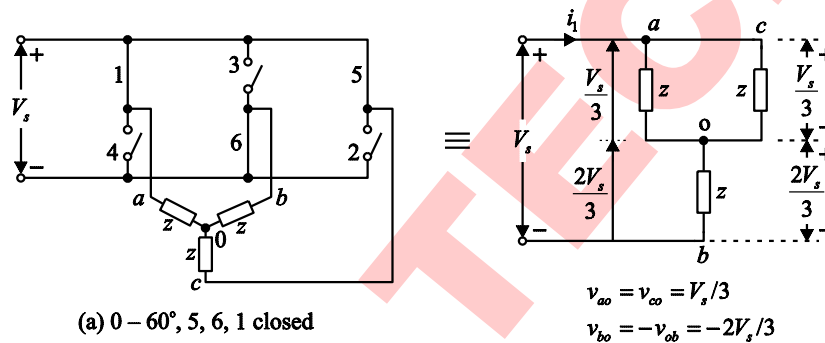


Fig. 2 : Voltage waveforms for 180° mode 3-phase VSI

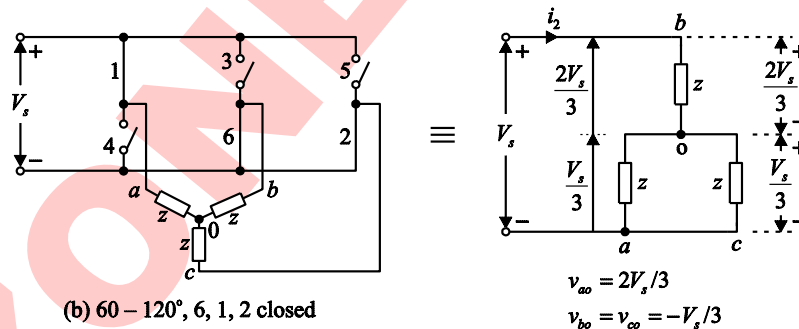
It means that T1 conducts for 180° and T4 for the next 180° of a cycle. Thyristors in the upper group, i.e. T1, T3, T5 conduct at an interval of 120°. It implies that if T1 is fired at $\omega t = 0^\circ$, then T3 must be fired at $\omega t = 120^\circ$ and T5 at $\omega t = 240^\circ$. Same is true for lower group of SCRs. On the basis of this firing scheme, a table is prepared as shown at the top of Fig. In this table, first row shows that T1 from upper group conducts for 180°, T4 for the next 180° and then again T1 for 180° and so on. In the second row, T3 from the upper group is shown to start conducting 120° after T1 starts conducting. After T3 conduction for 180°, T6 conducts for the next 180° and again T3 for the next 180° and so on. Further in the third row, T5 from the upper group starts conducting 120° after T3 or 240° after T1. After T5 conduction for 180°, T2 conducts for the next 180°, T5 for the next 180° and so on.

These output voltages are plotted in Fig. 2. In this manner, the variation of phase voltages v_{a0} , v_{c0} as obtained in Fig. 3 up to step IV and similarly for other steps, is plotted in Fig. 2. It is clear that for each cycle of output voltage of each phase, six steps are required and each step has a duration of 60°.

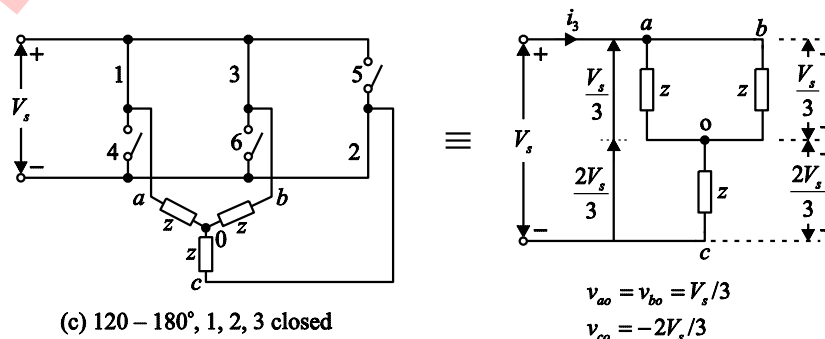
Step - I :



Step - II :



Step - III :



Step - IV :

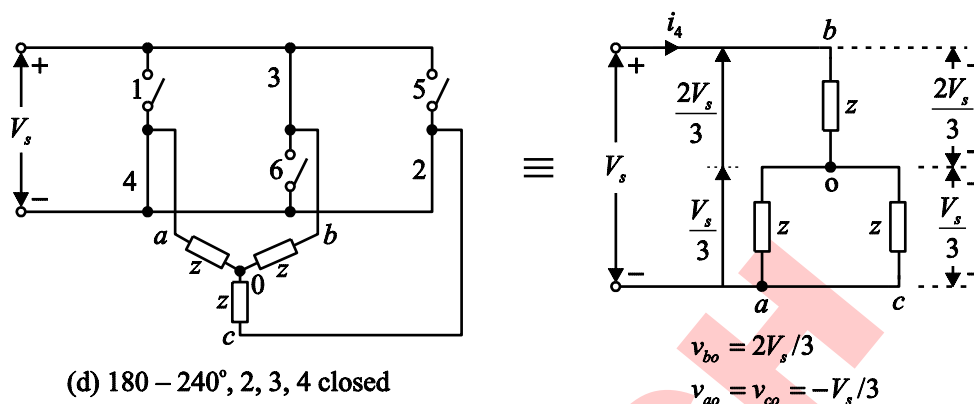
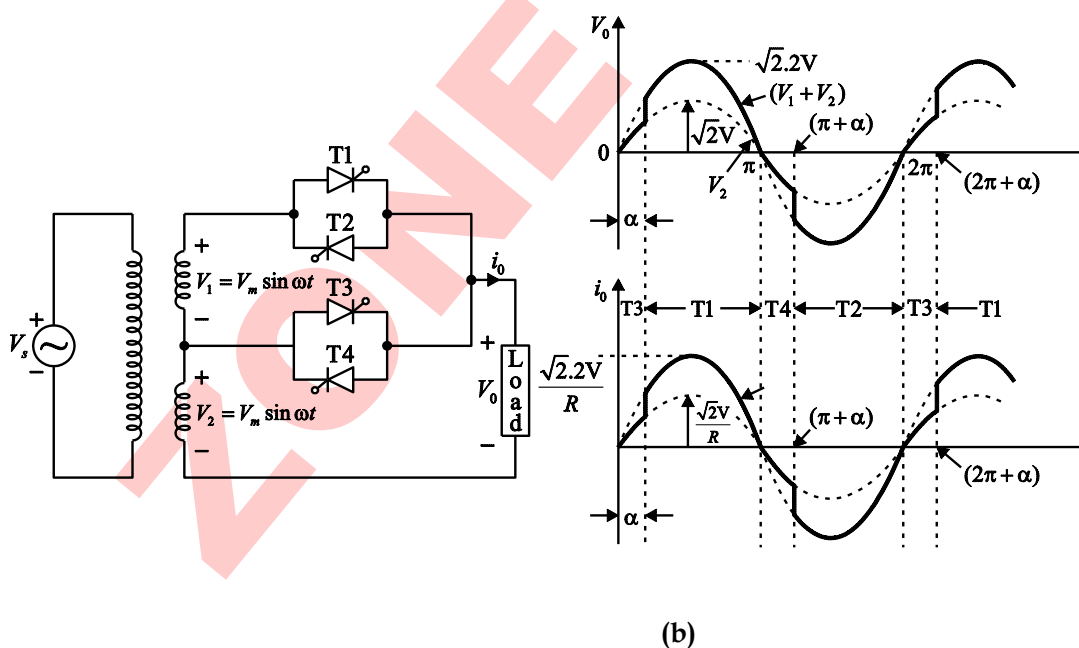


Fig. 3 : Equivalent circuits for a 3-phase six-step 180° mode inverter with a balanced star-connected load

59.

Sol. A two-stage sequence control of ac voltage regulators employs two stages in parallel as shown in Fig. (a). The turns ratio from primary to each secondary is taken as unity. This means that for source voltage $V_s = V_m \sin \omega t$, $V_1 = V_2 = V_m \sin \omega t$ and sum of two secondary voltages is $2V_m \sin \omega t$.

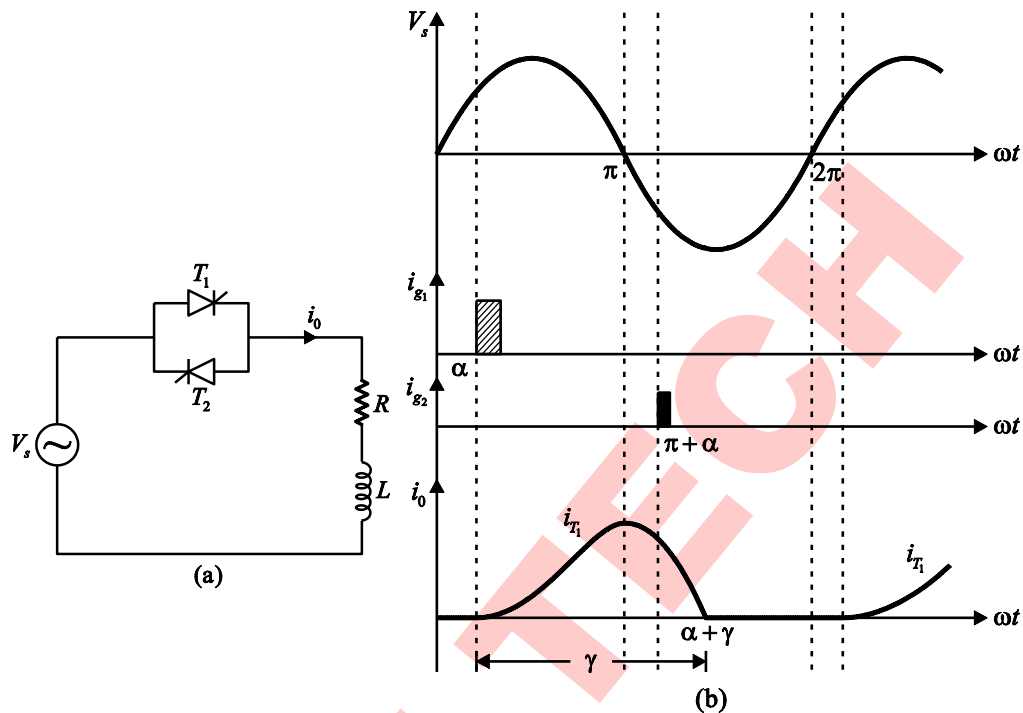


The load is R for obtaining output voltage control from zero to rms value V , use only thyristor pair T_3, T_4 in Fig. For zero output voltage α is 180° for T_3, T_4 and for V , α is zero. For output voltage control from V to $2V$, α for thyristor pair T_3, T_4 is always zero and for thyristor pair T_1, T_2 ; α is varied from zero to 180° .

The main advantage of two-stage sequence control of ac voltage controller over single-phase full-wave ac voltage controller is the reduction of harmonics in the load and line currents. For resistance load, the load current waveform is identical with output voltage waveform.

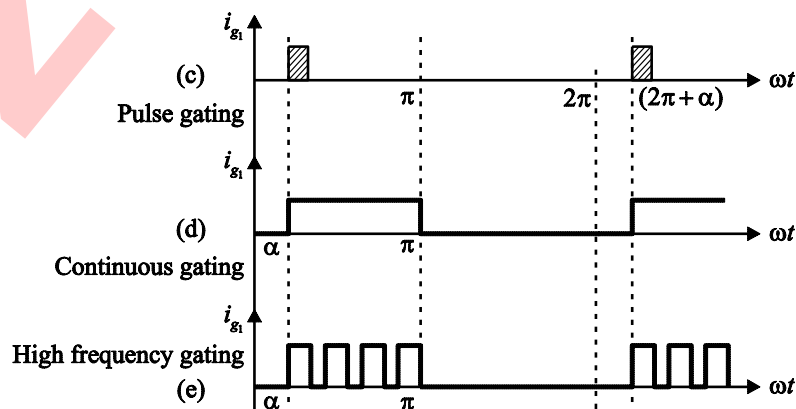
60.

Sol.

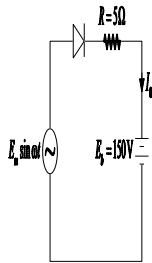


1. Pulse gating is not suitable for RL loads. The reason for this can be explained by referring to figure (b). At α , T_1 is fixed and the current grows as shown. At $\pi + \alpha$, T_1 is fixed. At $\pi + \alpha$, T_2 is fired. As T_1 is still conducting, voltage drop across T_1 reverse biases T_2 at $\pi + \alpha$, T_2 therefore not turned on.

At $(\alpha + \gamma)$, i_{T1} decays to zero and T_1 stops conducting, as a result T_2 gets forward biased but the gate pulse i_{g2} applied to T_2 at $\pi + \alpha$ is already zero and therefore T_2 does not get turned on.



2. This difficulty can be overcome by applying continuous gate signal to the SCRs T_1 and T_2 so that when i_{T_1} becomes zero at $(\alpha + \gamma)$, T_2 get turned on due to the presence of continuous gate signal.

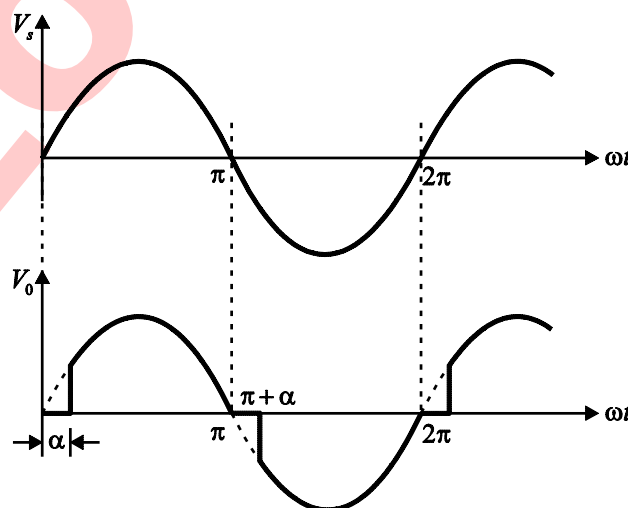
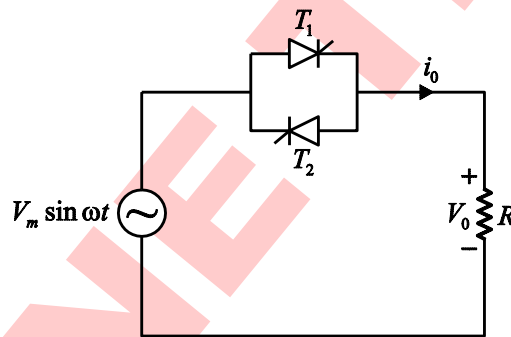


3. In practice continuous gating is undesirable as it leads to more heating of the SCR gate and at the same time, it increases the size of the pulse transformer. High frequency carrier gating mitigates the disadvantages of continuous gate signal and ensures thyristor turn on.

61.

Sol. (i) Phase control method:

For lighting load, phase control method is used.



During positive half cycle, T_1 is triggered at firing angle α , it conducts from $\omega t = \alpha$ to π for the load. During negative half cycle T_2 is triggered at $\omega t = \pi + \alpha$, it conducts from $\omega t = \pi + \alpha$ to 2π .

Rms value of output voltage,

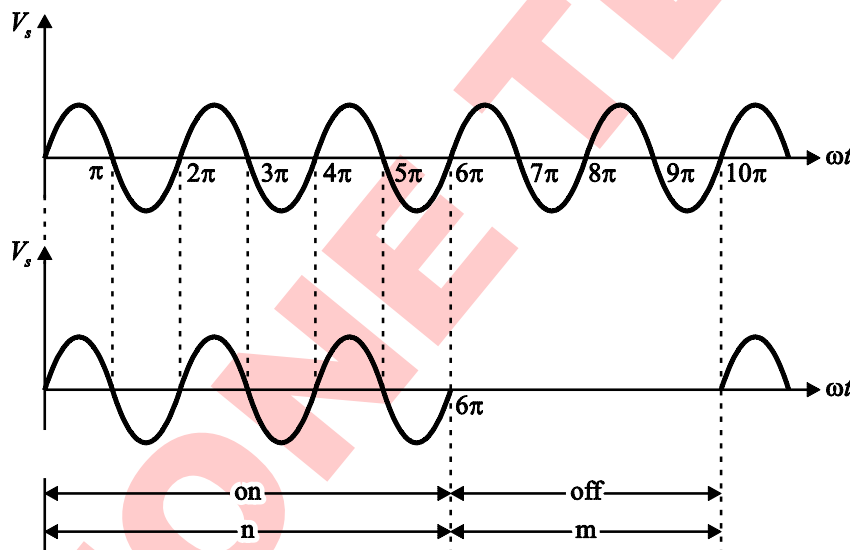
$$V_{or} = \left[\frac{1}{\pi} \int_{\alpha}^{\pi} V_m^2 \sin^2 \omega t \cdot d(\omega t) \right]^{1/2} = \frac{V_m}{\sqrt{2}} \left[\frac{1}{\pi} \left\{ (\pi - \alpha) + \frac{\sin 2\alpha}{2} \right\} \right]^{1/2}$$

By varying α , the rms value of the voltage across the lighting load is controlled.

(ii) Integral cycle control:

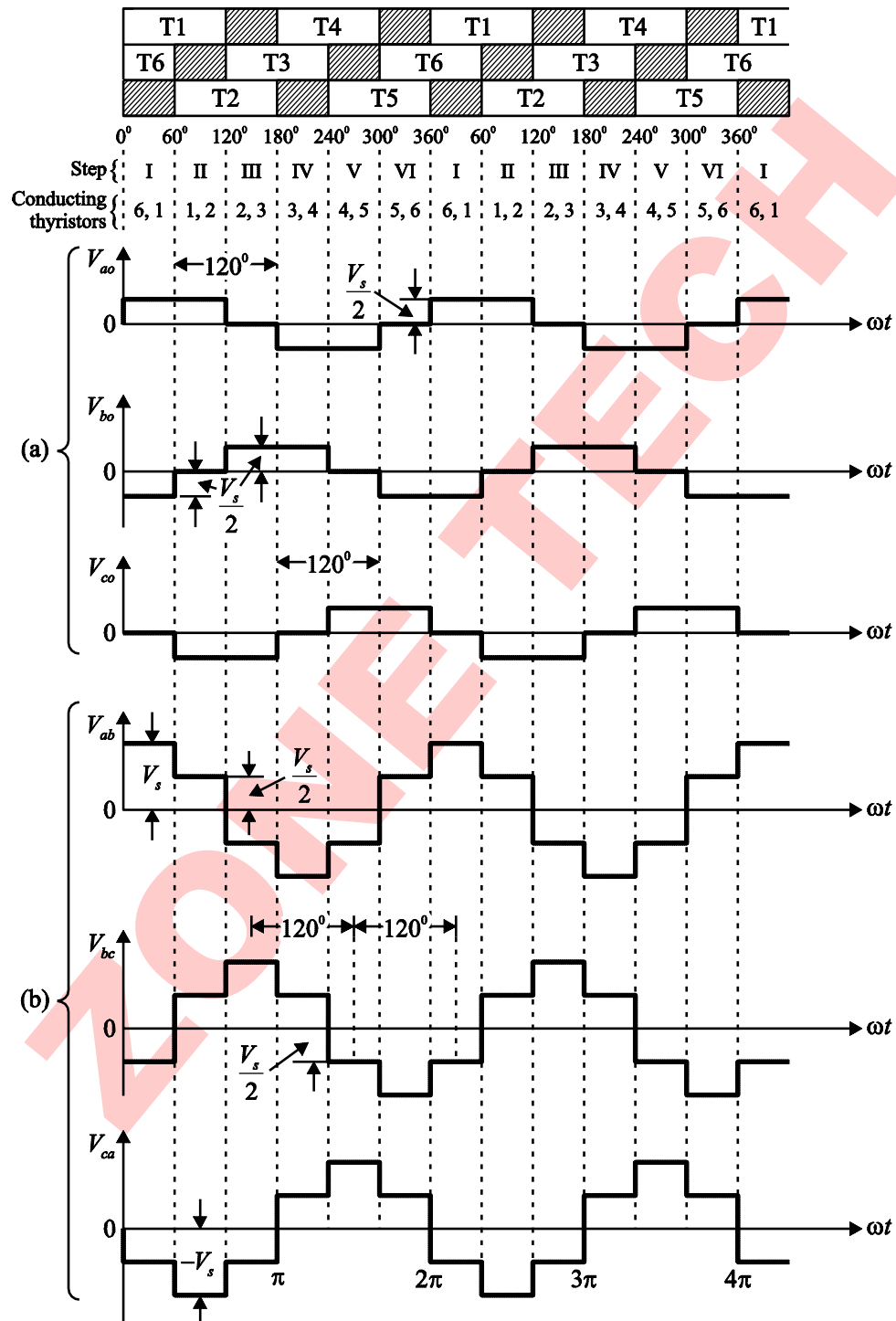
For heating loads, almost no variation in temperature will be noticed if control is achieved connecting the load to source for some on-cycle and then disconnecting the load for some off-cycles. This form of power control is called integral cycle control. So integral cycle control consists of switching on the supply to load for an integral number of cycles and then switching off the supply for a further number of integral cycles.

The source energises the load for n cycles. When gate pulses are withdrawn, load remains off for m cycles. By varying the number of n and m cycles, power delivered to load can be regulated as desired.



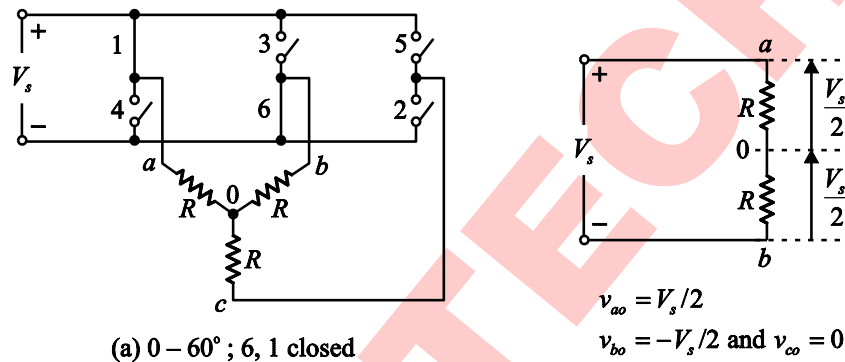
62.

Sol.

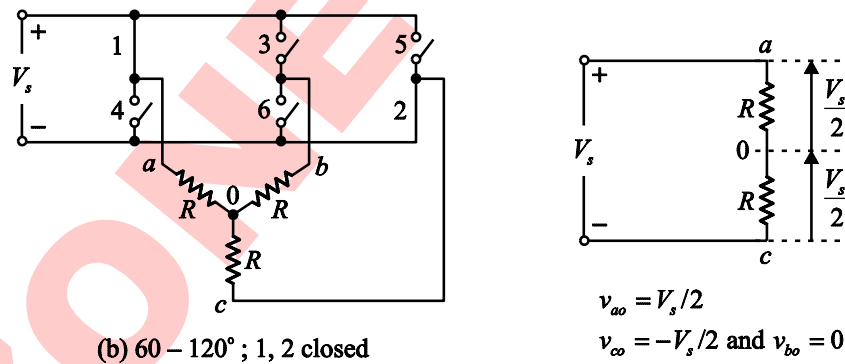


For this inverter, a table giving the sequence of firing the six thyristors is prepared as shown in the top of Fig. In this table, first rows shows that T1 conducts for 120° and for the next 60° , neither T1 nor T4 conducts. Now T4 is turned on at $\omega t = 180^\circ$ and it further conducts for 120° , i.e. from $\omega t = 180^\circ$ to $\omega t = 300^\circ$. This means that for 60° interval from $\omega t = 120^\circ$ to $\omega t = 180^\circ$, series connected SCRs do not conduct. At $\omega t = 300^\circ$, T4 is turned off, then 60° interval elapses before T1 is turned on again at $\omega t = 360^\circ$. In the second row, T3 is turned on at $\omega t = 120^\circ$ as in 180° mode inverter. Now T3 conducts for 120° , then 60° interval elapses during which neither T3 nor T6 conducts. At $\omega t = 300^\circ$, T6 is turned on, it conducts for 120° and then 60° interval elapses after which T3 is turned on again. The third row is also completed similarly. This table shows that T6, T1 should be gated for step I; T1, T2 for step II; T2, T3 for step III and so on. During each steps only two thyristors conduct for this inverter - one from the upper group and one from the lower group.

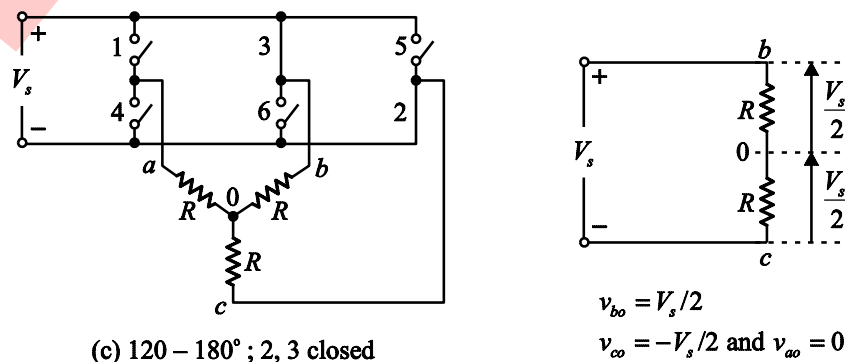
Step-I :



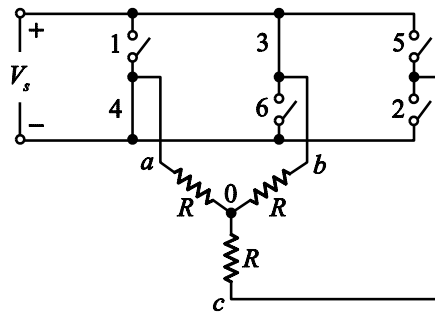
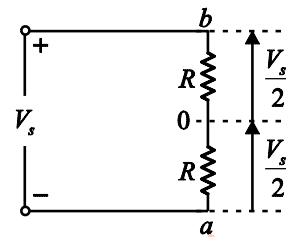
Step-II :



Step - III :



Step-IV :

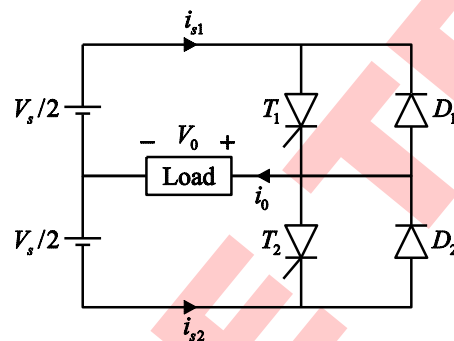
(d) $180 - 240^\circ$; 3, 4 closed

$$v_{bo} = V_s/2$$

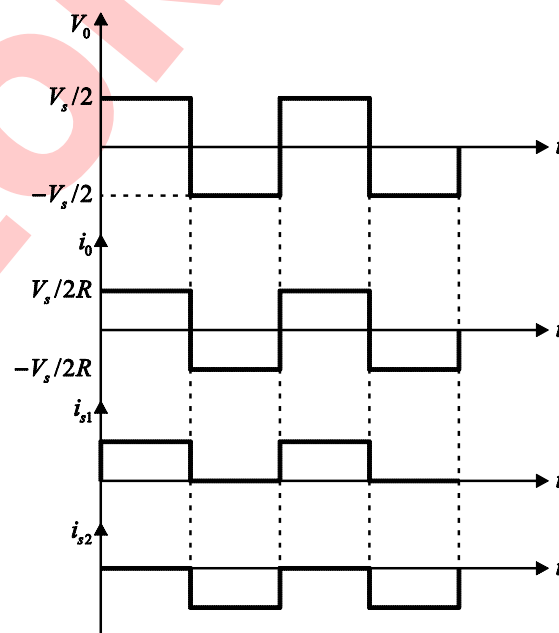
$$v_{ao} = -V_s/2 \text{ and } v_{co} = 0$$

Fig. Equivalent circuits for a 3-phase six-step 120° mode inverter with balanced star-connected resistive load.

63.
Sol.



Single phase half-bridge inverter



The rms value of output voltage

$$V_{0(\text{rms})} = \frac{V_s}{2} = \frac{400}{2} = 200 \text{ V}$$

$$I_{0(\text{rms})} = \frac{V_s}{2R} = \frac{400}{2 \times 10} = 20 \text{ A}$$

Power delivered to the load = $200 \times 20 = 4000 \text{ W}$

When (T_1) is conducting, power to load is delivered by upper source $\left(\frac{V_s}{2}\right)$ and when (T_2) is 'ON' lower source deliver power to load.

Power delivered by each source = $\frac{V_s}{2} I_s(\text{av}) = \frac{V_s}{2} [I_{s1}(\text{av}) + I_{s2}(\text{av})]$

Here $I_{s1} = \left(\frac{V_s}{2R}\right) \times \frac{1}{2} = \frac{V_s}{4R}$

Similarly $I_{s2} = \left(\frac{V_s}{2R}\right) \times \frac{1}{2} = \frac{V_s}{4R}$

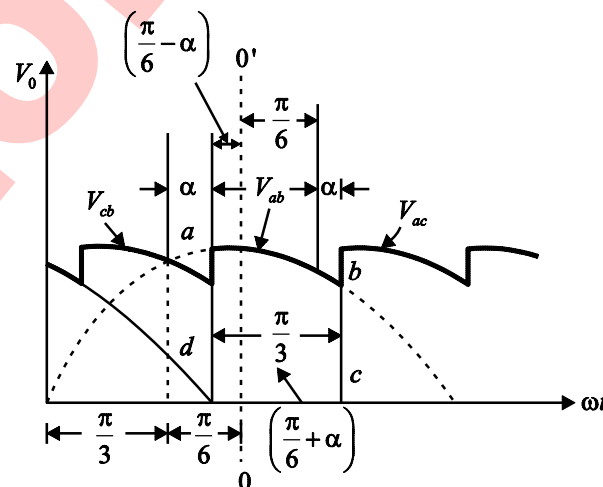
$\therefore$ Power delivered by each source = $\left(\frac{V_s}{2}\right) \times \left(\frac{V_s}{2R}\right)$

$$= \left(\frac{400}{2}\right) \times \left(\frac{400}{2 \times 10}\right) = 200 \times 20 = 4000 \text{ W}$$

$\therefore$ $\text{pf} = \frac{\text{power delivered to load}}{\text{supplied (VI)}} = 1 \text{ (unity power factor)}$

64.

Sol.



Average output voltage can be obtained by finding the dashed area abed over periodic cycle with $00'$ as the origin at the maximum value of V_{ab} . V_0 is given by

$$V_0 = \frac{3}{\pi} \int_{-\left(\frac{\pi}{6}-\alpha\right)}^{\left(\frac{\pi}{6}+\alpha\right)} V_{ml} \cos \omega t (d\omega t)$$

$$= \frac{3V_{ml}}{\pi} \left[\sin \left(\alpha + \frac{\pi}{6} \right) + \sin \left(\alpha - \frac{\pi}{6} \right) \right] = \frac{3V_{ml}}{\pi} \cos \alpha$$

For $\alpha < \pi/3$

$$V_0 = \frac{3}{\pi} \int_{-\left(\frac{\pi}{6}-\alpha\right)}^{\pi/2} V_{ml} \cos \omega t \cdot d\omega t$$

$$= \frac{3}{\pi} [V_{ml} \sin \omega t]_{-\left(\frac{\pi}{6}-\alpha\right)}^{\pi/2} = \frac{3V_{ml}}{\pi} \left(1 + \cos \left(\alpha + \frac{\pi}{3} \right) \right)$$

The output voltage waveform for $\alpha = \frac{\pi}{2}$, is illustrated below.

