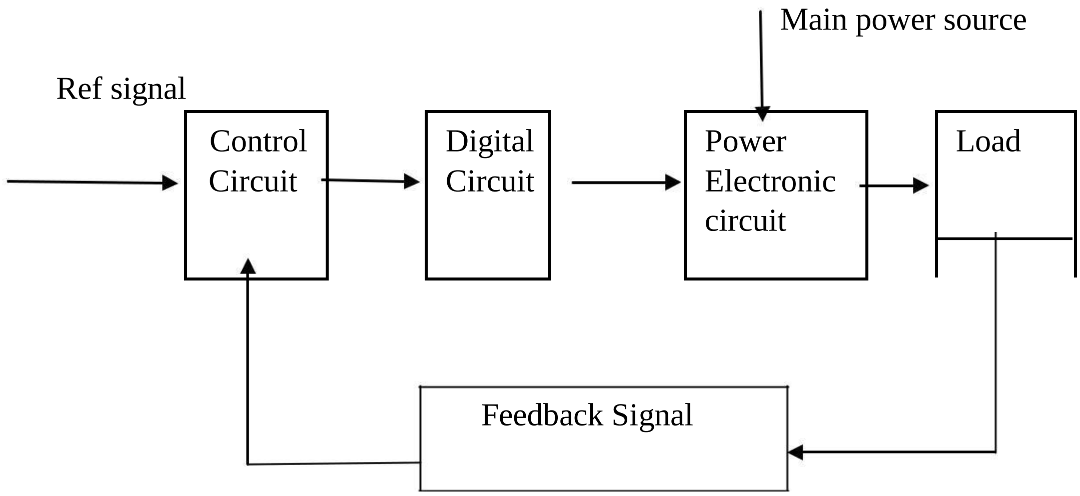
UNIT – I

POWER SEMICONDUCTOR DEVICES

The control of electric motor drives requires control of electric power. Power electronics have eased the concept of power control. Power electronics signifies the word power electronics and control or we can say the electronic that deal with power equipment for power control.



Power electronics based on the switching of power semiconductor devices. With the development of power semiconductor technology, the power handling capabilities and switching speed of power devices have been improved tremendously.

APPLICATIONS OF POWER ELECTRONICS

Advertising, air con ditioning, aircraft power supplies, alarms, appliances – (domestic and industrial), audio am plifiers, battery chargers, blenders, blowers, boilers, burglar alarms, cement kiln, chemic al processing, clothes dryers, computers, conveyors, cranes and hoists, dimmers (light dimmers), displays, electric door openers, electric dryers, electric fans, electric vehicles , electromagnets, electro mechanical electro plating, electronic ignition, electrostatic p recipitators, elevators, fans, flashers, food mixers, food warmer trays, fork lift trucks, furnaces, games, garage door openers, gas turbine starting, generator exciters, grin ders, hand power tools, heat controls, high frequency lighting, HVDC transmission, in duction heating, laser power supplies, latching relays, light flashers, linear induction motor controls, locomotives, machine tools, magnetic recording, magnets, mass transit railway system, mercury arc lamp ballasts, mining, model trains, motor controls, motor drives, movie projectors, nuclear reactor control rod, oil well drilling, oven controls, paper mills, particle accelerators, phonographs, photo copiers, power suppliers, printing press, pumps and compressors, radar/sonar power supplies, refrigerators, regulators, RF amplifiers, security systems, servo systems, sewing machines, solar power supplies, solid-state contactors, solid- state relays, static circuit breakers, static relays, steel mills, synchronous motor starting, TV circuits, temperature controls, timers and toys, traffic signal controls, trains, TV deflection circuits, ultrasonic generators, UPS, vacuum cleaners, VAR compensation, vending machines, VLF transmitters, voltage regulators, washing machines, welding equipment.

POWER ELECTRONIC APPLICATIONS

COMMERCIAL APPLICATIONS

Heating Systems Ventilating, Air Conditioners, Central Refrigeration, Lighting, Computers and Office equipments, Uninterruptible Power Supplies (UPS), Elevators, and Emergency Lamps.

DOMESTIC APPLICATIONS

Cooking Equipments, Lighting, Heating, Air Conditioners, Refrigerators & Freezers, Personal Computers, Entertainment Equipments, UPS.

INDUSTRIAL APPLICATIONS

Pumps, compressors, blowers and fans. Machine tools, arc furnaces, induction furnaces, lighting control circuits, industrial lasers, induction heating, welding equipments.

AEROSPACE APPLICATIONS

Space shuttle power supply systems, satellite power systems, aircraft power systems.

TELECOMMUNICATIONS

Battery chargers, power supplies (DC and UPS), mobile cell phone battery chargers.

TRANSPORTATION

Traction control of electric vehicles, battery chargers for electric vehicles, electric locomotives, street cars, trolley buses, automobile electronics including engine controls.

UTILITY SYSTEMS

High voltage DC transmission (HVDC), static VAR compensation (SVC), Alternative energy sources (wind, photovoltaic), fuel cells, energy storage systems, induced draft fans and boiler feed water pumps.

POWER SEMICONDUCTOR DEVICES

- Power Diodes.
- Power Transistors (BJT's).
- Power MOSFETS.
- IGBT's.
- Thyristors

  Thyristors are a family of p-n-p-n structured power semiconductor switching devices

- SCR's (Silicon Controlled Rectifier)

  The silicon controlled rectifier is the most commonly and widely used member of the thyristor family. The family of thyristor devices include SCR's, Diacs, Triacs, SCS, SUS, LASCR's and so on.

POWER SEMICONDUCTOR DEVICES USED IN POWER ELECTRONICS

The first thyristor or the SCR was developed in 1957. The conventional Thyristors (SCR's) were exclusively used for power control in industrial applications until 1970. After 1970, various types of power semiconductor devices were developed and became commercially available. The power semiconductor devices can be divided broadly into five types

- Power Diodes.
- Thyristors.

- Power BJT's.
- Power MOSFET's.
- Insulated Gate Bipolar Transistors (IGBT's).
- Static Induction Transistors (SIT's).

The Thyristors can be subdivided into different types

- Forced-commutated Thyristors (Inverter grade Thyristors)
- Line-commutated Thyristors (converter-grade Thyristors)
- Gate-turn off Thyristors (GTO).
- Reverse conducting Thyristors (RCT's).
- Static Induction Thyristors (SITH).
- Gate assisted turn-off Thyristors (GATT).
- Light activated silicon controlled rectifier (LASCR) or Photo SCR's.
- MOS-Controlled Thyristors (MCT's).

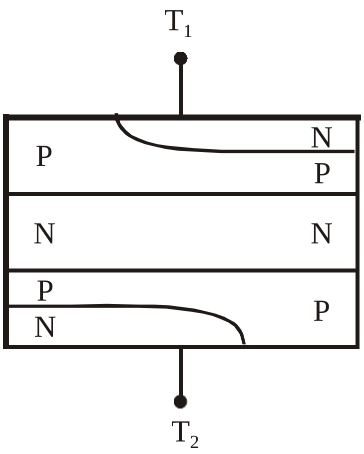
- Time required for emitter current to diffuse across the base region into the collector region once the base emitter junction is forward biased. The turn on time t_{on} ranges from 10 to 300 ns. Base current is normally more than the minimum required to saturate the transistor. As a result excess minority carrier charge is stored in the base region.

When the input voltage is reversed from V_{B1} to $-V_{B2}$ the base current also abruptly changes but the collector current remains constant for a short time interval t_S called the storage time.

The reverse base current helps to discharge the minority charge carries in the base region and to remove the excess stored charge form the base region. Once the excess stored charge is removed the baser region the base current begins to fall towards zero. The fall-time t_f is the time taken for the collector current to fall from 90% to 10% of $I_{C(sat)}$. The turn off time t_{off} is the sum of storage time and the fall time. $t_{off} = t_s + t_f$

DIAC

A diac is a two terminal five layer semi-conductor bi-directional switching device. It can conduct in both directions. The device consists of two p-n-p-n sections in anti parallel as shown in figure. T_1 and T_2 are the two terminals of the device.



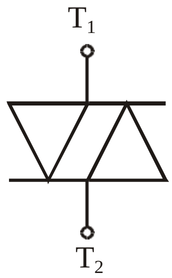


Fig.: Diac Structure

Fig.: Diac symbol

Figure above shows the symbol of diac. Diac will conduct when the voltage applied across the device terminals T_1& T_2exceeds the break over voltage..

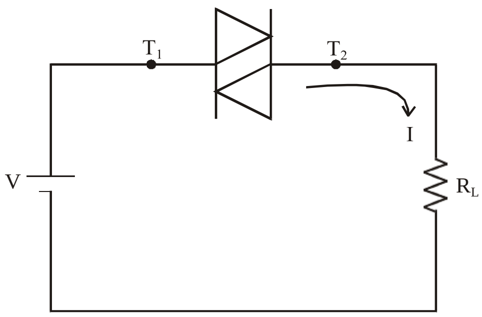


Fig. 1.1

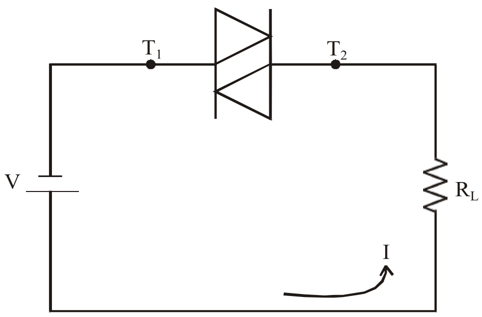


Fig. 1.2

Figure 1.1 shows the circuit diagram with T_1 positive with respect to T_2. When the voltage

across the device is less than the break over voltage V_{B01} a very small amount of current called leakage current flows through the device. During this period the device is in non-conducting or blocking mode. But once the voltage across the diac exceeds the break over voltage V_{B01} the diac turns on and begins to conduct. Once it starts conducting the current through diac becomes large and the device current has to be limited by connecting an external load resistance R_L, at

the same time the voltage across the diac decreases in the conduction state. This explain the forward characteristics.

Figure 1.2 shows the circuit diagram with T_2 positive with respect to T_1. The reverse

characteristics obtained by varying the supply voltage are identical with the forward characteristic as the device construction is symmetrical in both the directions.

In both the cases the diac exhibits negative resistance switching characteristic during conduction. i.e., current flowing through the device increases whereas the voltage across it decreases.

Figure below shows forward and reverse characteristics of a diac. Diac is mainly used for triggering triacs.

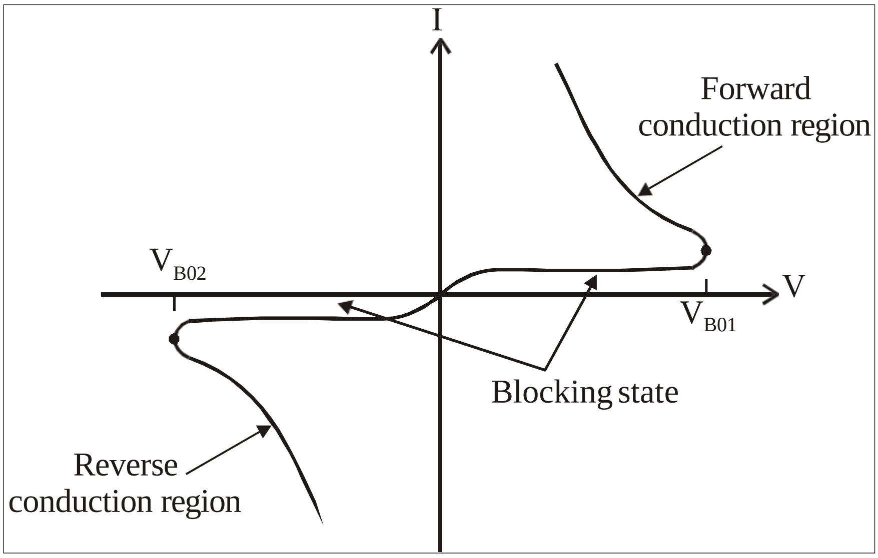


Fig.: Diac Characteristics

TRIAC

A triac is a three terminal bi-directional switching thyristor device. It can conduct in both directions when it is triggered into the conduction state. The triac is equivalent to two SCRs connected in anti-parallel with a common gate. Figure below shows the triac structure. It consists of three terminals viz., MT_2, MT_1 and gate G.

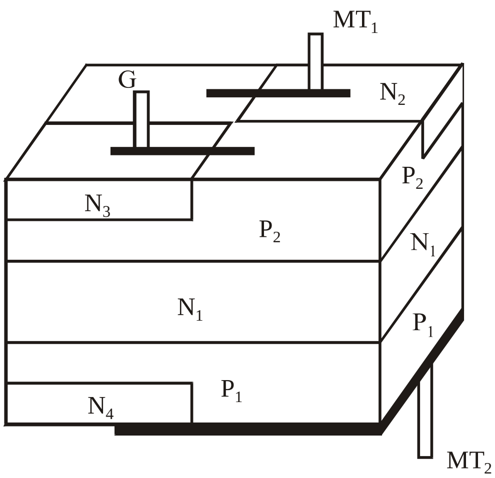


Fig. : Triac Structure

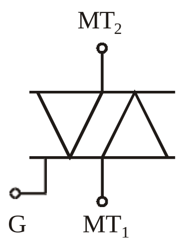


Fig. : Triac Symbol

The gate terminal G is near the MT_1 terminal. Figure above shows the triac symbol. MT_1 is the reference terminal to obtain the characteristics of the triac. A triac can be operated in four different modes depending upon the polarity of the voltage on the terminal MT_2 with respect to MT_1 and based on the gate current polarity.

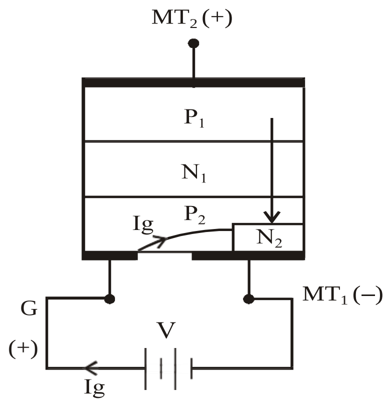
The characteristics of a triac is similar to that of an SCR, both in blocking and conducting states. A SCR can conduct in only one direction whereas triac can conduct in both directions.

TRIGGERING MODES OF TRIAC

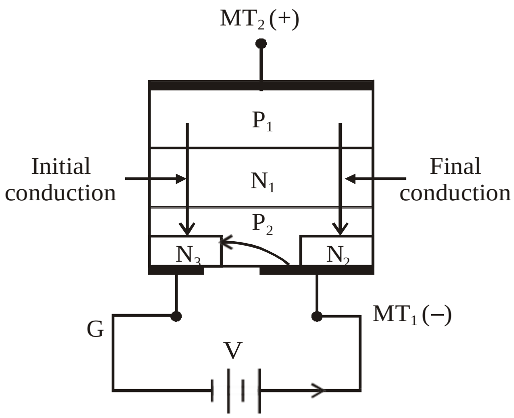
MODE 1 : MT_2 positive, Positive gate current (I^+ mode of operation)

When MT_2 and gate current are positive with respect to MT_1, the gate current flows through

P_2-N_2 junction as shown in figure below. The junction P_1-N_1 and P_2-N_2 are forward biased but junction N_1-P_2 is reverse biased. When sufficient number of charge carriers are injected in P_2 layer by the gate current the junction N_1-P_2 breakdown and triac starts conducting through $P_1N_1P_2N_2$ layers. Once triac starts conducting the current increases and its V-I characteristics is similar to that of thyristor. Triac in this mode operates in the first-quadrant.



MODE 2 : MT_2 positive, Negative gate current (I^- mode of operation)

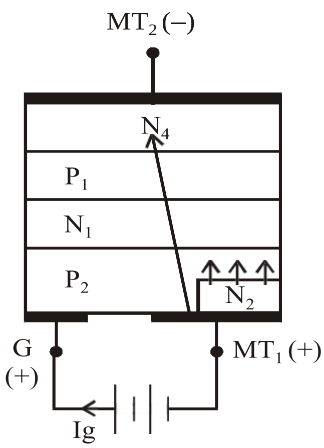


Ig

When MT_2 is positive and gate G is negative with respect to MT_1 the gate current flows through P_2-N_3 junction as shown in figure above. The junction P_1-N_1 and P_2-N_3 are forward biased but junction N_1-P_2 is reverse biased. Hence, the triac initially starts conducting through $P_1N_1P_2N_3$ layers. As a result the potential of layer between P_2-N_3 rises towards the potential of MT_2. Thus, a potential gradient exists across the layer P_2 with left hand region at a higher potential than the right hand region. This results in a current flow in P_2 layer from left to right, forward biasing the P_2N_2 junction. Now the right hand portion P_1-N_1 - P_2-N_2 starts conducting. The device operates in first quadrant. When compared to Mode 1, triac with MT_2 positive and negative gate current is less sensitive and therefore requires higher gate current for triggering.

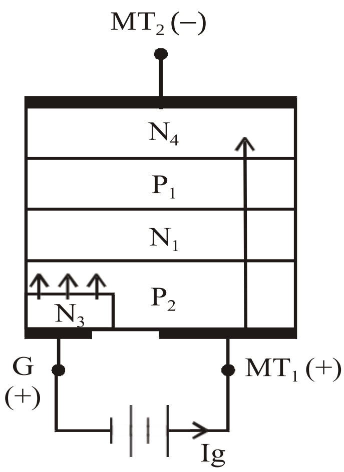
MODE 3 : MT_2 negative, Positive gate current (III^+ **mode of operation)**

When MT_2 is negative and gate is positive with respect to MT_1 junction P_2N_2 is forward biased and junction P_1-N_1 is reverse biased. N_2 layer injects electrons into P_2 layer as shown by arrows in figure below. This causes an increase in current flow through junction P_2-N_1. Resulting in breakdown of reverse biased junction N_1-P_1. Now the device conducts through layers $P_2N_1P_1N_4$ and the current starts increasing, which is limited by an external load.



The device operates in third quadrant in this mode. Triac in this mode is less sensitive and requires higher gate current for triggering.

MODE 4 : MT_2 negative, Negative gate current (III^- mode of operation)



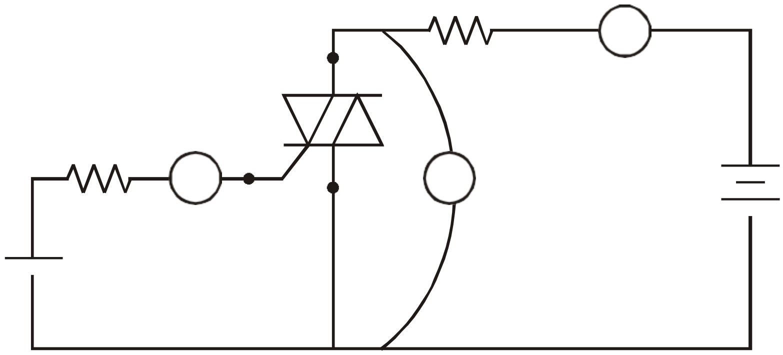
In this mode both MT_2 and gate G are negative with respect to MT_1, the gate current flows through P_2N_3 junction as shown in figure above. Layer N_3 injects electrons as shown by arrows into P_2 layer. This results in increase in current flow across P_1N_1 and the device will turn ON due to increased current in layer N_1. The current flows through layers $P_2N_1P_1N_4$. Triac is more sensitive in this mode compared to turn ON with positive gate current. (Mode 3).

Triac sensitivity is greatest in the first quadrant when turned ON with positive gate current and also in third quadrant when turned ON with negative gate current. when MT_2 is positive with respect to MT_1 it is recommended to turn on the triac by a positive gate current. When MT_2 is negative with respect to MT_1 it is recommended to turn on the triac by negative gate current.

Therefore Mode 1 and Mode 4 are the preferred modes of operation of a triac (I^+ mode and III^- mode of operation are normally used).

TRIAC CHARACTERISTICS

Figure below shows the circuit to obtain the characteristics of a triac. To obtain the characteristics in the third quadrant the supply to gate and between MT_2 and MT_1 are reversed.



+

Figure below shows the V-I Characteristics of a triac. Triac is a bidirectional switching device. Hence its characteristics are identical in the first and third quadrant. When gate current is increased the break over voltage decreases.

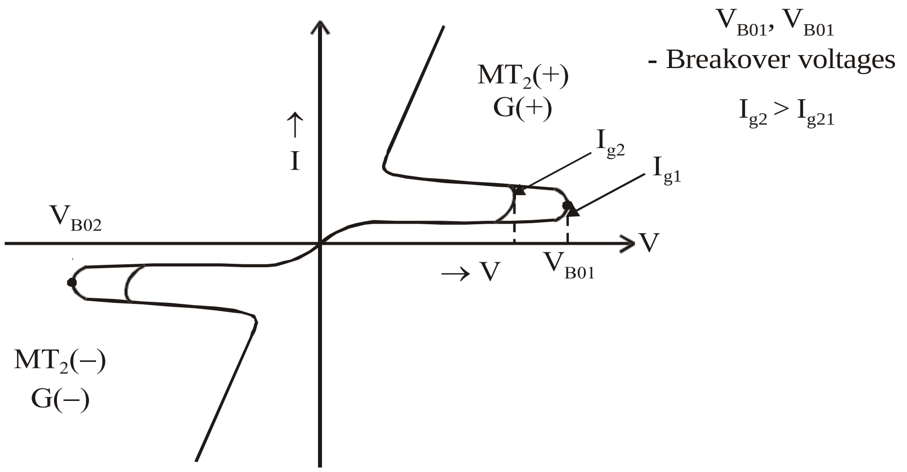
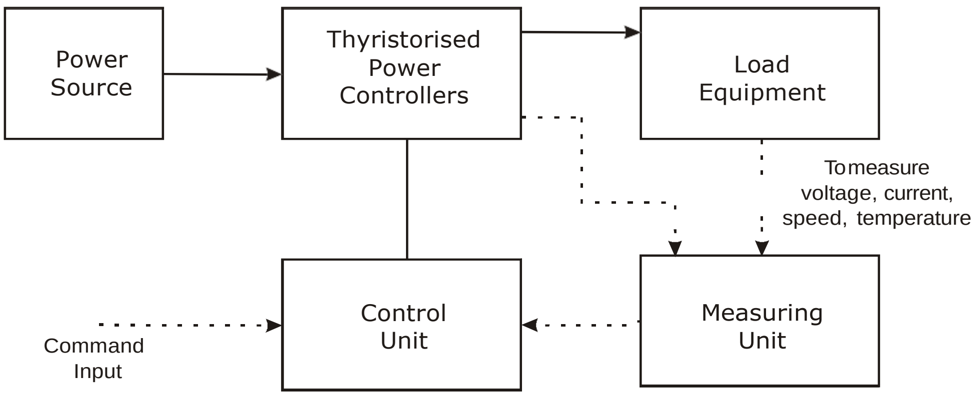


Fig.: Triac Characteristic

Triac is widely used to control the speed of single phase induction motors. It is also used in domestic lamp dimmers and heat control circuits, and full wave AC voltage controllers.

THYRISTORISED POWER CONTROLLERS

Block diagram given below, shows the system employing a thyristorised power controller. The main power flow between the input power source and the load is shown by solid lines.



Thyristorised power controllers are widely used in the industry. Old/conventional controllers including magnetic amplifiers, mercury arc rectifiers, thyratrons, ignitrons, rotating amplifiers, resistance controllers have been replaced by thyristorised power controllers in almost all the applications.

A typical block diagram of a thyristorised power converter is shown in the above figure.

The thyristor power converter converts the available power from the source into a suitable form to run the load or the equipment. For example the load may be a DC motor drive which requires DC voltage for its operation. The available power supply is AC power supply as is often the case. The thyristor power converter used in this case is a AC to DC power converter which converts the input AC power into DC output voltage to feed to the DC motor. Very often a measuring unit or an instrumentation unit is used so as to measure and monitor the output parameters like the output voltage, the load current, the speed of the motor or the temperature etc. The measuring unit will be provided with meters and display devices so that the output parameters can be seen and noted. The control unit is employed to control the output of the thyristorised power converter so as to adjust the output voltage / current to the desired value to obtain optimum performance of the load or equipment. The signal from the control unit is used to adjust the phase angle / trigger angle of the Thyristors in the power controller so as to vary the output voltage to the desired value.

SOME IMPORTANT APPLICATIONS OF THYRISTORISED POWER CONTROLLERS

- Control of AC and DC motor drives in rolling mills, paper and textile mills, traction vehicles, mine winders, cranes, excavators, rotary kilns, ventilation fans, compression etc.
- Uninterruptible and stand by power supplies for critical loads such as computers, special high tech power supplies for aircraft and space applications.
- Power control in metallurgical and chemical processes using arc welding, induction heating, melting, resistance heating, arc melting, electrolysis, etc.
- Static power compensators, transformer tap changers and static contactors for industrial power systems.
- Power conversion at the terminals of a HVDC transmission systems.
- High voltage supplies for electrostatic precipitators and x-ray generators.
- Illumination/light control for lighting in stages, theaters, homes and studios.
- Solid state power controllers for home/domestic appliances.

ADVANTAGES OF THYRISTORISED POWER CONTROLLERS

- High efficiency due to low losses in the Thyristors.
- Long life and reduced/minimal maintenance due to the absence of mechanical wear.
- Control equipments using Thyristors are compact in size.
- Easy and flexibility in operation due to digital controls.
- Faster dynamic response compared to the electro mechanical converters.
- Lower acoustic noise when compared to electro magnetic controllers, relays and contactors.

DISADVANTAGES OF THYRISTORISED POWER CONTROLLERS

- All the thyristorised power controllers generate harmonics (unwanted frequency components) due to the switching ON and OFF of the thyristors. These harmionics adversely affect the performance of the load connected to them. For example when the load are motors, there are additional power losses (harmonic power loss) torque harmonics, and increase in acoustic noise.
- The generated harmonics are injected into the supply lines and thus adversely affect the other loads/equipments connected to the supply lines.
- In some applications example: traction, there is interference with the commutation circuits due to the power supply line harmonics and due to electromagnetic radiation.
- The thyristorised AC to DC converters and AC to AC converters can operate at low power factor under some conditions.
- Special steps are then taken for correcting the line supply power factor (by installing PF improvement apparatus).
- The thyristorised power controllers have no short time over loading capacity and therefore they must be rated for maximum loading conditions. This leads to an increase in the cost of the equipment.
- Special protection circuits must be employed in thyristorised power controllers in order to protect and safe guard the expensive thyristor devices. This again adds to the system cost.

TYPES OF POWER CONVERTERS or THYRISTORISED POWER CONTROLLERS

For the control of electric power supplied to the load or the equipment/machinery or for power conditioning the conversion of electric power from one form to other is necessary and the switching characteristic of power semiconductor devices (Thyristors) facilitate these conversions

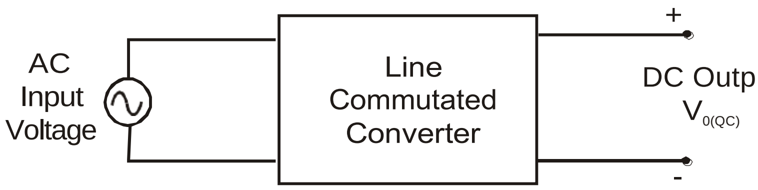
The thyristorised power converters are referred to as the static power converters and they perform the function of power conversion by converting the available input power supply in to output power of desired form.

The different types of thyristor power converters are

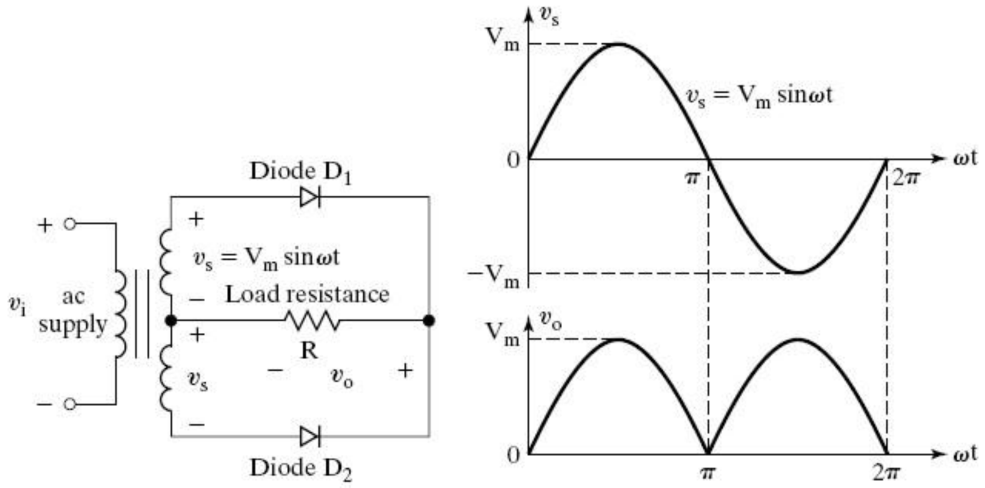
- Diode rectifiers (uncontrolled rectifiers).
- Line commutated converters or AC to DC converters (controlled rectifiers)
- AC voltage (RMS voltage) controllers (AC to AC converters).

- Cyclo converters (AC to AC converters at low output frequency).
- DC choppers (DC to DC converters).
- Inverters (DC to AC converters).

LINE COMMUTATED CONVERTERS (AC TO DC CONVERTERS)

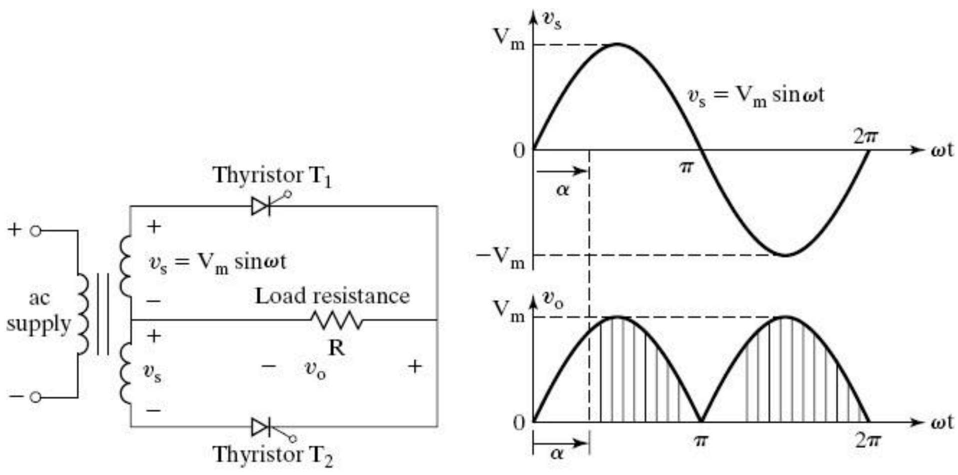


These are AC to DC converters. The line commutated converters are AC to DC power converters. These are also referred to as controlled rectifiers. The line commutated converters (controlled rectifiers) are used to convert a fixed voltage, fixed frequency AC power supply to obtain a variable DC output voltage. They use natural or AC line commutation of the Thyristors.



(a) Circuit diagram (b) Voltage waveforms

Fig: A Single Phase Full Wave Uncontrolled Rectifier Circuit (Diode Full Wave Rectifier) using a Center Tapped Transformer



(a) Circuit diagram (b) Voltage waveforms

Fig: A Single Phase Full Wave Controlled Rectifier Circuit (using SCRs) using a Center Tapped Transformer

Different types of line commutated AC to DC converters circuits are

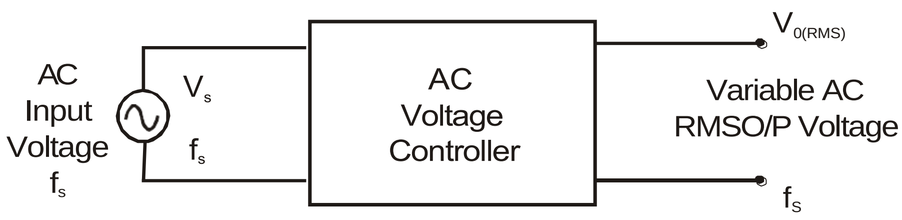
- Diode rectifiers – Uncontrolled Rectifiers
- Controlled rectifiers using SCR's.
    - Single phase controlled rectifier.
    - Three phase controlled rectifiers.

Applications of Line Commutated Converters

AC to DC power converters are widely used in

- Speed control of DC motor in DC drives.
- UPS.
- HVDC transmission.
- Battery Chargers.

AC VOLTAGE REGULATORS OR RMS VOLTAGE CONTROLLERS (AC TO AC CONVERTERS)



The AC voltage controllers convert the constant frequency, fixed voltage AC supply into variable AC voltage at the same frequency using line commutation.

AC regulators (RMS voltage controllers) are mainly used for

- Speed control of AC motor.
- Speed control of fans (domestic and industrial fans).
- AC pumps.

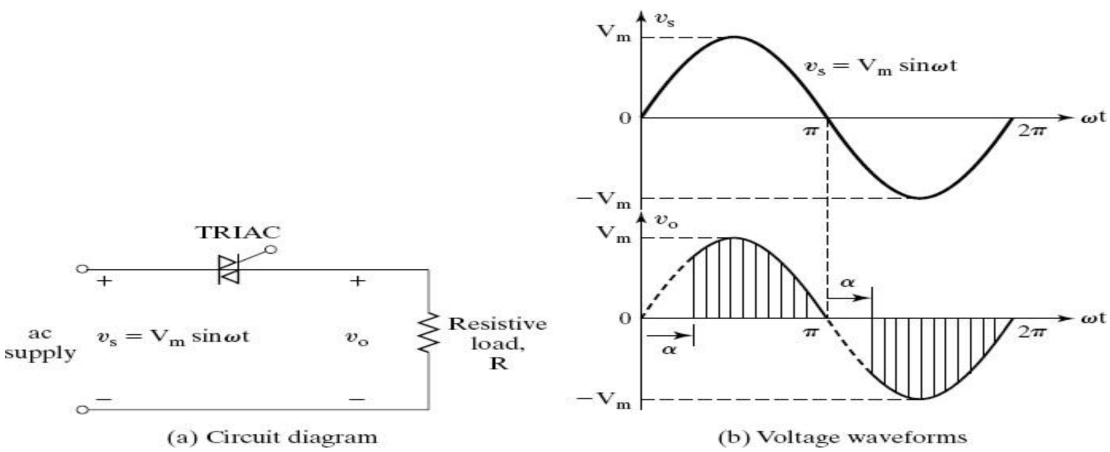
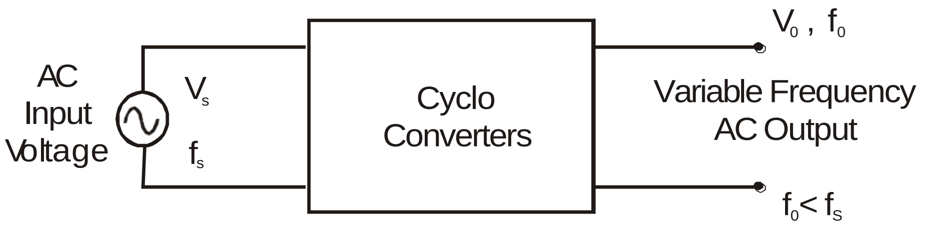


Fig: A Single Phase AC voltage Controller Circuit (AC-AC Converter using a TRIAC)

CYCLO CONVERTERS (AC TO AC CONVERTERS WITH LOW OUTPUT FREQUENCY)

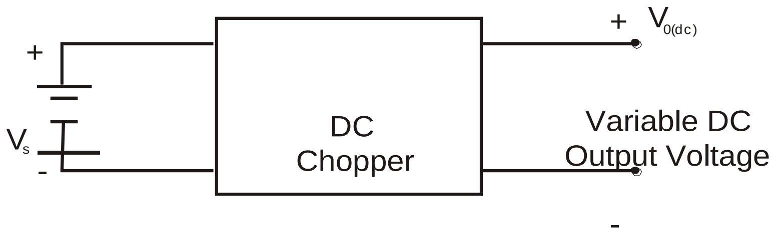


The cyclo converters convert power from a fixed voltage fixed frequency AC supply to a variable frequency and variable AC voltage at the output.

The cyclo converters generally produce output AC voltage at a lower output frequency. That is output frequency of the AC output is less than input AC supply frequency.

Applications of cyclo converters are traction vehicles and gearless rotary kilns.

CHOPPERS (DC TO DC CONVERTERS)



The choppers are power circuits which obtain power from a fixed voltage DC supply and convert it into a variable DC voltage. They are also called as DC choppers or DC to DC converters. Choppers employ forced commutation to turn off the Thyristors. DC choppers are further classified into several types depending on the direction of power flow and the type of commutation. DC choppers are widely used in

- Speed control of DC motors from a DC supply.
- DC drives for sub-urban traction.
- Switching power supplies.

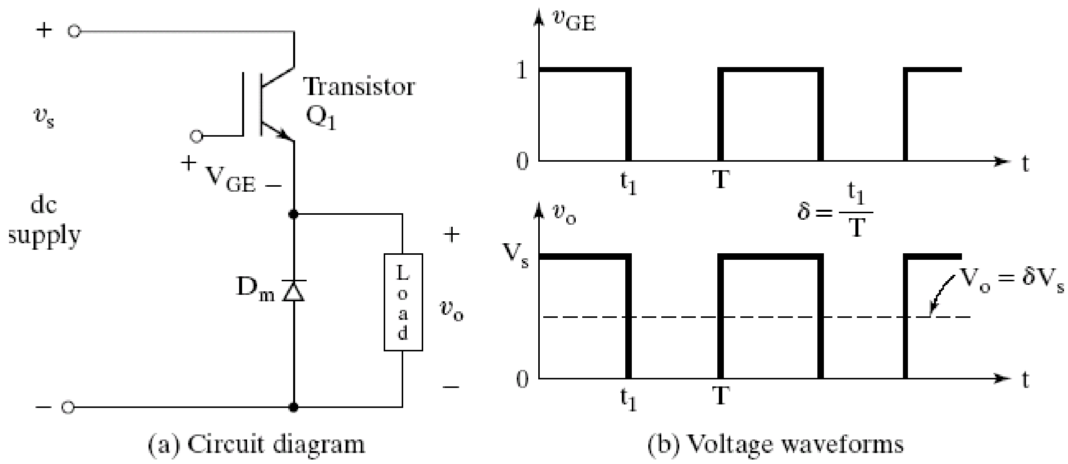
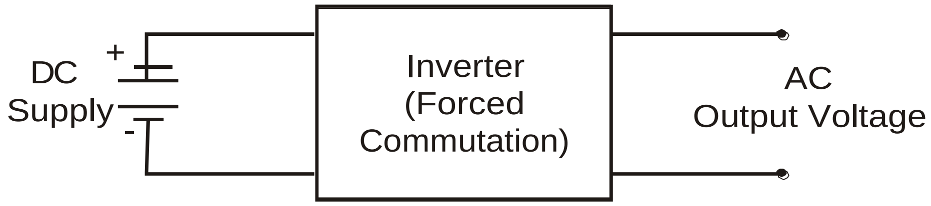


Fig: A DC Chopper Circuit (DC-DC Converter) using IGBT

INVERTERS (DC TO AC CONVERTERS)



The inverters are used for converting DC power from a fixed voltage DC supply into an AC output voltage of variable frequency and fixed or variable output AC voltage. The inverters also employ force commutation method to turn off the Thyristors.

Application of inverters are in

- Industrial AC drives using induction and synchronous motors.
- Uninterrupted power supplies (UPS system) used for computers, computer labs.

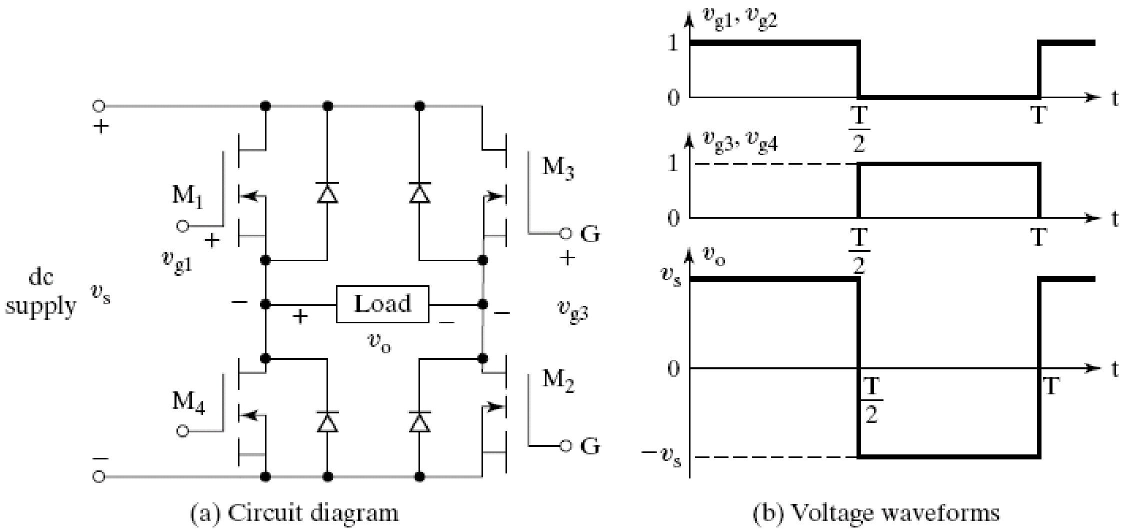


Fig: Single Phase DC-AC Converter (Inverter) using MOSFETS

DESIGN OF POWER ELECTRONICS CIRCUITS

The design and study of power electronic circuits involve

- Design and study of power circuits using Thyristors, Diodes, BJT's or MOSFETS.
- Design and study of control circuits.
- Design and study of logic and gating circuits and associated digital circuits.
- Design and study of protection devices and circuits for the protection of thyristor power devices in power electronic circuits.

The power electronic circuits can be classified into six types

- Diode rectifiers (uncontrolled rectifiers)
- AC to DC converters (Controlled rectifiers)
- AC to AC converters (AC voltage controllers)

- DC to DC converters (DC choppers)
- DC to AC converters (Inverters)
- Static Switches (Thyristorized contactors)

THYRISTORS

A thyristor is the most important type of power semiconductor devices. They are extensively used in power electronic circuits. They are operated as bi-stable switches from non-conducting to conducting state.

A thyristor is a four layer, semiconductor of p-n-p-n structure with three p-n junctions. It has three terminals, the anode, cathode and the gate.

The word thyristor is coined from thyratron and transistor. It was invented in the year 1957 at Bell Labs. The Different types of Thyristors are

- Silicon Controlled Rectifier (SCR).
- TRIAC
- DIAC
- Gate Turn Off Thyristor (GTO)

SILICON CONTROLLED RECTIFIER (SCR)

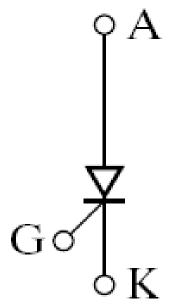


Fig.: Symbol

The SCR is a four layer three terminal device with junctions J_1, J_2, J_3 as shown. The construction of SCR shows that the gate terminal is kept nearer the cathode. The approximate thickness of each layer and doping densities are as indicated in the figure. In terms of their lateral dimensions Thyristors are the largest semiconductor devices made. A complete silicon wafer as large as ten centimeter in diameter may be used to make a single high power thyristor.

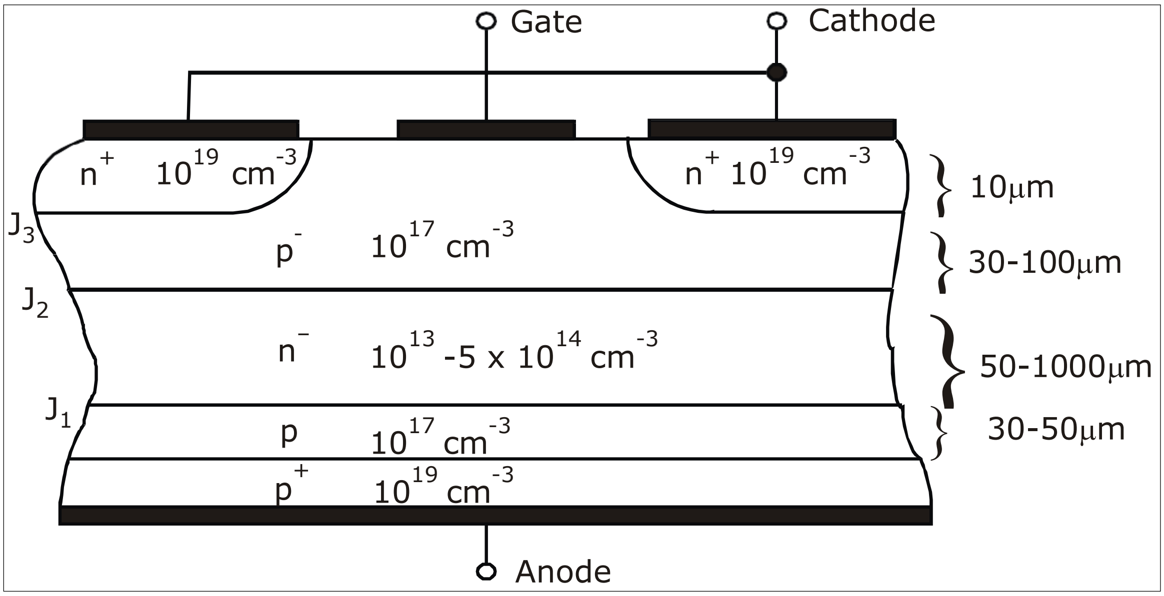


Fig.: Structure of a generic thyristor

QUALITATIVE ANALYSIS

When the anode is made positive with respect the cathode junctions J_1 & J_3 are forward biased and junction J_2 is reverse biased. With anode to cathode voltage V_{AK} being small, only leakage current flows through the device. The SCR is then said to be in the forward blocking state. If V_{AK} is further increased to a large value, the reverse biased junction J_2 will breakdown due to avalanche effect resulting in a large current through the device. The voltage at which this phenomenon occurs is called the forward breakdown voltage V_{BO}. Since the other junctions J_1 & J_3 are already forward biased, there will be free movement of carriers across all three junctions resulting in a large forward anode current. Once the SCR is switched on, the voltage drop across it is very small, typically 1 to 1.5V. The anode current is limited only by the external impedance present in the circuit.

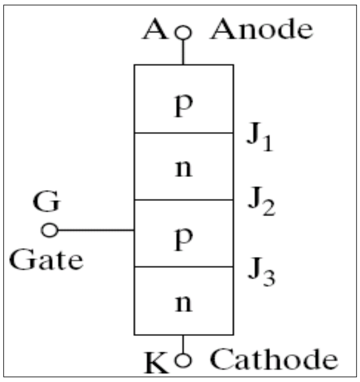
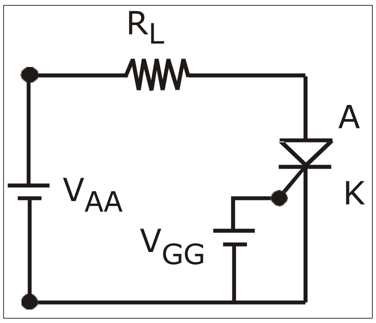


Fig.: Simplified model of a thyristor

Although an SCR can be turned on by increasing the forward voltage beyond V_{BO}, in practice, the forward voltage is maintained well below V_{BO} and the SCR is turned on by applying a positive voltage between gate and cathode. With the application of positive gate voltage, the leakage current through the junction J_2 is increased. This is because the resulting gate current consists mainly of electron flow from cathode to gate. Since the bottom end layer is heavily doped as compared to the p-layer, due to the applied voltage, some of these electrons reach junction J_2 and add to the minority carrier concentration in the p-layer. This raises the reverse leakage current and results in breakdown of junction J_2 even though the applied forward voltage is less than the breakdown voltage V_{BO}. With increase in gate current breakdown occurs earlier.

V-I CHARACTERISTICS



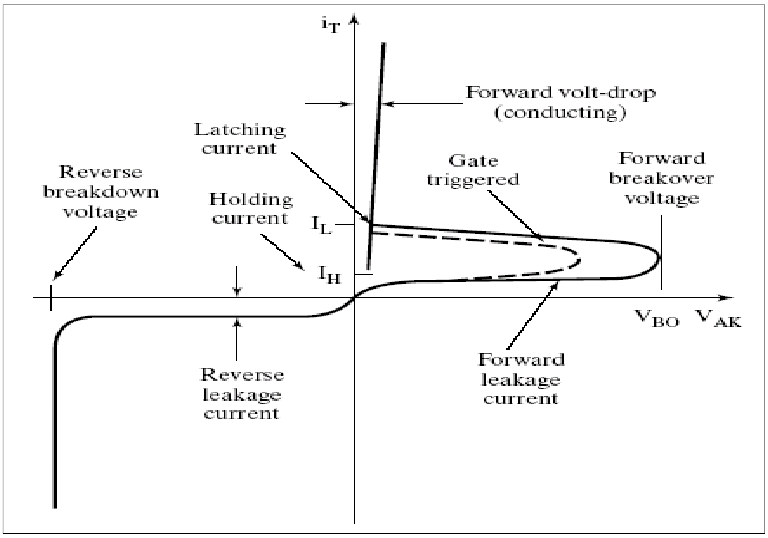


Fig: V-I Characteristics

A typical V-I characteristics of a thyristor is shown above. In the reverse direction the thyristor appears similar to a reverse biased diode which conducts very little current until avalanche breakdown occurs. In the forward direction the thyristor has two stable states or modes of operation that are connected together by an unstable mode that appears as a negative resistance on the V-I characteristics. The low current high voltage region is the forward blocking state or the off state and the low voltage high current mode is the on state. For the forward blocking state the quantity of interest is the forward blocking voltage V_{BO} which is defined for zero gate current. If a positive gate current is applied to a thyristor then the transition or break over to the on state will occur at smaller values of anode to cathode voltage as shown. Although not indicated the gate current does not have to be a dc current but instead can be a pulse of current having some minimum time duration. This ability to switch the thyristor by means of a current pulse is the reason for wide spread applications of the device.

However once the thyristor is in the on state the gate cannot be used to turn the device off. The only way to turn off the thyristor is for the external circuit to force the current through the device to be less than the holding current for a minimum specified time period.

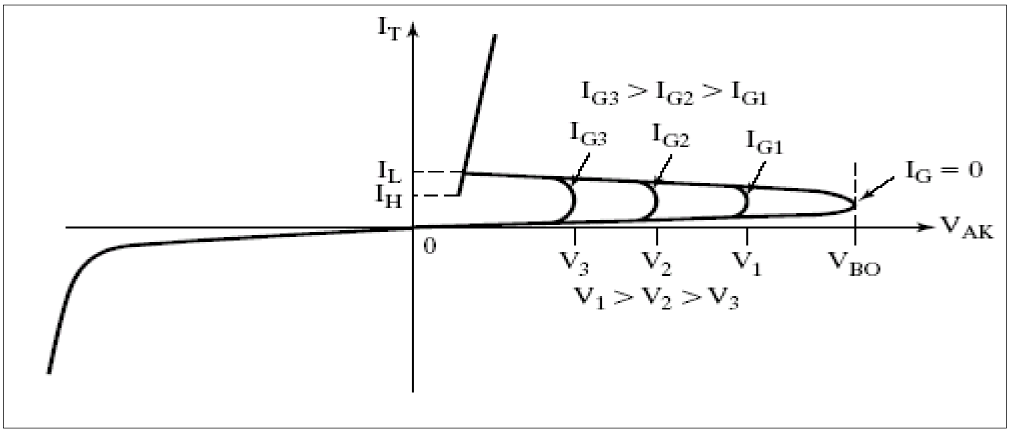


Fig.: Effects on gate current on forward blocking voltage

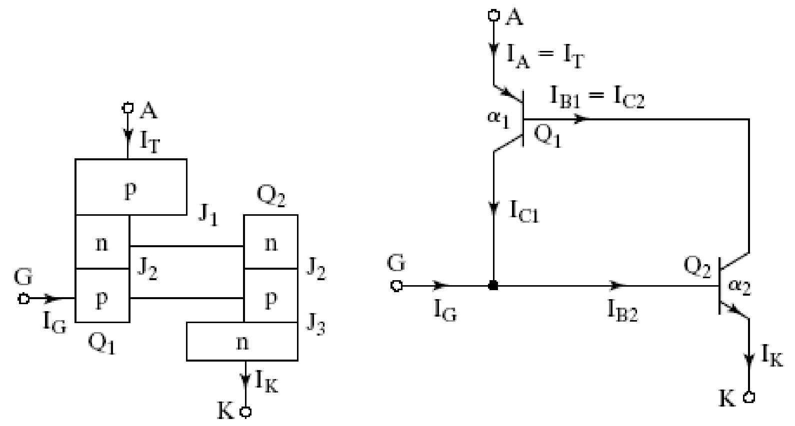
HOLDING CURRENT I_H

After an SCR has been switched to the on state a certain minimum value of anode current is required to maintain the thyristor in this low impedance state. If the anode current is reduced below the critical holding current value, the thyristor cannot maintain the current through it and reverts to its off state usually I_μ is associated with turn off the device.

LATCHING CURRENT I_L

After the SCR has switched on, there is a minimum current required to sustain conduction. This current is called the latching current. I_L associated with turn on and is usually greater than holding current.

TWO TRANSISTOR MODEL



(a) Basic structure (b) Equivalent circuit

The general transistor equations are,

$$I_C = \beta I_B + (1+\beta)\, I_{CBO}$$
$$I_C = \alpha I_E + I_{CBO}$$
$$I_E = I_C + I_B$$
$$I_B = I_E\,(1-\alpha) - I_{CBO}$$

The SCR can be considered to be made up of two transistors as shown in above figure.

Considering PNP transistor of the equivalent circuit,

$$I_{E_1} = I_A,\ I_C = I_{C_1},\ \alpha = \alpha_1,\ I_{CBO} = I_{CBO_1},\ I_B = I_{B_1}$$
$$\therefore \quad I_{B_1} = I_A\,(1-\alpha_1) - I_{CBO_1} \qquad ---(1)$$

Considering NPN transistor of the equivalent circuit,

$$I_C = I_{C_2},\ I_B = I_{B_2},\ I_{E_2} = I_K = I_A + I_G$$
$$I_{C_2} = \alpha_2 I_k + I_{CBO_2}$$
$$I_{C_2} = \alpha_2\,(I_A + I_G) + I_{CBO_2} \qquad ---(2)$$

From the equivalent circuit, we see that

$$\therefore \quad I_{C_2} = I_{B_1}$$
$$\Rightarrow \quad I_A = \frac{\alpha_2 I_g + I_{CBO1} + I_{CBO2}}{1-(\alpha_1+\alpha_2)}$$

Two transistors analog is valid only till SCR reaches ON state

Case 1: When $I_g = 0$,

$$I_A = \frac{I_{CBO_1} + I_{CBO_2}}{1-(\alpha_1+\alpha_2)}$$

The gain α_1 of transistor T_1 varies with its emitter current $I_E = I_A$. Similarly varies with $I_E = I_A + I_g = I_K$. In this case, with $I_g = 0$, α_2 varies only with I_A. Initially when the applied forward voltage is small, $(\alpha_1 + \alpha_2) < 1$.

If however the reverse leakage current is increased by increasing the applied forward voltage, the gains of the transistor increase, resulting in $(\alpha_1 + \alpha_2) \rightarrow 1$.

From the equation, it is seen that when $(\alpha_1 + \alpha_2) = 1$, the anode current I_A tends towards ∞ .

This explains the increase in anode current for the break over voltage V_{B0}.

Case 2: With gate current I_g applied.

When sufficient gate drive is applied, we see that $I_{B_2} = I_g$ is established. This in turn results in a current through transistor T_2, this increases α_2 of T_2. But with the existence of $I_{C_2} = \beta_2 I_{\beta_2} = \beta_2 I_g$, a current through T, is established. Therefore, $I_{C_1} = \beta_1 I_{B_1} = \beta_1 \beta_2 I_{B_2} = \beta_1 \beta_2 I_g$. This current in turn is connected to the base of T_2. Thus the base drive of T_2 is increased which in turn increases the base drive of T_1, therefore regenerative feedback or positive feedback is established between the two transistors. This causes $(\alpha_1 + \alpha_2)$ to tend to unity therefore the anode current begins to grow towards a large value. This regeneration continues even if I_g is removed this characteristic of SCR makes it suitable for pulse triggering; SCR is also called a Lathing Device.

SWITCHING CHARACTERISTICS (DYNAMIC CHARACTERISTICS)

THYRISTOR TURN-ON CHARACTERISTICS

When the SCR is turned on with the application of the gate signal, the SCR does not conduct fully at the instant of application of the gate trigger pulse. In the beginning, there is no appreciable increase in the SCR anode current, which is because, only a small portion of the silicon pellet in the immediate vicinity of the gate electrode starts conducting. The duration between 90% of the peak gate trigger pulse and the instant the forward voltage has fallen to 90% of its initial value is called the gate controlled / trigger delay time t_{gd}. It is also defined as the duration between 90% of the gate trigger pulse and the instant at which the anode current rises to 10% of its peak value. t_{gd} is usually in the range of 1μsec.

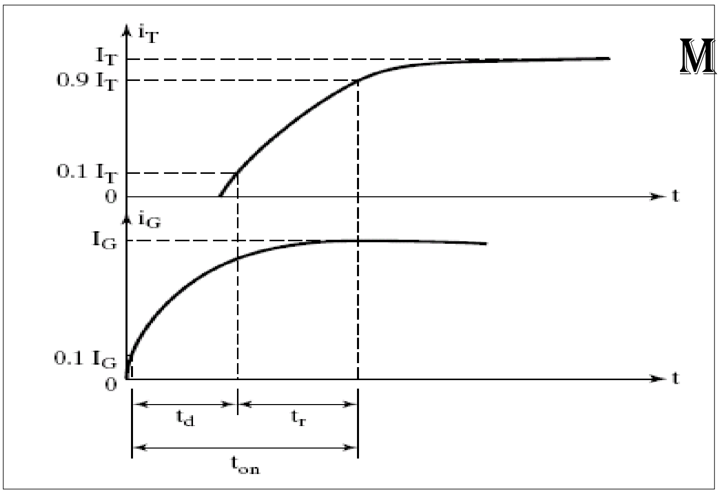
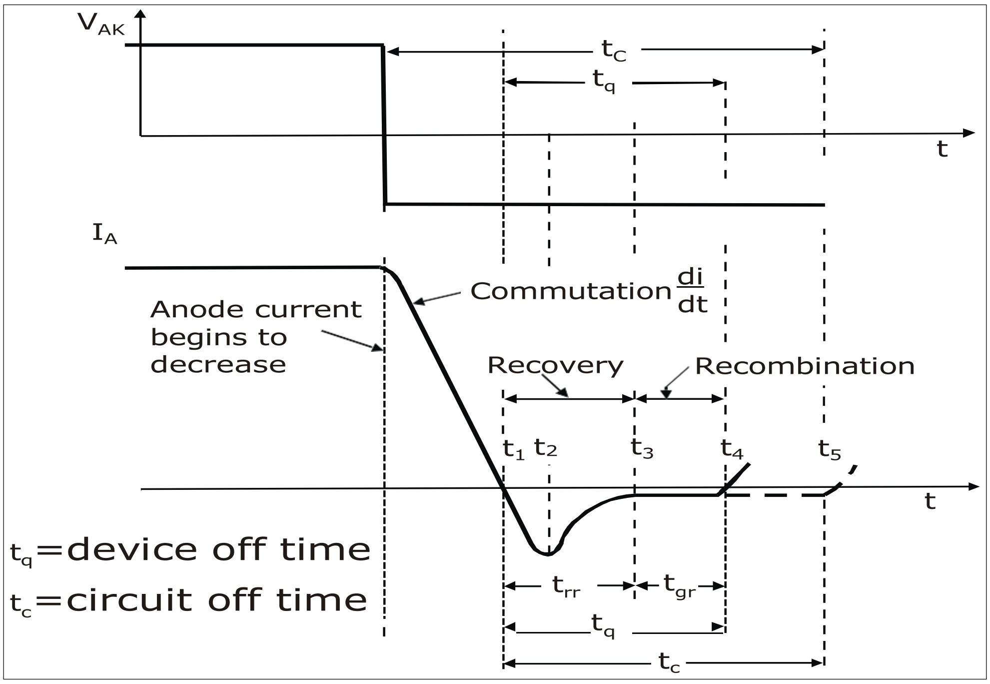


Fig.: Turn-on characteristics

Once t_{gd} has lapsed, the current starts rising towards the peak value. The period during which the anode current rises from 10% to 90% of its peak value is called the rise time. It is also defined as the time for which the anode voltage falls from 90% to 10% of its peak value. The summation of t_{gd} and t_r gives the turn on time t_{on} of the thyristor.

THYRISTOR TURN OFF CHARACTERISTICS



When an SCR is turned on by the gate signal, the gate loses control over the device and the device can be brought back to the blocking state only by reducing the forward current to a level below that of the holding current. In AC circuits, however, the current goes through a natural zero value and the device will automatically switch off. But in DC circuits, where no neutral zero value of current exists, the forward current is reduced by applying a reverse voltage across anode and cathode and thus forcing the current through the SCR to zero.

As in the case of diodes, the SCR has a reverse recovery time t_{rr} which is due to charge storage in the junctions of the SCR. These excess carriers take some time for recombination resulting in the gate recovery time or reverse recombination time t_{gr}. Thus, the turn-off time t_q is the sum of the durations for which reverse recovery current flows after the application of reverse voltage and the time required for the recombination of all excess carriers present. At the end of the turn off time, a depletion layer develops across J_2 and the junction can now withstand the forward voltage. The turn off time is dependent on the anode current, the magnitude of reverse V_g applied ad the magnitude and rate of application of the forward voltage. The turn off time for converte grade SCR's is 50 to 100μsec and that for inverter grade SCR's is 10 to 20μsec.

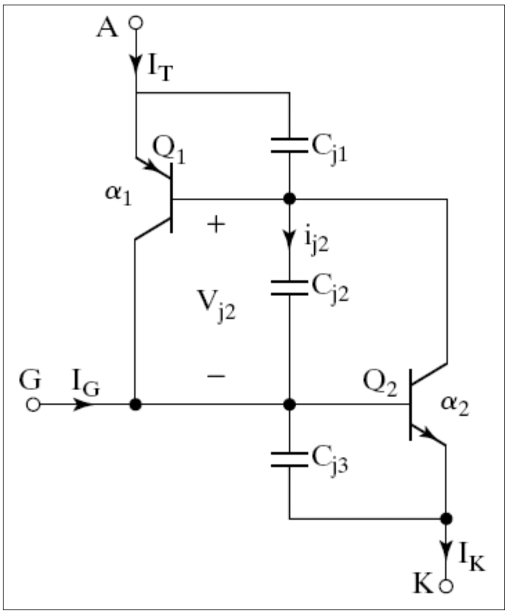
To ensure that SCR has successfully turned off , it is required that the circuit off time t_c be greater than SCR turn off time t_q.

THYRISTOR TURN ON

- **Thermal Turn on:** If the temperature of the thyristor is high, there will be an increase in charge carriers which would increase the leakage current. This would cause an increase in α_1 & α_2 and the thyristor may turn on. This type of turn on many cause thermal run away and is usually avoided.
- **Light**: If light be allowed to fall on the junctions of a thyristor, charge carrier concentration would increase which may turn on the SCR.
- **LASCR:** Light activated SCRs are turned on by allowing light to strike the silicon wafer.
- **High Voltage Triggering:** This is triggering without application of gate voltage with only application of a large voltage across the anode-cathode such that it is greater than the forward breakdown voltage V_{BO}. This type of turn on is destructive and should be avoided.
- **Gate Triggering:** Gate triggering is the method practically employed to turn-on the thyristor. Gate triggering will be discussed in detail later.
- $\frac{dv}{dt}$ **Triggering:** Under transient conditions, the capacitances of the p-n junction will influence the characteristics of a thyristor. If the thyristor is in the blocking state, a rapidly rising voltage applied across the device would cause a high current to flow through the device resulting in turn-on. If i_{j_2} is the current throught the junction j_2 and C_{j_2} is the junction capacitance and V_{j_2} is the voltage across j_2, then

$$i_{j_2} = \frac{dq_2}{dt} = \frac{d}{dt}\left(C_{j_2} V_{j_2}\right) = \frac{C_{j_2} dV_{J2}}{dt} + V_{j_2}\frac{dC_{j_2}}{dt}$$

From the above equation, we see that if $\frac{dv}{dt}$ is large, 1_{j_2} will be large. A high value of charging current may damage the thyristor and the device must be protected against high $\frac{dv}{dt}$. The manufacturers specify the allowable $\frac{dv}{dt}$.



THYRISTOR RATINGS

First Subscript	Second Subscript	Third Subscript
D $\rightarrow$ off state	W $\rightarrow$ working	M $\rightarrow$ Peak Value
T $\rightarrow$ ON state	R $\rightarrow$ Repetitive	
F $\rightarrow$ Forward	S $\rightarrow$Surge or non-repetitive	
R $\rightarrow$ Reverse		

VOLTAGE RATINGS

V_{DWM} : This specifies the peak off state working forward voltage of the device. This specifies the maximum forward off state voltage which the thyristor can withstand during its working.

V_{DRM} : This is the peak repetitive off state forward voltage that the thyristor can block repeatedly in the forward direction (transient).

V_{DSM} : This is the peak off state surge / non-repetitive forward voltage that will occur across the thyristor.

V_{RWM} : This the peak reverse working voltage that the thyristor can withstand in the reverse direction.

V_{RRM} : It is the peak repetitive reverse voltage. It is defined as the maximum permissible instantaneous value of repetitive applied reverse voltage that the thyristor can block inreverse direction.

V_{RSM}: Peak surge reverse voltage. This rating occurs for transient conditions for a specified time duration.

V_T: On state voltage drop and is dependent on junction temperature.

V_{TM}: Peak on state voltage. This is specified for a particular anode current and junction temperature.

$\frac{dv}{dt}$ rating: This is the maximum rate of rise of anode voltage that the SCR has to withstand and which will not trigger the device without gate signal (refer $\frac{dv}{dt}$ triggering).

CURRENT RATING

$I_{Taverage}$: This is the on state average current which is specified at a particular temperature.

I_{TRMS}: This is the on-state RMS current.

Latching current, I_L: After the SCR has switched on, there is a minimum current required to sustain conduction. This current is called the latching current. I_L associated with turn on and is usually greater than holding current

Holding current, I_H: After an SCR has been switched to the on state a certain minimum value of anode current is required to maintain the thyristor in this low impedance state. If the anode current is reduced below the critical holding current value, the thyristor cannot maintain the current through it and reverts to its off state usually I_μ is associated with turn off the device.

$\frac{di}{dt}$ rating: This is a non repetitive rate of rise of on-state current. This maximum value of rate of rise of current is which the thyristor can withstand without destruction. When thyristor is switched on, conduction starts at a place near the gate. This small area of conduction spreads rapidly and if rate of rise of anode current $\frac{di}{dt}$ is large compared to the spreading velocity of carriers, local hotspots will be formed near the gate due to high current density. This causes the junction temperature to rise above the safe limit and the SCR may be damaged permanently. The $\frac{di}{dt}$ rating is specified in $A|\mu \sec$.

GATE SPECIFICATIONS

I_{GT}: This is the required gate current to trigger the SCR. This is usually specified as a DC value.

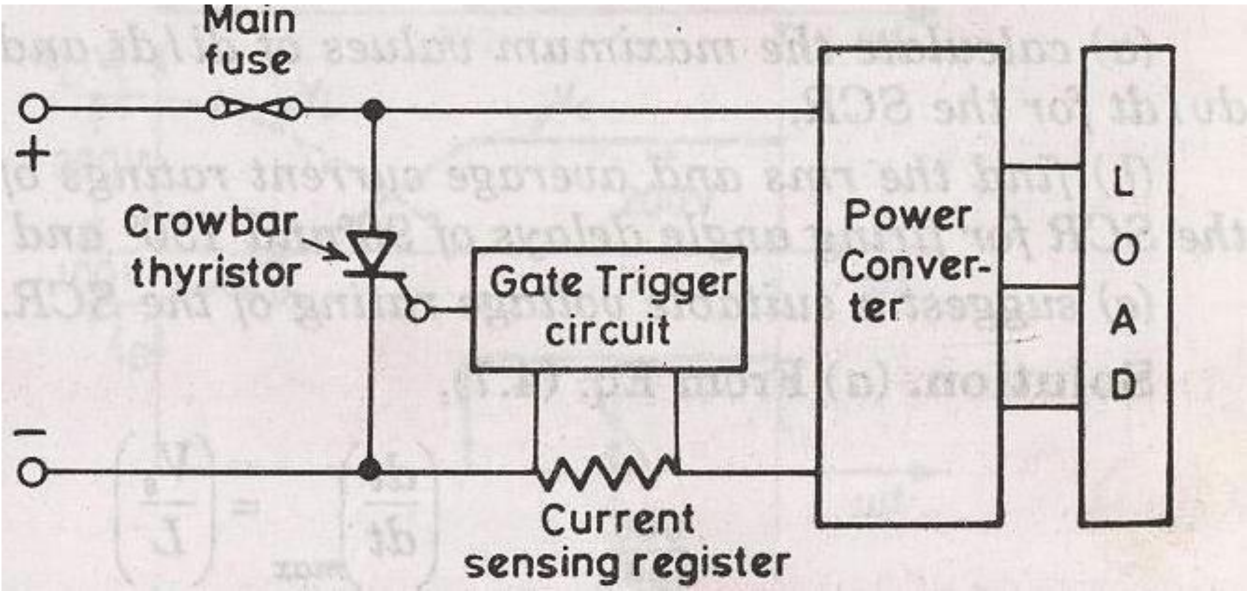
V_{GT}: This is the specified value of gate voltage to turn on the SCR (dc value).

V_{GD}: This is the value of gate voltage, to switch from off state to on state. A value below this will keep the SCR in off state.

Q_{RR}: Amount of charge carriers which have to be recovered during the turn off process.

R_{thjc}: Thermal resistance between junction and outer case of the device.
ELECTRONIC CROWBAR PROTECTION

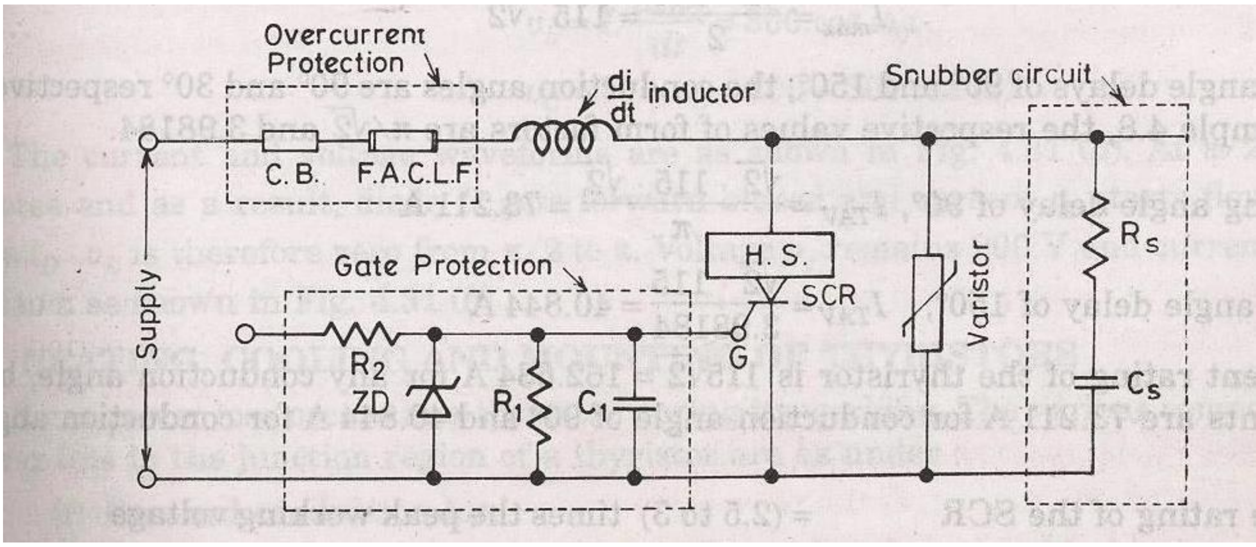
For overcurrent protection of power converter using SCR, electronic crowbar are used. It provide rapid isolation of power converter before any damage occurs.



HEAT PROTECTION-
To protect the SCR

1. From the local spots
2. Temp rise
   SCRs are mounted over heat sinks.

GATE PROTECTION-



Gate circuit should also be protected from

1. Overvoltages
2. Overcurrents

Overvoltage across the gate circuit causes the false triggering of SCR

GATE TRIGGERING METHODS

Types

The different methods of gate triggering are the following

- R-triggering.
- RC triggering.
- UJT triggering.

RESISTANCE TRIGGERING

A simple resistance triggering circuit is as shown. The resistor R_1 limits the current through the gate of the SCR. R_2 is the variable resistance added to the circuit to achieve control over the triggering angle of SCR. Resistor 'R' is a stabilizing resistor. The diode D is required to ensure that no negative voltage reaches the gate of the SCR.

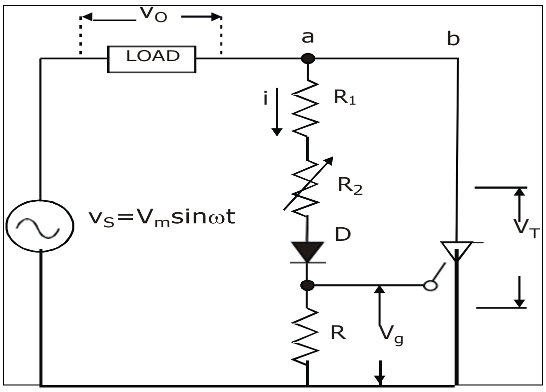


Fig.: Resistance firing circuit

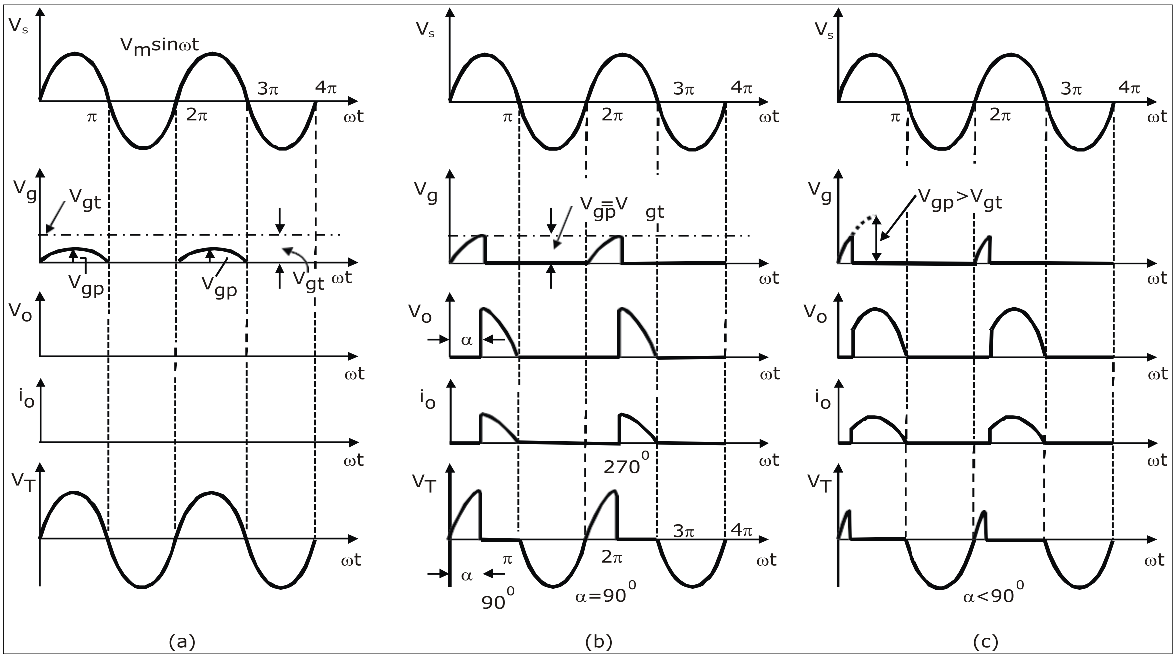


Fig.: Resistance firing of an SCR in half wave circuit with dc load

(a) No triggering of SCR (b) $\alpha = 90^0$ (c) $\alpha < 90^0$

Design

With $R_2 = 0$, we need to ensure that $\frac{V_m}{R_1} < I_{gm}$, where I_{gm} is the maximum or peak gate current of the SCR. Therefore $R_1 \geq \frac{V_m}{I_{gm}}$.

Also with $R_2 = 0$, we need to ensure that the voltage drop across resistor 'R' does not exceed V_{gm} , the maximum gate voltage

$$V_{gm} \geq \frac{V_m R}{R_1 + R}$$

$$\therefore \quad V_{gm}R_1 + V_{gm}R \geq V_m R$$

$$\therefore \quad V_{gm}R_1 \geq R\left(V_m - V_{gm}\right)$$

$$R \leq \frac{V_{gm}R_1}{V_m - V_{gm}}$$

OPERATION

Case 1: $V_{gp} < V_{gt}$

V_{gp}, the peak gate voltage is less then V_{gt} since R_2 is very large. Therefore, current 'I' flowing through the gate is very small. SCR will not turn on and therefore the load voltage is zero and v_{scr} is equal to V_s. This is because we are using only a resistive network. Therefore, output will be in phase with input.

ase 2: $V_{gp} = V_{gt}$, $R_2 \rightarrow$ optimum value.

When R_2 is set to an optimum value such that $V_{gp} = V_{gt}$, we see that the SCR is triggered at 90^0 (since V_{gp} reaches its peak at 90^0 only). The waveforms shows that the load voltage is zero till 90^0 and the voltage across the SCR is the same as input voltage till it is triggered at 90^0.

Case 3: $V_{gp} > V_{gt}$, $R_2 \rightarrow$ small value.

The triggering value V_{gt} is reached much earlier than 90^0. Hence the SCR turns on earlier than V_S reaches its peak value. The waveforms as shown with respect to $V_s = V_m \sin \omega t$.

At $\quad \omega t = \alpha, V_S = V_{gt}, V_m = V_{gp} \left(\because V_{gt} = V_{gp} \sin\alpha \right)$

Therefore $\quad \alpha = \sin^{-1}\left(\dfrac{V_{gt}}{V_{gp}}\right)$

But $\quad V_{gp} = \dfrac{V_m R}{R_1 + R_2 + R}$

Therefore $\quad \alpha = \sin^{-1}\left[\dfrac{V_{gt}\left(R_1 + R_2 + R\right)}{V_m R}\right]$

Since V_{gt}, R_1, R are constants $\alpha \alpha R_2$

RESISTANCE CAPACITANCE TRIGGERING

RC HALF WAVE

Capacitor 'C' in the circuit is connected to shift the phase of the gate voltage. D_1 is used to prevent negative voltage from reaching the gate cathode of SCR.

In the negative half cycle, the capacitor charges to the peak negative voltage of the supply $\left(-V_m\right)$ through the diode D_2. The capacitor maintains this voltage across it, till the supply voltage crosses zero. As the supply becomes positive, the capacitor charges through resistor 'R' from initial voltage of $-V_m$, to a positive value.

When the capacitor voltage is equal to the gate trigger voltage of the SCR, the SCR is fired and the capacitor voltage is clamped to a small positive value.

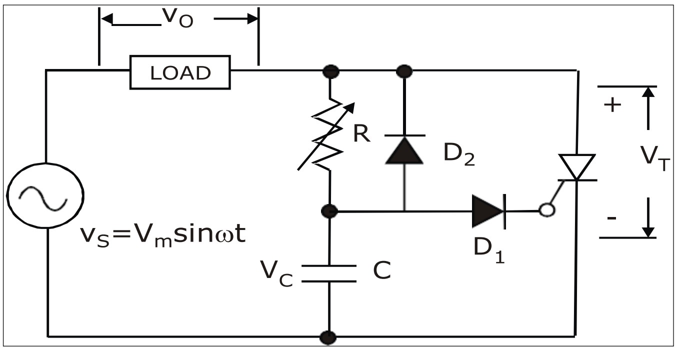


Fig.: RC half-wave trigger circuit

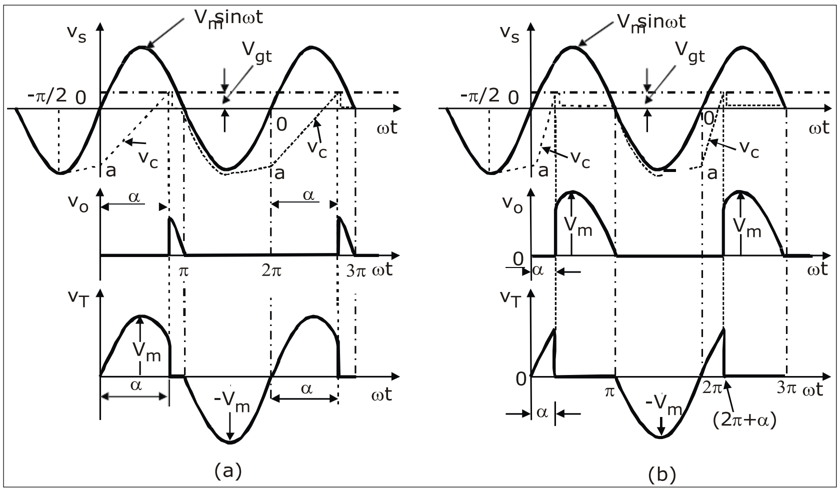


Fig.: Waveforms for RC half-wave trigger circuit

(a) High value of R **(b) Low value of R**

Case 1: R $\rightarrow$ Large.

When the resistor 'R' is large, the time taken for the capacitance to charge from $-V_m$ to V_{gt} is large, resulting in larger firing angle and lower load voltage.

Case 2: R $\rightarrow$ Small

When 'R' is set to a smaller value, the capacitor charges at a faster rate towards V_{gt} resulting in early triggering of SCR and hence V_L is more. When the SCR triggers, the voltage drop across it falls to 1 – 1.5V. This in turn lowers, the voltage across R & C. Low voltage across the SCR during conduction period keeps the capacitor discharge during the positive half cycle.

DESIGN EQUATION

From the circuit $V_C = V_{gt} + V_{d1}$. Considering the source voltage and the gate circuit, we can

PE

write

$v_s = I_{gt} R + V_C$. SCR fires when

$v_s \geq I_{gt} R + V_C$ that is $v_S \geq I_g R + V_{gt} + V_{d1}$. Therefore $R \leq \frac{v_s - V_{gt} - V_{d1}}{I_{gt}}$. The RC time constant for zero output voltage that is maximum firing angle for power frequencies is empirically gives as $RC \geq 1.3\left(\frac{T}{2}\right)$.

RC FULL WAVE

A simple circuit giving full wave output is shown in figure below. In this circuit the initial voltage from which the capacitor 'C' charges is essentially zero. The capacitor 'C' is reset to this voltage by the clamping action of the thyristor gate. For this reason the charging time constant RC must be chosen longer than for half wave RC circuit in order to delay the triggering. The RC value is empirically chosen as $RC \geq \frac{50T}{2}$. Also $R \leq \frac{v_s - V_{gt}}{I_{gt}}$.

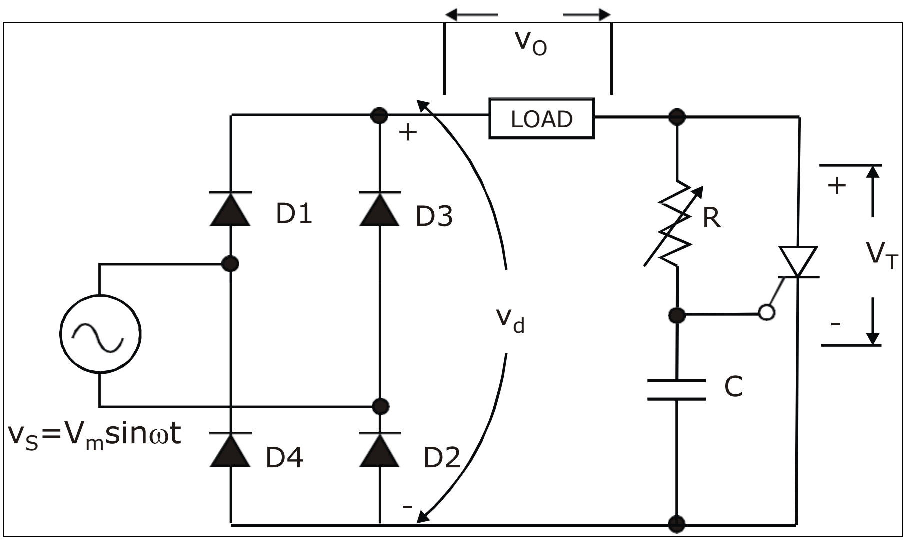


Fig: RC full-wave trigger circuit

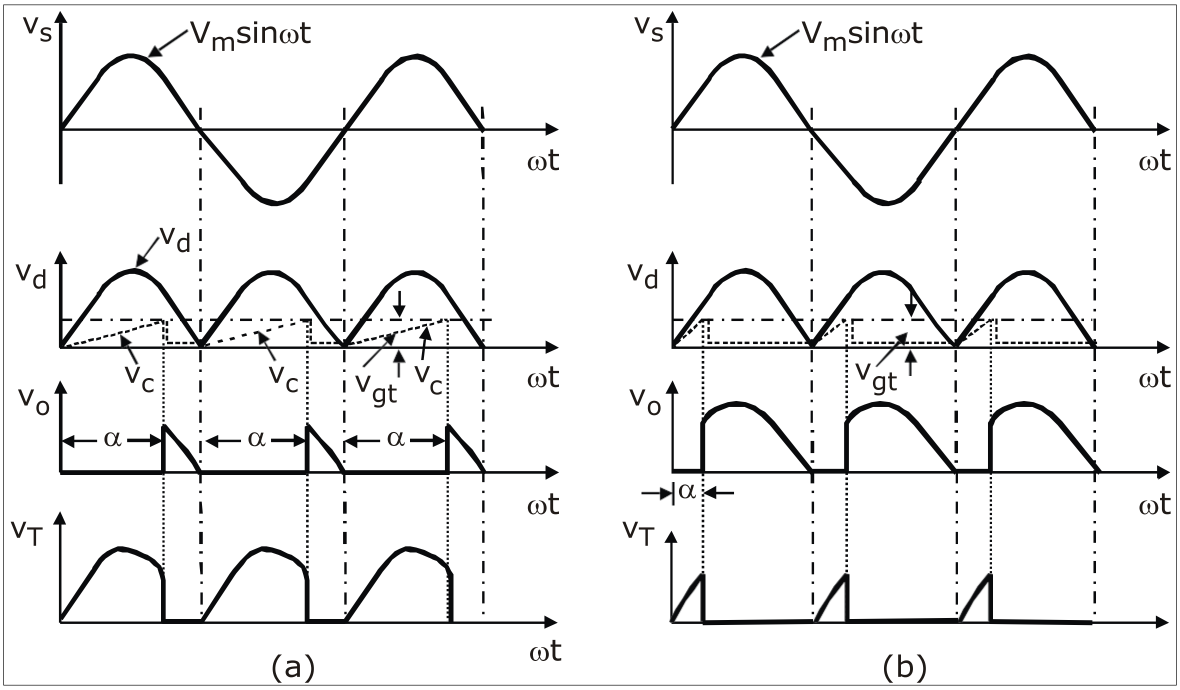


Fig: Wave-forms for RC full-wave trigger circuit

(a) High value of R **(b) Low value of R**

PROBLEM

1. Design a suitable RC triggering circuit for a thyristorised network operation on a 220V, 50Hz supply. The specifications of SCR are $V_{gt\,\min} = 5V$, $I_{gt\,\max} = 30mA$.

$$R = \frac{v_s - V_{gt} - V_D}{I_g} = 7143.3\Omega$$

Therefore $RC \geq 0.013$

$R \leq 7.143k\Omega$

$C \geq 1.8199\mu F$

UNI-JUNCTION TRANSISTOR (UJT)

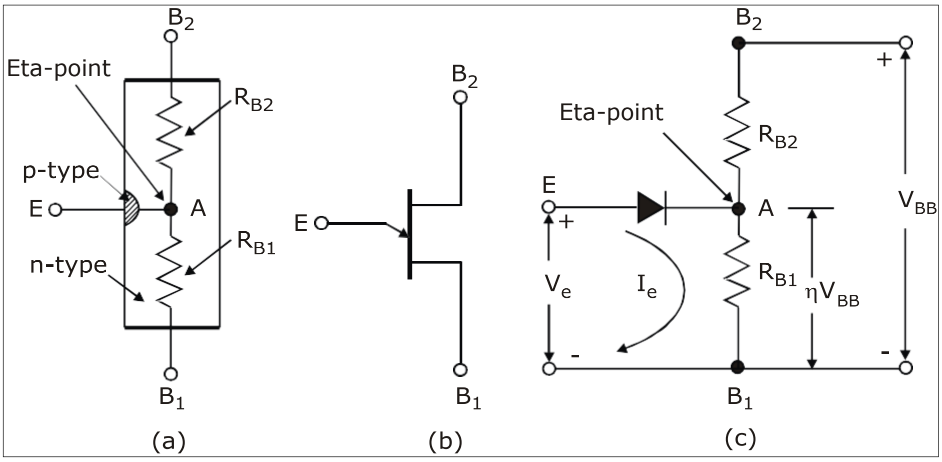


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Fig.: (a) Basic structure of UJT (b) Symbolic representation (c) Equivalent circuit

UJT is an n-type silicon bar in which p-type emitter is embedded. It has three terminals base1, base2 and emitter 'E'. Between B_1 and B_2 UJT behaves like ordinary resistor and the internal resistances are given as R_{B1} and R_{B2} with emitter open $R_{BB} = R_{B1} + R_{B2}$. Usually the p-region is heavily doped and n-region is lightly doped. The equivalent circuit of UJT is as shown. When V_{BB} is applied across B_1 and B_2, we find that potential at A is

$$V_{AB1} = \frac{V_{BB}R_{B1}}{R_{B1} + R_{B2}} = \eta V_{BB} \left[\eta = \frac{R_{B1}}{R_{B1} + R_{B2}} \right]$$

η is intrinsic stand off ratio of UJT and ranges between 0.51 and 0.82. Resistor R_{B2} is between 5 to 10KΩ.

OPERATION

When voltage V_{BB} is applied between emitter 'E' with base 1 B_1 as reference and the emitter voltage V_E is less than $(V_D + \eta V_{BE})$ the UJT does not conduct. $(V_D + \eta V_{BB}$ is designated as V_P)

which is the value of voltage required to turn on the UJT. Once V_E is equal to $V_P \equiv \eta V_{BE} + V_D$, then UJT is forward biased and it conducts.

The peak point is the point at which peak current I_P flows and the peak voltage V_P is across the UJT. After peak point the current increases but voltage across device drops, this is due to the fact that emitter starts to inject holes into the lower doped n-region. Since p-region is heavily doped compared to n-region. Also holes have a longer life time, therefore number of carriers in the base region increases rapidly. Thus potential at 'A' falls but current I_E increases rapidly. R_{B1} acts as a decreasing resistance.

The negative resistance region of UJT is between peak point and valley point. After valley point, the device acts as a normal diode since the base region is saturated and R_{B1} does not decrease again.

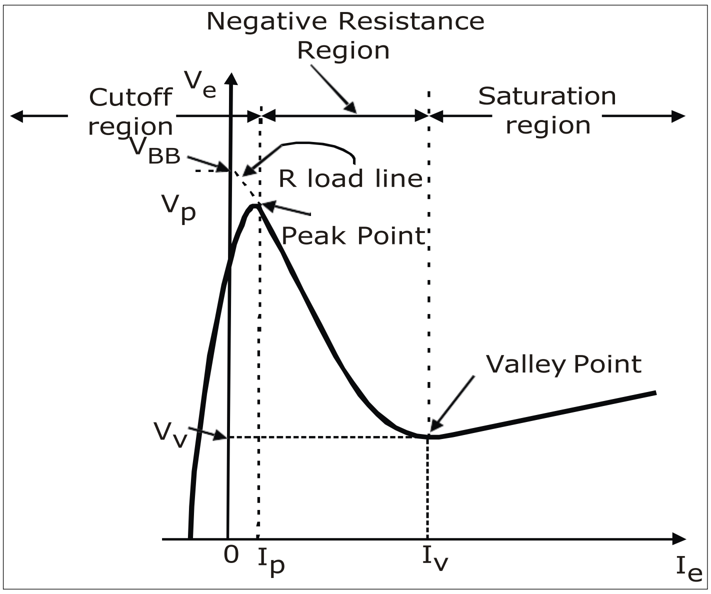


Fig.: V-I Characteristics of UJT

UJT RELAXATION OSCILLATOR

UJT is highly efficient switch. The switching times is in the range of nanoseconds. Since UJT exhibits negative resistance characteristics it can be used as relaxation oscillator. The circuit diagram is as shown with R_1 and R_2 being small compared to R_{B1} and R_{B2} of UJT.

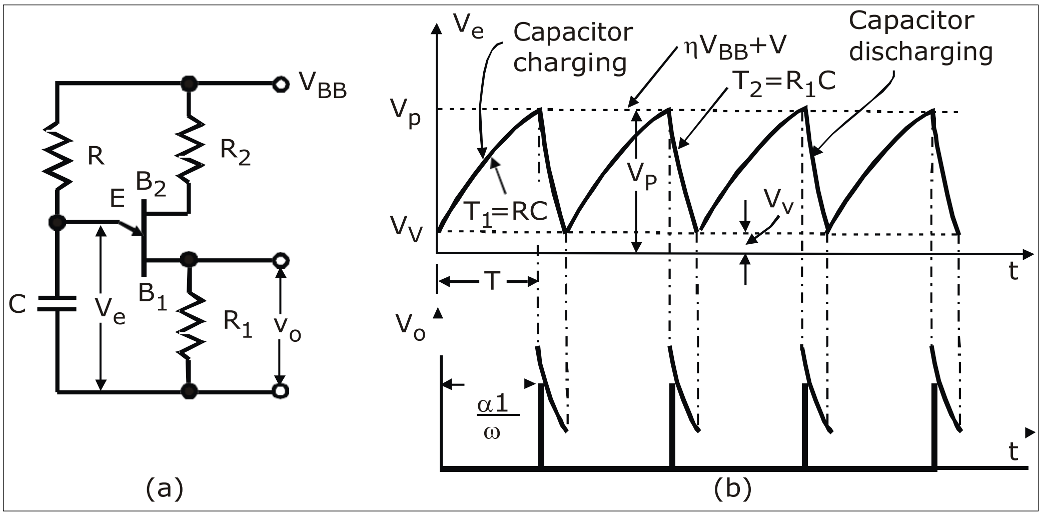


Fig.: UJT oscillator (a) Connection diagram and (b) Voltage waveforms

OPERATION

When V_{BB} is applied, capacitor 'C' begins to charge through resistor 'R' exponentially towards V_{BB}. During this charging emitter circuit of UJT is an open circuit. The rate of charging is $\tau_1 = RC$. When this capacitor voltage which is nothing but emittervoltage V_E reaches the peak point $V_P = \eta V_{BB} + V_D$, the emitter base junction is forward biased and UJT turns on. Capacitor

PE

PE

'C' rapidly discharges through load resistance

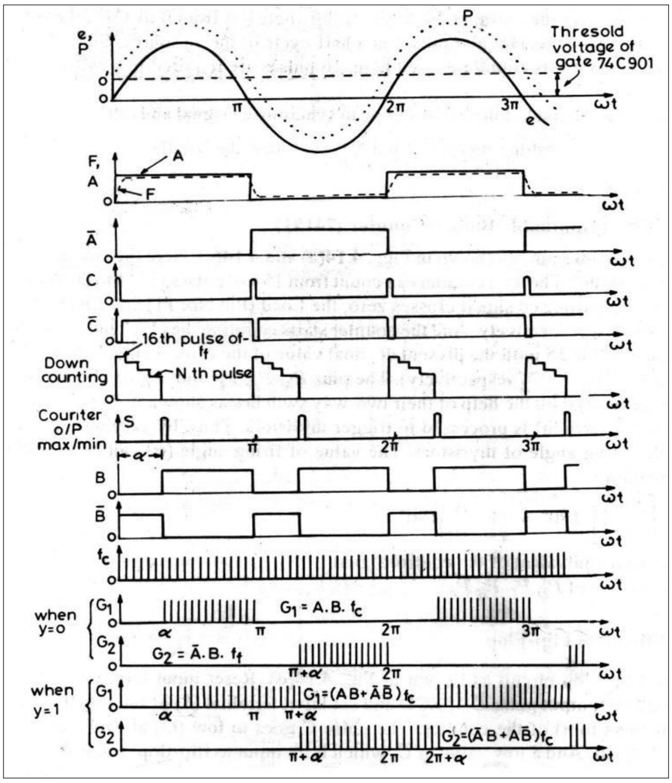
PE

R_1 with time constant $\tau_2 = R_1 C (\tau_2$

τ_1) . When
emitter voltage decreases to valley point V_v, UJT turns off. Once again the capacitor will charge towards V_{BB} and the cycle continues. The rate of charging of the capacitor will be determined by the resistor R in the circuit. If R is small the capacitor charges faster towards V_{BB} and thus reaches V_P faster and the SCR is triggered at a smaller firing angle. If R is large the capacitor takes a longer time to charge towards V_P the firing angle is delayed. The waveform for both cases is as shown below.

EXPRESSION FOR PERIOD OF OSCILLATION 'T'

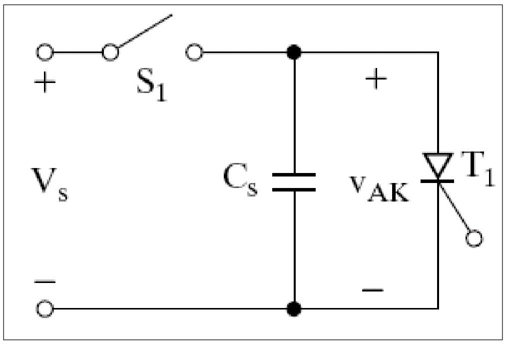
The period of oscillation of the UJT can be derived based on the voltage across the capacitor. Here we assume that the period of charging of the capacitor is lot larger than than the discharging time.

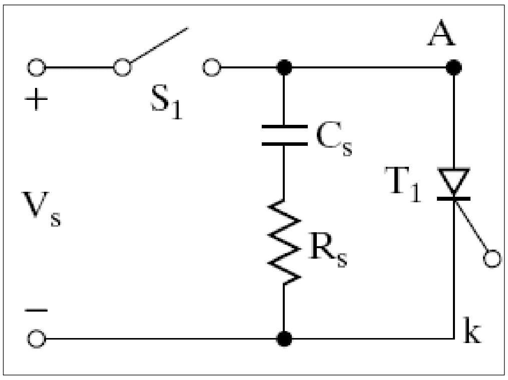


$\frac{dv}{dt}$ PROTECTION

The $\frac{dv}{dt}$ across the thyristor is limited by using snubber circuit as shown in figure (a) below. If switch S_1 is closed at $t = 0$, the rate of rise of voltage across the thyristor is limited by the capacitor C_S. When thyristor T_1 is turned on, the discharge current of the capacitor is limited by the resistor R_S as shown in figure (b) below.

Fig. (a)





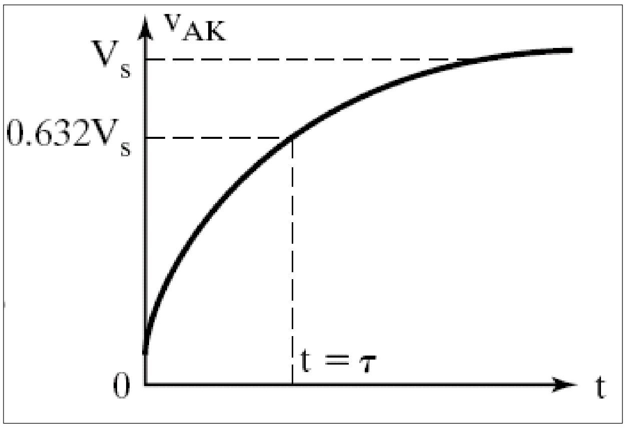


Fig. (c)

The voltage across the thyristor will rise exponentially as shown by fig (c) above.

From fig. (b) above, circuit we have (for SCR off)

$$V_S = i(t)\,R_S + \frac{1}{C}\int i(t)\,dt + V_c(0) \quad [\textit{for}\ t=0]$$

Therefore $$i(t) = \frac{V_S}{R_S} e^{-t/\tau_s}, \text{ where } \tau_s = R_S C_S$$

Also $$V_T(t) = V_S - i(t) R_S$$

$$V_T(t) = V_S - \frac{V_S}{R_S} e^{-t/\tau_s} R_S$$

Therefore $$V_T(t) = V_S - V_S e^{-t/\tau_s} = V_S\left[1 - e^{-t/\tau_s}\right]$$

At t = 0, $$V_T(0) = 0$$

At $t = \tau_s$, $$V_T(\tau_s) = 0.632V_S$$

Therefore $$\frac{dv}{dt} = \frac{V_T(\tau_s) - V_T(0)}{\tau_s} = \frac{0.632V_S}{R_S C_S}$$

And $R_S = \frac{V_S}{I_{TD}}$.

I_{TD} is the discharge current of the capacitor.

It is possible to use more than one resistor for $\frac{dv}{dt}$ and discharging as shown in the figure (d) below. The $\frac{dv}{dt}$ is limited by R_1 and C_S. $R_1 + R_2$ limits the discharging current such that $I_{TD} = \frac{V_S}{R_1 + R_2}$

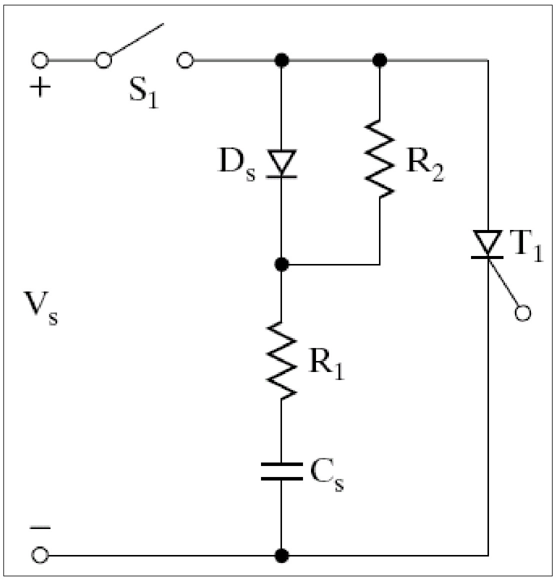


Fig. (d)

The load can form a series circuit with the snubber network as shown in figure (e) below. The damping ratio of this second order system consisting RLC network is given as,

$$\delta = \frac{\alpha}{\omega_0} = \frac{R_S + R}{2}\sqrt{\frac{C_S}{L_S + L}}$$

, where L_S is stray inductance and L, R is load inductance and resistance respectively.

To limit the peak overshoot applied across the thyristor, the damping ratio should be in the range of 0.5 to 1. If the load inductance is high, R_S can be high and C_S can be small to retain the desired value of damping ratio. A high value of R_S will reduce discharge current and a low value of C_S reduces snubber loss. The damping ratio is calculated for a particular circuit R_S and C_S can be found.

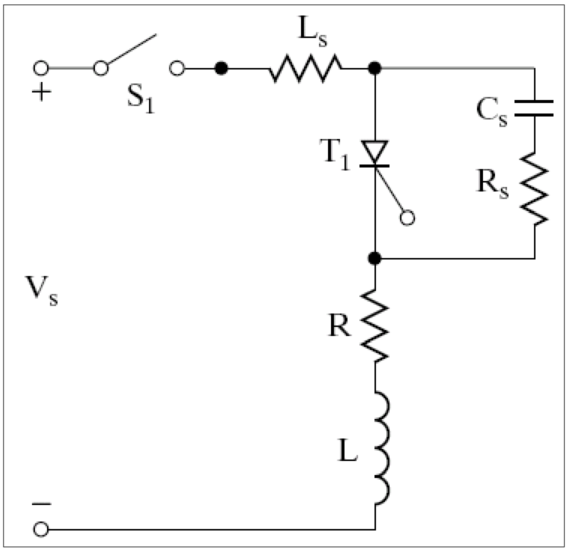
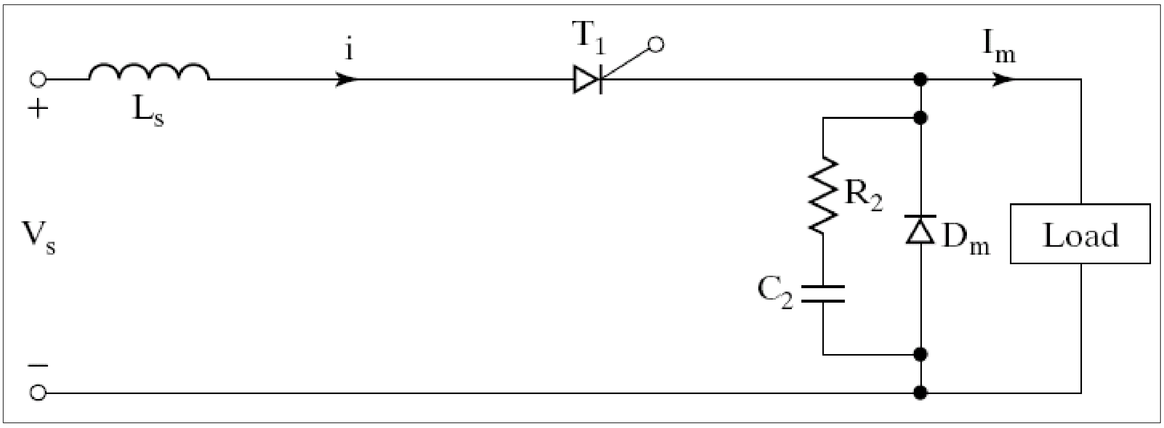


Fig. (e)

$\frac{di}{dt}$ PROTECTION



Practical devices must be protected against high $\frac{di}{dt}$. As an example let us consider the circuit shown above, under steady state operation D_m conducts when thyristor T_1 is off. If T_1 is fired when D_m is still conducting $\frac{di}{dt}$ can be very high and limited only by the stray inductance of the circuit. In practice the $\frac{di}{dt}$ is limited by adding a series inductor L_S as shown in the circuit above. Then the forward $\frac{di}{dt} = \frac{V_S}{L_S}$.

SERIES AND PARALLEL OPERATION

SCR ratings have improved considerably since its introduction in 1957. Presently, SCRs with voltage and current rating of 10kV and 3kA are available. However, for some industrial applications, the demand for voltage and current ratings is so high that a single SCR cannot fulfill such requirements. In such cases, SCRs are connected in series in order to meet the high voltage demand and in parallel for fulfilling high current demand.

The string efficiency that is a term used for measuring the degree of utilization of SCRs in a string.

$$\text{String efficiency} = \frac{\text{Actual voltage / current rating}}{(n_s\text{ , no. of SCRs}) \times \text{voltage / current rating of one SCR}}$$

Usually the above ratio is less than one. Since SCRs of same ratings and specifications do not have identical characteristics unequal voltage / current sharing is bound to occur for all SCRs in a string. Therefore the string efficiency can never be equal to one.

DERATING FACTOR (DRF)

The use of an extra unit will improve the reliability of a string. A measure of the reliability of the string is given by a factor called derating factor defined as

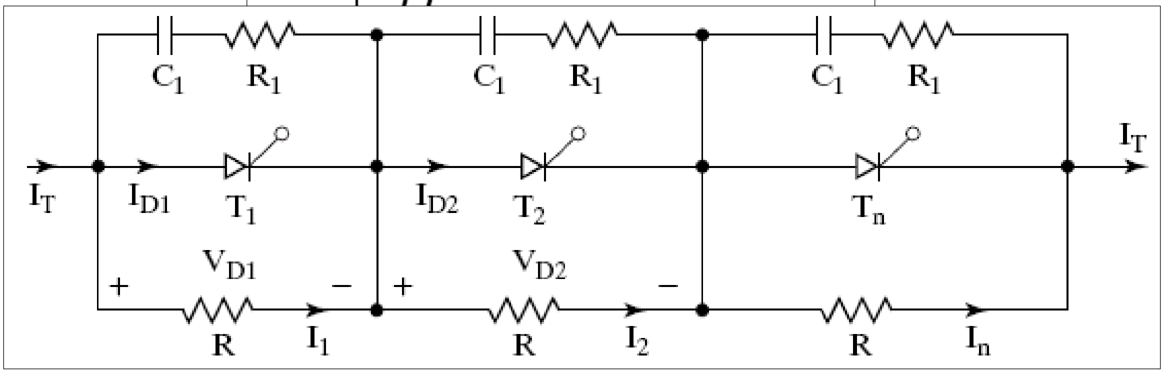
DRF = 1 - String efficiency

SERIES OPERATION OF SCRS

For high voltage applications two or more Thyristors can be connected in series to provide the required voltage rating. However due to production spread the characteristics of Thyristors of the same type are not identical.

STATIC EQUALIZATION

As seen from V-I characteristics, two identical Thyristors to be used in a string do not have the same off state current for same off-state voltages. If these SCRs are used in a string as such, unequal voltage distribution would occur. In order to overcome this, we could connect resistors across individual SCRs to meet the requirement of equal off state currents for the same off state voltage. But this is not practical therefore we use the same resistor ‘R’ across each SCR to get fairly uniform voltage distribution.



We see that, equal resistors ‘R’ are connected across individual SCR’s which are connected in series. Let n_s be the number of SCRs connected. Let I_T be the total current that the string carries and individual SCRs have leakage currents $I_{D1}, I_{D2}..............., I_{Dn}$.

As seen from the V-I characteristics, even though the voltage across each SCR is the same, the leakage current in the off state differ. Let $I_{D1} < I_{D2}$. Since SCR1 has lower leakage current

compared to other SCRs, it will block a higher voltage compared to other SCRs.

Let the leakage current of other SCRs, be such that $I_{D2} = I_{D3} = \square, I_{Dn}$. Therefore

$$I_{D1} = I_T - I_1 .... (1)$$

$$I_{D2} = I_T - I_2 .... (2)$$

If V_{D1} is the voltage across SCR1, then $V_{D1} = I_1 R$, and voltage across the rest of the SCRs is $(n_s - 1) I_2 R$.

Therefore total voltage across the string = $V_s = I_1 R + (n_s - 1) I_2 R$.

But from equation (2) $\quad V_S = I_1 R + (n_s - 1)(I_T - I_{D2})$

But from equation (1) $\quad I_T = I_{D1} + I_1$

Therefore

PE $V_S = I_1R + (n_s - 1)[I_{D1} + I_1 - I_{D2}]R$

$$V_S = V_{D1} + (n_s - 1)\ I_1R + (n_s - 1)[I_{D1} - I_{D2}]R$$

But $I_{D1} < I_{D2}$, $\therefore$ $V_S = V_{D1} + (n_s - 1)\, I_1 R - (n_s - 1)(I_{D2} - I_{D1})\, R$

$I_{D2} - I_{D1} = \Delta I_D$ = difference between leakage currents of SCR1 and the rest of the SCRs.

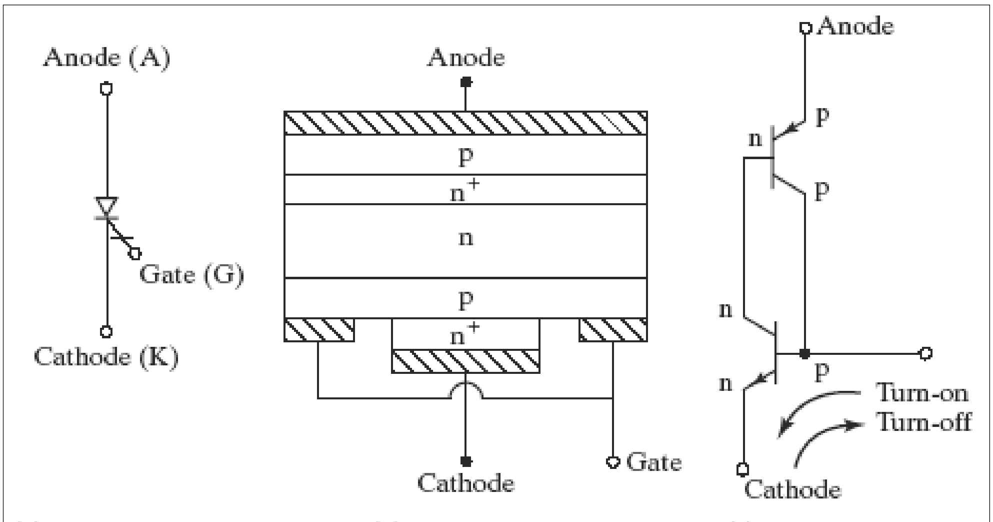
Therefore $V_S = V_{D1} + (n_s - 1)V_{D1} - (n_S - 1)\Delta I_D R$

$$V_S = n_s V_{D1} - (n_s - 1)\Delta I_D R \qquad ...(3)$$

Also $R = \dfrac{n_s V_{D1} - V_S}{(n_s - 1)\,\Delta I_D}$

From equation (3), considering the worst case condition of $I_{D1} = 0$, $V_{D1(\max)} = \dfrac{V_S + (n_s - 1)\, R I_{D2}}{n_s}$

GATE TURN-OFF THYRISTORS



(a) GTO symbol (b) Cross section (c) Equivalent circuit

A gate-turn-off thyristor (GTO) like an SCR can be turned on by applying a positive gate signal. However, it can be turned off by a negative gate signal. A GTO is a latching device and can be built with current and voltage ratings similar to those of an SCR. A GTO is turned on by applying a short positive pulse and turned off by a short negative pulse to its gate. The GTOs have advantages over SCRs.

Elimination of commutating components in forced commutation, resulting in reduction in cost, weight, and volume.

Reduction in acoustic and electro-magnetic noise due to the elimination of commutation chokes.

Faster turn-off permitting high switching frequencies and

Improved efficiency of converters.

In low power applications GTOs have the following advantages over bipolar transistors.

A higher blocking voltage capability.

A high ratio of peak controllable current to average current.

A high ratio of surge peak current to average current, typically 10:1.

A high on-state gain (anode current/gate current), typically 600; and

A pulsed gate signal of short duration.

Under surge conditions, a GTO goes into deeper saturation due to regenerative action. On the other hand, a bipolar transistor tends to come out of saturation.

A GTO has low gain during turn-off, typically 6, and requires a relatively high negative current pulse to turn off. It has higher on-state voltage than that of SCRs. The on-state voltage of typical 550A, 1200V GTO is typically 3.4V.

Controllable peak on-state current I_{TGQ} is the peak value of on-state current which can be turned off by gate control. The off state voltage is reapplied immediately after turn-off and the reapplied dv/dt is only limited by the snubber capacitance. Once a GTO is turned off, the load current I_L, which is diverted through and charges the snubber capacitor, determines the reapplied $dv|dt$.

$$\frac{dv}{dt} = \frac{I_L}{C_s}$$

Where C_s is the snubber capacitance

BIDIRECTIONAL TRIODE THYRISTORS

A TRIAC conducts in both directions unlike the SCR. Since it conducts in both directions, the terminals are named as MT1 and MT2 and the Gate. As seen from the diagram the gate 'G' is near terminal MT1. The cross hatched strip shows that 'G' is connected to n_3 as well as p_2.

Similarly terminal MT1 is connected to p_2 as well as n_2 and terminal MT2 is connected to join p_1 and n_4.

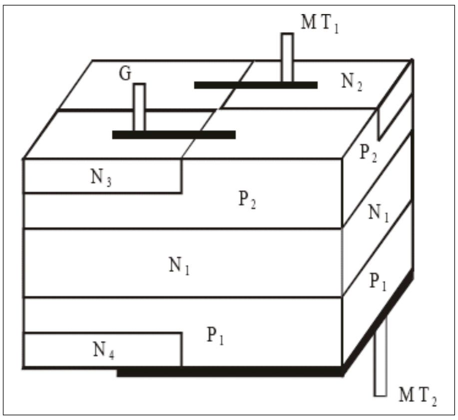


Fig.: Triac Structure

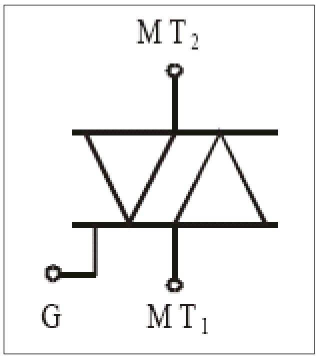


Fig.: Triac Symbol

With no signal to the gate the triac will block both half cycles of the applied AC voltage in case the peak value of this voltage is less than the breakover voltage in either direction. However the triac can be turned on by applying a positive voltage with respect to terminal MT1. For convenience sake MT1 is taken as the reference terminal.

There are four modes of operation of the triac.

OPERATION

MODE (I): MT2 POSITIVE, GATE POSITIVE

When gate current is positive with respect to MT1, gate current mainly flows through $p_2 n_2$ junction like in ordinary SCR. When the gate current has injected sufficient charge into the p_2 layer the traic starts conducting through $p_1 n_1 p_2 n_2$ layers. This shows that when MT2 and gate are positive with respect to MT1 triac acts like a conventional thyristor. The quadrant of operation is the first quadrant

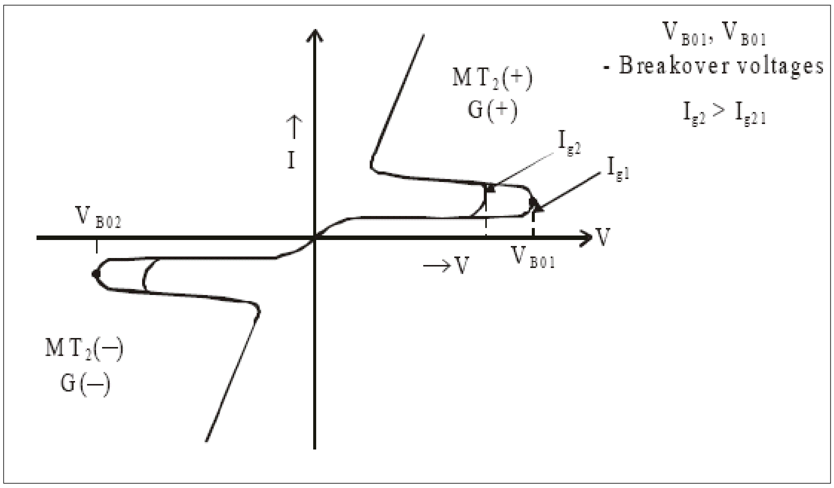


Fig.: V-I

Characteristics of TRIAC

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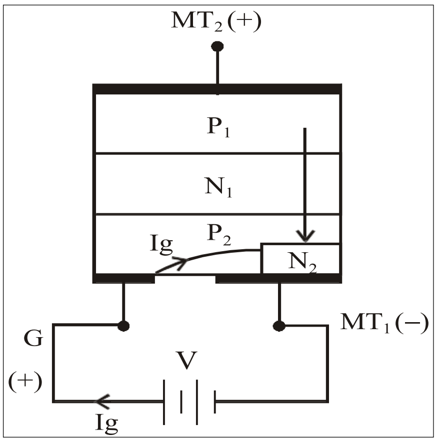
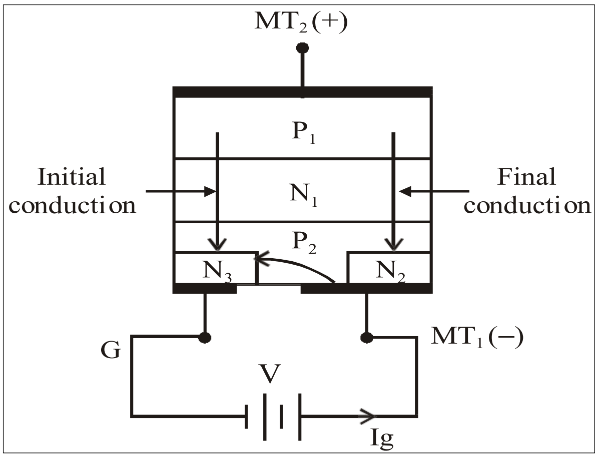


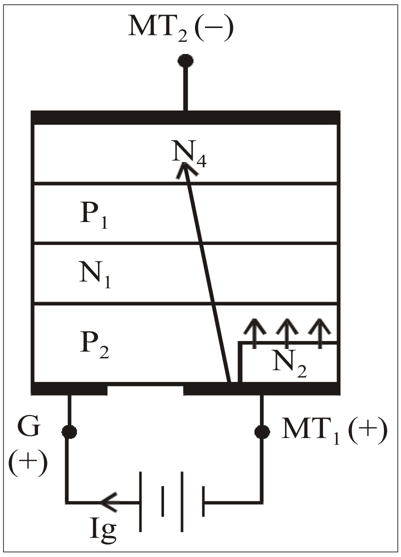
MODE (II): MT2 POSITIVE, GATE NEGATIVE

When gate terminal is negative with respect to MT1 gate current flows through $p_2 n_3$ junction and forward biases this junction. As a result the triac starts conducting through $p_1 n_1 p_2 n_3$ initially. With the conduction of $p_1 n_1 p_2 n_3$ the voltage drop across this path falls but the potential of layer between $p_2 n_3$ rises towards the anode potential of MT2. As the right hand portion of p_2 is clamped at cathode potential of MT1 a potential gradient exists across layer p_2. Its left hand side being at a higher potential than its right hand side a current is thus established in layer p_2 from left to right which forward bias the $p_2 n_2$ junction and finally the main structure $p_1 n_1 p_2 n_2$ begins to conduct. The structure of $p_1 n_1 p_2 n_3$ may be regarded as an auxiliary SCR and the structure $p_1 n_1 p_2 n_2$ as the main SCR. It can be stated that the anode current of the auxiliary SCR serves as the gate current of the main SCR. This mode of operation is less sensitive as compared to the previous mode since more gate current is required.



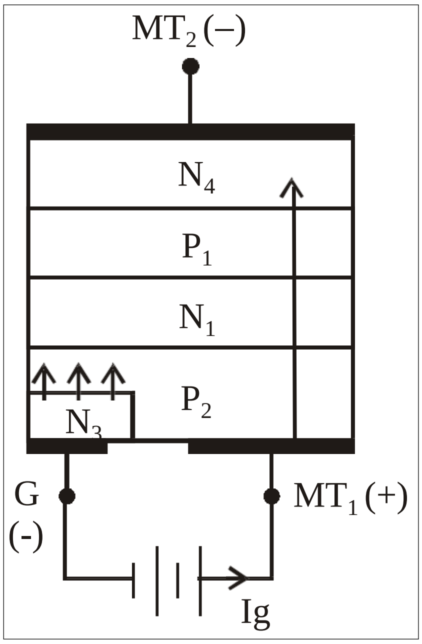
MODE (III): MT2 NEGATIVE, GATE POSITIVE

The gate current I_G forward bias $p_2 n_2$ junction. Layer n_2 injects negative electrons $(e^{-}s)$ into the p_2 layer as shown . With n_2 layer acting as a remote gate the structure $p_2 n_1 p_1 n_4$ eventually turns on. As usual the current after conduction is limited by the external load. Since in this mode the triac is turned on by a remote gate n_2, the device is less sensitive in the III quadrant with positive gate current.



MODE (IV): MT2 NEGATIVE, GATE NEGATIVE

In this mode the layer n_3 acts as a remote gate. The gate current forward bias the $n_3 p_2$ junction. Finally the structure $p_2 n_1 p_1 n_4$ is turned on. Though the triac is turned on by a remote gate n_3 yet the device is more sensitive in this mode.



CONCLUSION

We can conclude that the sensitivity of the triac is greatest in first quadrant and third quadrant. Thus the triac is rarely operated in the first quadrant with negative gate current and in the third quadrant with positive gate current.

As the two conducting paths from MT1 to MT2 or from MT2 to MT1 interact with each other in the structure of the traic, voltage, current and frequency ratings are much lower as compared to conventional Thyristors. The maximum ratings are around 1200V, 300A.

APPLICATIONS

Triacs are used in heat control and for speed controls of small single phase series and induction motors.

PE

UNIT-II

Phase control technique – Single phase Line commutated converters

Unlike diode rectifiers, PCRs or phase controlled rectifiers has an advantage of regulating the output voltage. The diode rectifiers are termed as uncontrolled rectifiers. When these diodes are switched with Thyristors, then it becomes phase control rectifier. The o/p voltage can be regulated by changing the firing angle of the Thyristors. The main application of these rectifiers is involved in speed control of DC motor.

What is a Phase Controlled Rectifier?

The term PCR or Phase controlled rectifier is a one type of rectifier circuit in which the diodes are switched by Thyristors or SCRs (Silicon Controlled Rectifiers). Whereas the diodes offer no control over the o/p voltage, the Thyristors can be used to differ the output voltage by adjusting the firing angle or delay. A phase control Thyristor is activated by applying a short pulse to its gate terminal and it is deactivated due to line communication or natural. In case of heavy inductive load, it is deactivated by firing another Thyristor of the rectifier during the negative half cycle of i/p voltage.

Types of Phase Controlled Rectifier

The phase controlled rectifier is classified into two types based on the type of i/p power supply. And each kind includes a semi, full and dual converter.

Figure: 2.1. Classification of rectifiers

Single-phase Controlled Rectifier

PE

This type of rectifier which works from single phase AC i/p power

supply Single Phase Controlled Rectifiers are classified into

different types

Half wave Controlled Rectifier: This type of rectifier uses a single Thyristor device to provide o/p control only in one half cycle of input AC supply, and it offers low DC output.

Full wave Controlled Rectifier: This type of rectifier provides higher DC output

- Full wave controlled rectifier with a center tapped transformer requires two Thyristors.
- Full wave bridge controlled rectifiers do not need a center tapped transformer

Three-phase Controlled Rectifier

This type of rectifier which works from three phase AC i/p power supply

- A semi converter is a one quadrant converter that has one polarity of o/p voltage and current.
- A full converter is a a two quadrants converter that has polarity of o/p voltage can be either +ve or –ve but, the current can have only one polarity that is either +ve or -ve.
- Dual converter works in four quadrants – both o/p voltage and o/p current can have both the polarities.

Operation of Phase Controlled Rectifier

The basic working principle of a PCR circuit is explained using a single phase half wave PCR circuit with a RL load resistive shown in the following circuit.

A single phase half wave Thyristor converter circuit is used to convert AC to DC power conversion. The i/p AC supply is attained from a transformer to offer the required AC supply voltage to the Thyristor converter based on the o/p DC voltage required. In the above circuit, the primary and secondary AC supply voltages are denoted with VP and VS.

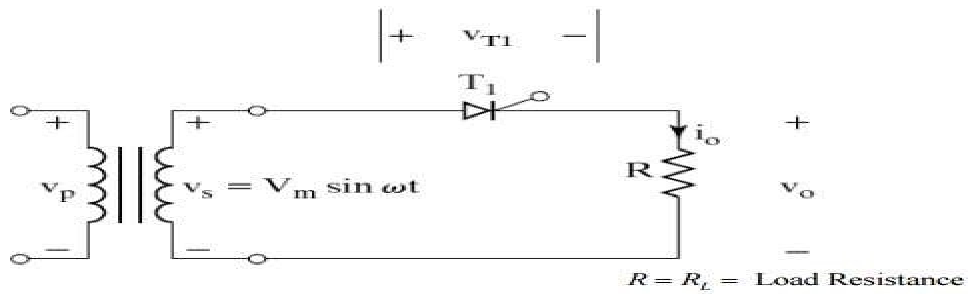


Figure: 2.2. Single phase half wave rectifier circuit

During the +ve half cycle of i/p supply when the upper end of the transformer secondary winding is at a + ve potential with respect to the lower end, the Thyristor is in a forward biased state.

The thyristor is activated at a delay angle of ωt =α, by applying an appropriate gate trigger pulse to the gate terminal of thyristor. When the thyristor is activated at a delay angle of ωt =α, the thyristor behaviors and assuming a perfect thyristor. The thyristor acts as a closed switch and the i/p supply voltage acts across the load when it conducts from ωt =α to π radians For a purely resistive load, the load current io that flows when the thyristor T1 is on, is given by the expression.

$$I_o = v_o / R_L, \text{ for } \alpha \leq \omega t \leq \pi$$

Applications of Phase Controlled Rectifier

Phase controlled rectifier applications include paper mills, textile mills using DC motor drives and DC motor control in steel mills.

- AC fed traction system using a DC traction motor.
- Electro-metallurgical and Electrochemical processes.
- Reactor controls.
- Magnet power supplies.
- Portable hand instrument drives.

- Flexible speed industrial drives.
- Battery charges.
- High voltage DC transmission.
- UPS (Uninterruptible power supply systems).

Operation of half converter with R and RL loads

Single Phase Half Wave Controlled Rectifier with 'R' load:

As shown in figure below primary of transformer is connected to ac mains supply with which SCR becomes forward bias in positive half cycle. T1 is triggered at an angle α, T1 conducts and voltage is applied across R.

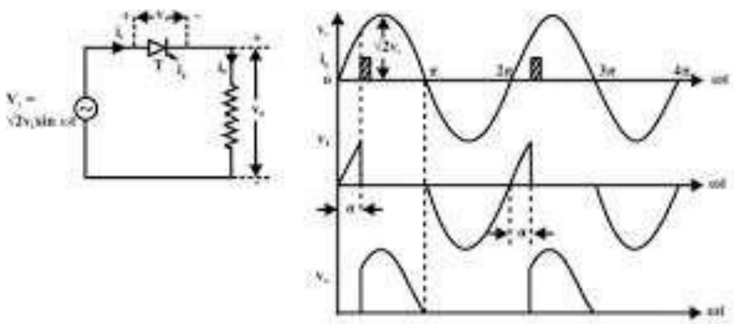


Figure: 2.3 Single phase half wave rectifier with R load with waveforms

The load current i_0 flows through 'R'
the waveforms for voltage & current are as shown
above. As load is resistive,
Output current is given as,

$$I_O = \frac{V_O}{R}$$

Hence shape of output current is same as output voltage
As T1 conducts only in positive half cycle as it is reversed bias in negative cycle, the ripple frequency of output voltage is-
fripple= 50 Hz (supply

frequency) Average output

voltage is given as,

PE

$$V_0(Avg) = \frac{1}{T}\int_0^T V_0(wt)\,dwt$$

i.e Area under one cycle.

Therefore T=2π&Vo(ωt) = Vm sinωt from α to π& for rest of the period Vo(ωt)=0

$$\therefore V_0(Avg) = \frac{1}{2\pi}\int_0^{2\pi} V_m sin(wt)\,dwt$$
$$= \frac{V_m}{2\pi}[-coswt]_\alpha^\pi$$
$$= \frac{V_m}{2\pi}(1+cos\alpha)$$

Power transferred to load,

$$P_0(Avg) = \frac{V_0^2(Avg)}{R}$$

Thus, power & voltage can be controlled by firing angle.

Single Phase Half Wave Controlled Rectifier with 'RL' load:

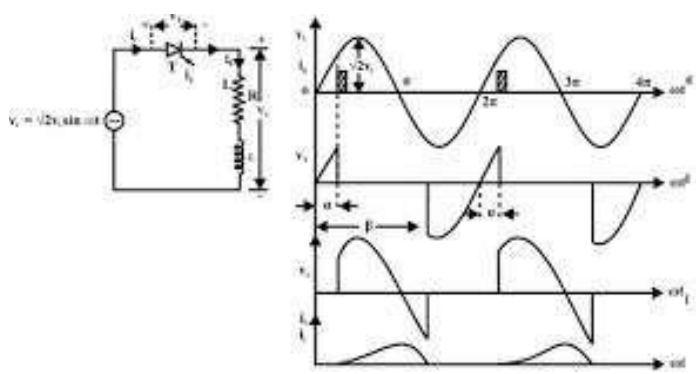


Figure: 2.4 Single phase half wave rectifier with RL load with waveforms

Figure above shows the single phase half wave rectifier with RL Load.

- Normally motors are inductive loads

L= armature of field coil inductance

R= Resistance of coil.

- In positive half cycle, SCR starts conduction at firing angle "α".
- Drop across SCR is small & neglected so output voltage is equal to supply voltage.
- Due to 'R_L' load, current through SCR increases slowly.
- At 'π', supply voltage is at zero where load current is at its max value.
- In positive half cycle, inductor stores energy & that generates the voltage.
- In negative half cycle, the voltage developed across inductor, forward biases SCR & maintains its conduction.
- Basically with the property of inductance it opposes change in current.
- Output current & supply current flows in same loop, so all the time $i_o=i_s$.
- After π the energy of inductor is given to mains & there is flow of 'io'. The energy reduces as if gets consumed by circuit so current also reduces.
- At 'β' energy stored in inductance is finished, hence 'i_o' becomes zero & 'T1' turns off.
- 'i_o' becomes zero from 'β' to '$2\pi+\alpha$' hence it is discontinuous conduction.

$$V_o(Avg) = \frac{1}{T}\int_o^T V_o(wt)\,dwt$$

i.e Area under one cycle.

Therefore $T=2\pi$ & $Vo(\omega t) = Vm\sin\omega t$ from α to π & for rest of the period $Vo(\omega t)=0$

$$\therefore V_o(Avg) = \frac{1}{2\pi}\int_o^{2\pi} V_m \sin(wt)\,dwt$$

$$= \frac{V_m}{2\pi}[-\cos wt]_\alpha^\pi$$

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

Power transferred to load,

$$P_o(Avg) = \frac{V_o^2(Avg)}{R}$$

Thus, power & voltage can be controlled by firing angle.

Single phase half controlled converter with RLE load

The diode D2 and D4 conducts for the positive and negative half cycle of the input voltage waveform respectively. On the other hand T1 starts conduction when it is fired in the positive half cycle of the input voltage waveform and continuous conduction till T3 is fired in the negative half cycle. Fig. shows the circuit diagram and the waveforms of a single phase half controlled converter supplying an R – L – E
load.

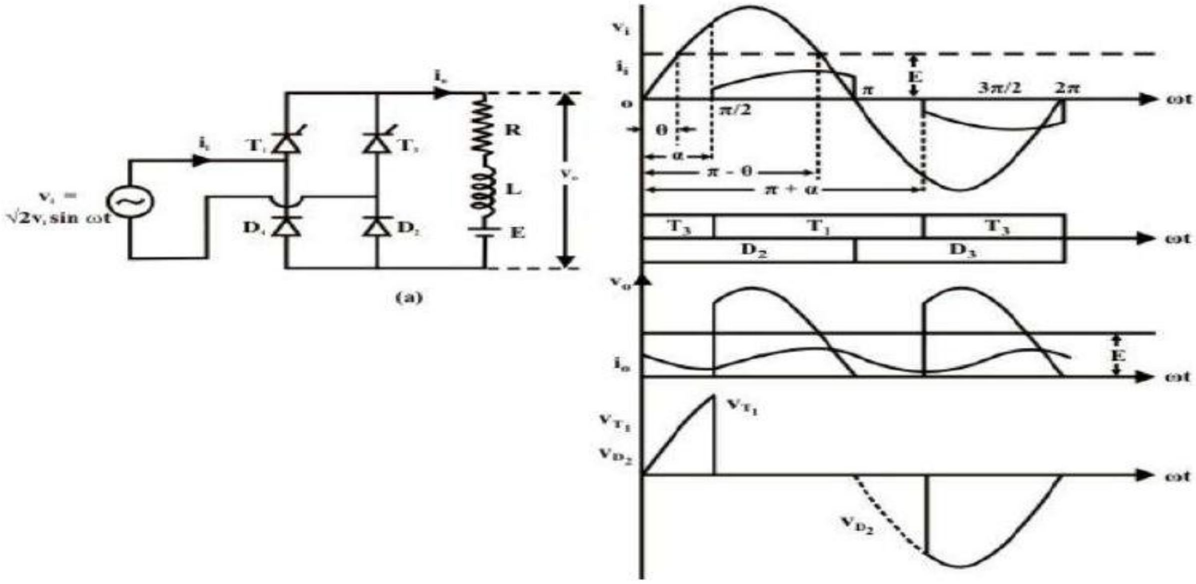


Figure: 2.5 single phase half controlled converter with RLE load

Referring to Fig T1 D2 starts conduction at ωt = α. Output voltage during this period becomes equal to

i. At ωt = π as vi tends to go negative D4 is forward biased and the load current commutates from D2 to D4 and freewheels through D4 and T1. The output voltage remains clamped to zero till T3 is fired at ωt = π + α. The T3 D4 conduction mode continues upto ωt = 2π. Where upon load current again free wheels through T3 and D2 while the load voltage is clamped to zero. From the discussion in the previous paragraph it can be concluded that the output voltage (hence the output current) is periodic over half the input cycle.

Single phase half controlled converter with RLE load and freewheeling diode

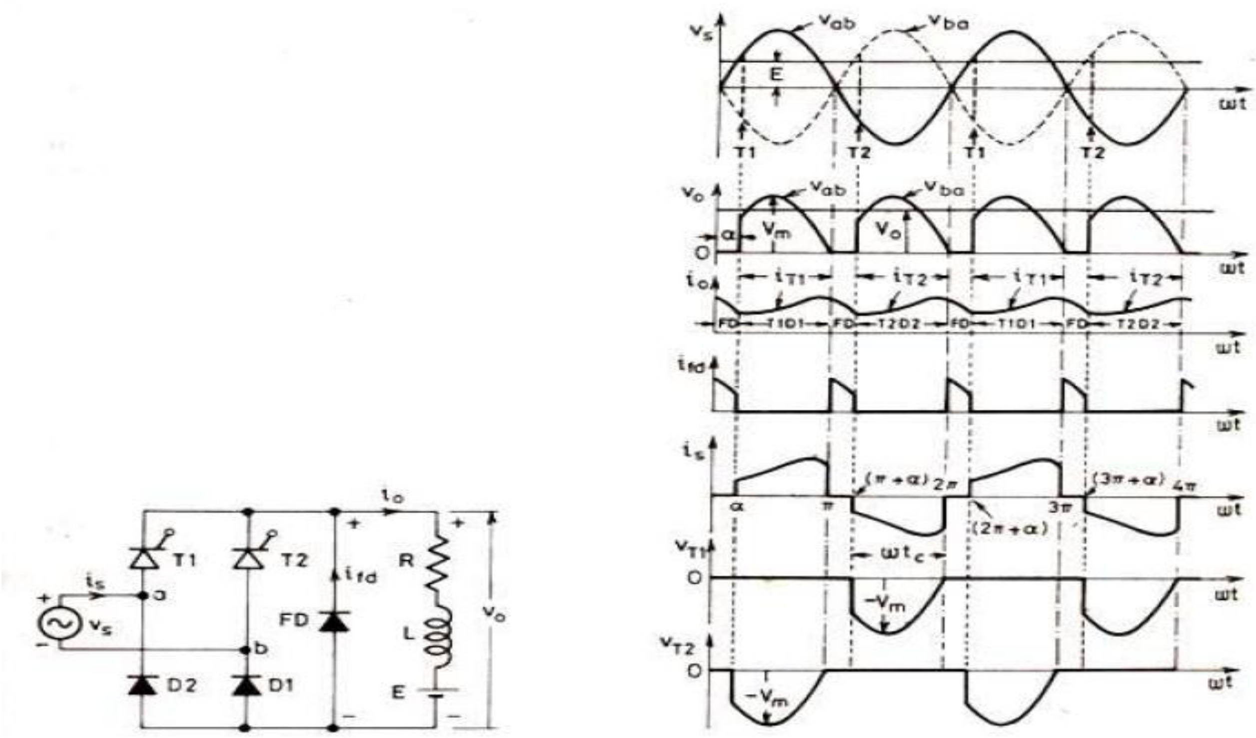


Figure: 2.6 single phase half controlled converter with RLE load and freewheeling diode

Single phase full wave controlled rectifier

Single Phase Full Wave Controlled Rectifier with 'R' load:

Figure below shows the Single phase Full Wave Controlled Rectifiers with R load

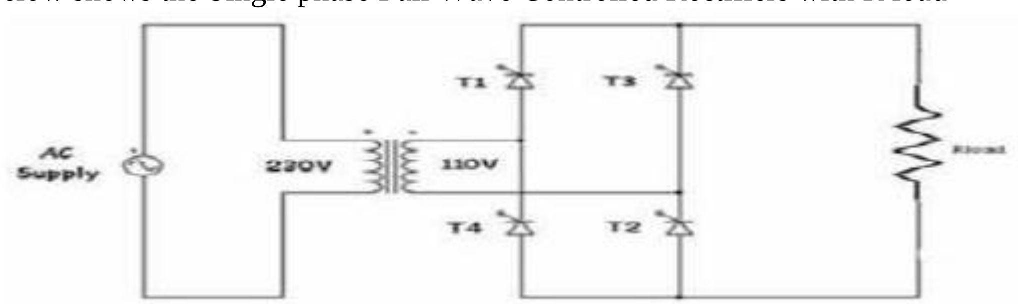


Figure: 2.7 single phase full converter circuit with R load

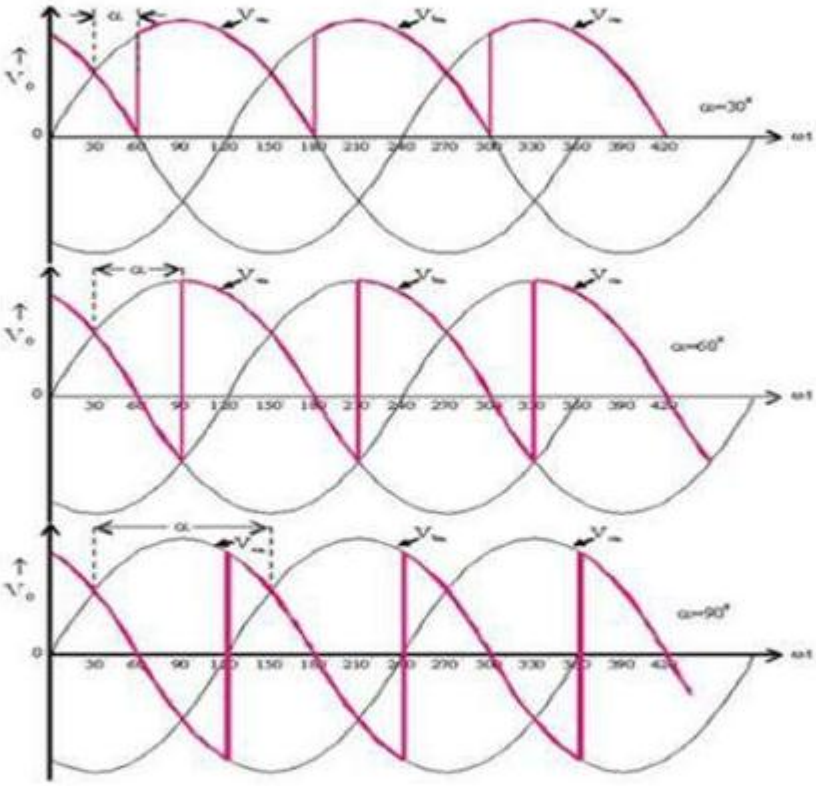


Figure: 2.8 single phase full converter circuit with R load input and output waveforms

• The single phase fully controlled rectifier allows conversion of single phase AC into DC. Normally this is used in various applications such as battery charging, speed control of DC motors and front end of UPS (Uninterruptible Power Supply) and SMPS (Switched Mode Power Supply).

• All four devices used are Thyristors. The turn-on instants of these devices are dependent on the firing signals that are given. Turn-off happens when the current through the device reaches zero and it is reverse biased at least for duration equal to the turn-off time of the device specified in the data sheet.

• In positive half cycle Thyristors T1 & T2 are fired at an angle α .

• When T1 & T2 conducts

Vo=Vs

IO=is=Vo/R=Vs/R

• In negative half cycle of input voltage, SCR's T3 &T4 are triggered at an angle of (π+α)

• Here output current & supply current are in opposite direction

∴ is=-io

T3 & T4 becomes off at 2π.

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi+\alpha} Vmsinwt\, d(wt) = \frac{2Vm}{\pi}\ cos\alpha$$

ingle Phase Full Wave Controlled Rectifier with RE load:

Figure below shows Single phase Full Wave Controlled Rectifiers with RL load.

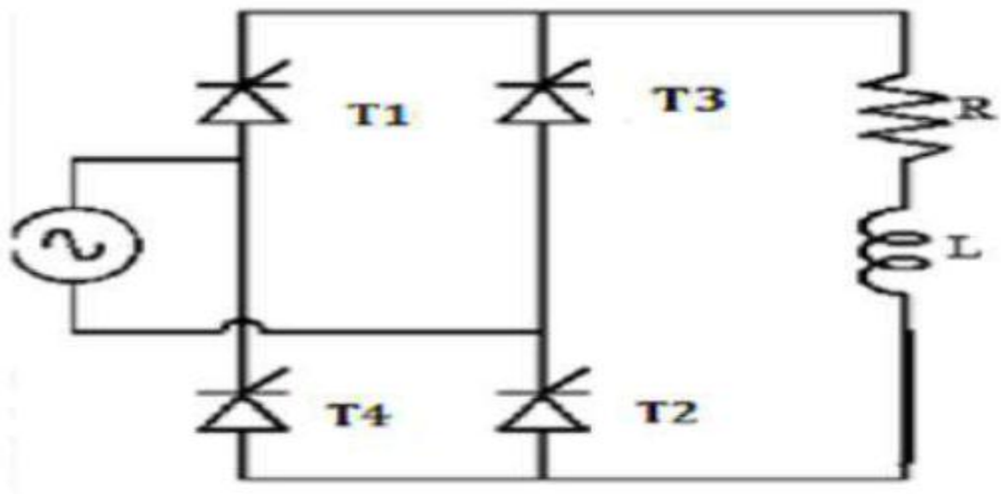


Figure: 2.9 single phase full converter circuit with RL load

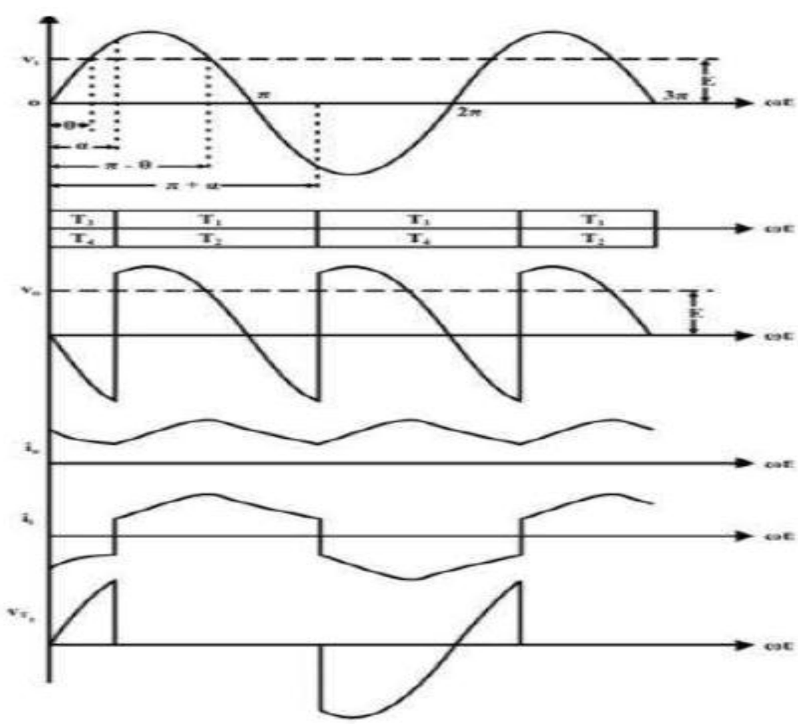


Figure: 2.10 single phase full converter circuit with RL load input and output waveforms

Operation of this mode can be divided between four modes

Mode 1 (α toπ)

• In positive half cycle of applied ac signal, SCR's T1 & T2 are forward bias & can be turned on at an angle α.

• Load voltage is equal to positive instantaneous ac supply voltage. The load current is positive, ripple free, constant and equal to Io.

• Due to positive polarity of load voltage & load current, load inductance will store energy.

Mode 2 (π toπ+α)

• At wt=π, input supply is equal to zero & after π it becomes negative. But inductance opposes any change through it.

• In order to maintain a constant load current & also in same direction. A self inducedemf appears across 'L' as shown.

• Due to this induced voltage, SCR's T1 & T2 are forward bais in spite the negative supply voltage.

• The load voltage is negative & equal to instantaneous ac supply voltage whereas load current is positive.

• Thus, load acts as source & stored energy in inductance is returned back to the ac supply.

Mode 3 (π+α to 2π)

• At wt=π+α SCR's T3 & T4 are turned on & T1, T2 are reversed bias.

• Thus , process of conduction is transferred from T1,T2 to T3,T4.

• Load voltage again becomes positive & energy is stored in inductor

• T3, T4 conduct in negative half cycle from (π+α) to 2π

• With positive load voltage & load current energy gets stored

Mode 4 (2π to 2π+α)

• At wt=2π, input voltage passes through zero.

• Inductive load will try to oppose any change in current if in order to maintain load current constant & in the same direction.

• Induced emf is positive & maintains conducting SCR's T3 & T4 with reverse polarity also.

• Thus VL is negative & equal to instantaneous ac supply voltage. Whereas load current continues to be positive.

• Thus load acts as source & stored energy in inductance is returned back to ac supply

- At $wt=\alpha$ or $2\pi+\alpha$, T3 & T4 are commutated and T1,T2 are turned on.

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi+\alpha} Vmsinwt\, d(wt) = \frac{2Vm}{\pi}\, cos\alpha$$

Single phase fully controlled converters with RLE load

The circuit diagram of a full wave bridge rectifier using thyristors in shown in figure below. It consists of four SCRs which are connected between single phase AC supply and a load.

This rectifier produces controllable DC by varying conduction of all SCRs.

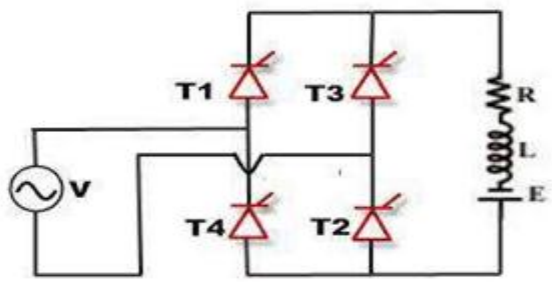


Figure: 2.11 single phase full converter circuit with RLE load

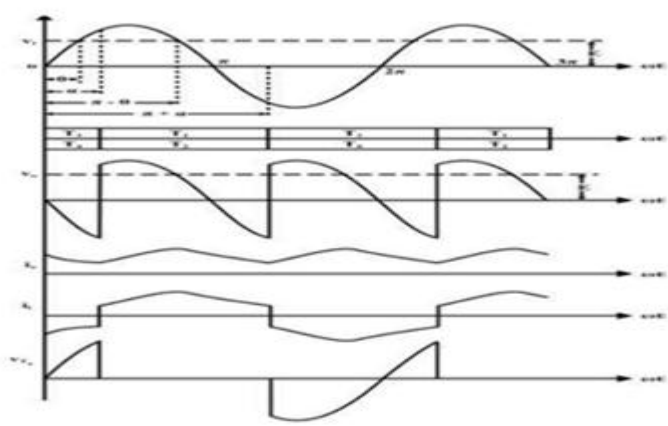


Figure: 2.12 single phase full converter circuit with RLE load input and output waveforms

In positive half-cycle of the input, Thyristors T1 and T2 are forward biased while T3 and T4 are reverse biased. Thyristors T1 and T2 are triggered simultaneously at some firing angle in the positive half cycle, and T3 and T4 are triggered in the negative half cycle.

The load current starts flowing through them when they are in conduction state. The load for this converter can be RL or RLE depending on the application.

By varying the conduction of each thyristor in the bridge, the average output of this converter gets controlled. The average value of the output voltage is twice that of half-wave rectifier.

The average output voltage is

$$V_0 = \frac{1}{\pi}\int_{\alpha}^{\pi+\alpha} Vmsinwt\ d(wt) = \frac{2Vm}{\pi}\ cos\alpha$$

Line commutated converters

For single phase half wave converter

1. Average DC load voltage: (V_{oavg})

$$V_{oavg} = V_0 = \frac{1}{T}\int_0^T Vmsinwt\ d(wt) \quad \text{where T is time period}$$

$$V_{oavg} = \frac{1}{2\pi}\left[\int_{\alpha}^{\pi} Vmsinwt\ d(wt) + \int_{\pi}^{2\pi+\alpha} 0\ d(wt)\right]$$

$$= \frac{1}{2\pi}\left[\int_{\alpha}^{\pi} Vmsinwt\ d(wt)\right]$$

$$= \frac{Vm}{2\pi}\left[-coswt\right]_{\alpha}^{\pi}$$

$$= \frac{Vm}{2\pi} - [cos\pi - cos\alpha]$$

$$= \frac{Vm}{2\pi}[1 + cos\alpha]$$

If $\alpha = 0$ $V_{oavg\ max} = \frac{Vm}{\pi}$

If $\alpha = 180$ $V_{oavg} = 0$

2. Average DC load current is given as

$$I_{oavg} = \frac{V0avg}{R}$$

DF

$$I_{oavg} = \frac{Vm}{2\pi R}[1 + cos\alpha]$$

3. RMS load voltage

$$V_{rms} = \left\{\frac{1}{T}\int_0^T Vm^2 sin^2 wt \; d(wt)\right\}^{1/2}$$

$$V_{rms} = \left\{\frac{1}{2\pi}\int_\alpha^\Pi Vm^2 sin^2 wt \; d(wt)\right\}^{1/2}$$

$$V_{rms} = \frac{Vm}{2\sqrt{\pi}}\left[(\pi - \alpha) + \frac{1}{2}sin2\alpha\right]^{1/2}$$

If $\alpha = 0$ $V_{rms} = \frac{Vm}{2}$

If $\alpha = 180$ $V_{rms} = 0$

The RMS voltage may be varied from 0 to $\frac{Vm}{2}$ by varying α from 180 to 0

4. Power delivered to the resistive load is given

P_L = (RMS load voltage)(RMS load current)

$= V_{rms} X \, I_{rms}$

$= \frac{Vrms^2}{R} = Irms^2 XR$

5. Input volt amperes = (RMS source voltage)(RMS line current)

$= V_s \, I_{rms}$

$= V_s \frac{\sqrt{2}Vs}{R2\sqrt{\pi}}\left[(\pi - \alpha) + \frac{1}{2}sin2\alpha\right]^{1/2}$

$= \frac{Vs^2}{\sqrt{2\pi}XR}\left[(\pi - \alpha) + \frac{1}{2}sin2\alpha\right]^{1/2}$

6. Input power factor: It is defined as the ratio of total mean input power to the total rms input volt amperes

$$\text{Input power factor} = \frac{\frac{\sqrt{2}Vs}{2\sqrt{\pi}}\left[(\pi-\alpha)+\frac{1}{2}sin2\alpha\right]^{1/2}}{Vs}$$

$$= \frac{1}{\sqrt{2\pi}}\left[(\pi-\alpha)+\frac{1}{2}sin2\alpha\right]^{1/2}$$

7. Form factor: Form factor is defined as the ratio of RMS voltage to the average DC voltage

$$\text{Form Factor} = \frac{\text{Vrms}}{\text{Vavg}}$$

8. Effective value of the AC component of the output voltage

$$V_{ac} = [\text{Vrms}^2 - \text{Vavg}^2\,]^{1/2}$$

9. Ripple factor (R_f)

   It is defined as the ratio of AC component to the DC. Where ripple is the amount of AC component present in DC component

$$R_f = \frac{\text{Vac}}{\text{Vavg}} = \frac{[\text{Vrms}^2 - \text{Vavg}^2\,]^{1/2}}{\text{Vavg}} = \left[\left(\frac{\text{Vrms}}{\text{Vavg}}\right)^2 - 1\right]^{1/2} = \sqrt{FF^2 - 1}$$

10. Transformer Utilization Factor (TUF):

    It is defined as the ratio of output DC power to the volt ampere rating of the transformer

$$\text{TUF} = \frac{\text{Pdc}}{\text{VA rating of secondary winding of the transformer}}$$

11. Rectifier efficiency:

    It is defined as the ratio of output DC power to the input ac power

$$\eta = \frac{\text{VavgIavg}}{\text{VrmsIrms}}$$

12. Peak inverse voltage (PIV):

    It is defined as the maximum voltage that an SCR can be subjected to in the reverse biased condition

    In the case of Half wave rectifier it is V_m

Effect of source inductance in single phase rectifier

Fig. below shows a single phase fully controlled converter with source inductance. For simplicity it has been assumed that the converter operates in the continuous conduction mode. Further, it has been assumed that the load current ripple is negligible and the load can be replaced by a dc current source the magnitude of which equals the average load current. Fig. shows the corresponding waveforms

It is assumed that the Thyristors T3 and T4 were conducting at t = 0. T1 and T2 are fired at $\omega t = \alpha$. If

there were no source inductance T3 and T4 would have commutated as soon as T1 and T2 are turned ON.

The input current polarity would have changed instantaneously. However, if a source inductance is present the commutation and change of input current polarity cannot be instantaneous. s. Therefore, when T1 and T2 are turned ON T3 T4 does not commutate immediately. Instead, for some interval all four Thyristors continue to conduct as shown in Fig. 2.14. This interval is called "overlap" interval.

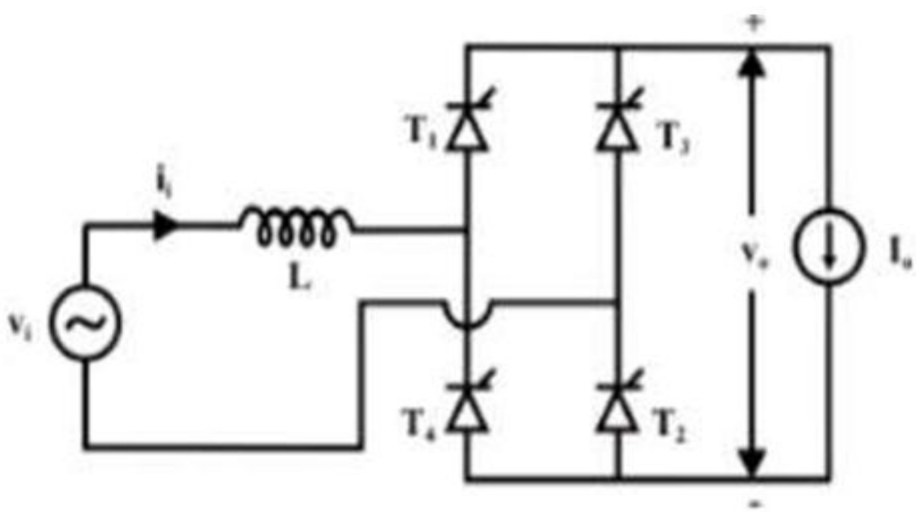


Figure: 2.13 single phase full converter circuit with source inductance

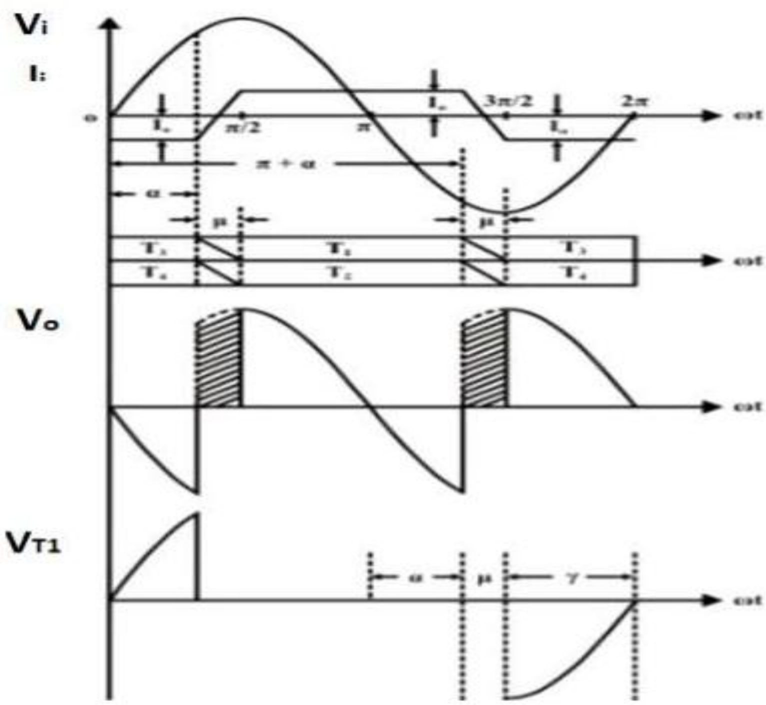


Figure: 2.14 single phase full converter output waveforms with source inductance

1. During overlap interval the load current freewheels through the thyristors and the output voltage is clamped to zero. On the other hand, the input current starts changing polarity as the current through T1 and T2 increases and T3 T4 current decreases. At the end of the overlap interval the current through T3 and T4 becomes zero and they commutate, T1 and T2 starts conducting the full load current
2. The same process repeats during commutation from T1 T2 to T3T4 at $\omega t = \pi + \alpha$. From Fig. 2.14 it is clear that, commutation overlap not only reduces average output dc voltage but also reduces the extinction angle γ which may cause commutation failure in the inverting mode of operation if α is very close to 180^o.
3. In the following analysis an expression of the overlap angle "μ" will be determined. From the equivalent circuit of the converter during overlap period.

$$L\frac{di_i}{dt} = v_i \ \ for \ \ \alpha \leq \omega t + \mu$$

$$i_i(\omega t = \alpha) = -I_0$$

$$i_i = I - \frac{\sqrt{2}V_i}{\omega L} cos\omega t$$

$$\therefore i_i|_{\omega t - \alpha} = I - \frac{\sqrt{2}V_i}{\omega L} cos\alpha = -I_0$$

$$I = \frac{\sqrt{2}V_i}{\omega L}\cos\alpha - I_0$$

$$\therefore \quad i_i = \frac{\sqrt{2}V_i}{\omega L}(\cos\alpha - \cos\omega t) - I_0$$

at $\omega t = \alpha + \mu \quad i_i = I_0$

$$I_0 = \frac{\sqrt{2}V_i}{\omega L}(\cos\alpha - \cos(\alpha + \mu)) - I_0$$

$$\therefore \quad \cos\alpha - \cos(\alpha + \mu) = \frac{\sqrt{2}\omega L}{V_0} I_0$$

$$V_0 = \frac{I}{\pi}\int_{\alpha}^{\alpha+\pi} V_i d\omega t$$

or $$V_0 = \frac{I}{\pi}\int_{\alpha+\mu}^{\alpha+\pi} \sqrt{2}v_i \sin \omega t d\omega t$$

$$= \frac{\sqrt{2}v_i}{\pi}[\cos(\alpha + \mu) - \cos(\pi + \alpha)$$

$$= \frac{\sqrt{2}v_i}{\pi}[\cos\alpha + \cos(\alpha + \mu)]$$

$$\therefore V_0 = 2\sqrt{2}\frac{v_i}{\pi}[cos\alpha - cos(\alpha + \mu)]$$ 59 | P a g e

$$\therefore V_0 = \frac{2\sqrt{2}}{\pi}v_i cos\alpha - \frac{2}{\pi}\omega L I_0$$

Operation of three phase half wave rectifier with R and RL loads

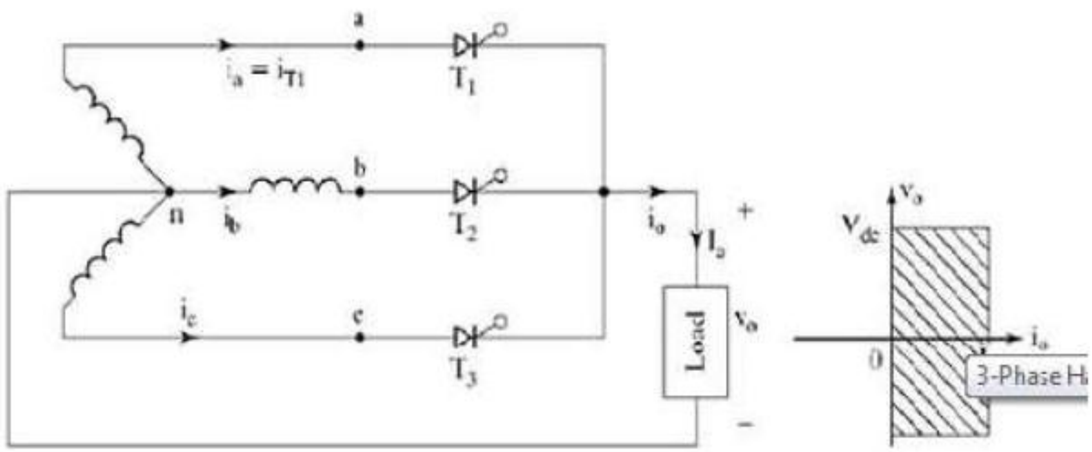


Figure: 2.16 circuit diagram three phase half wave rectifier

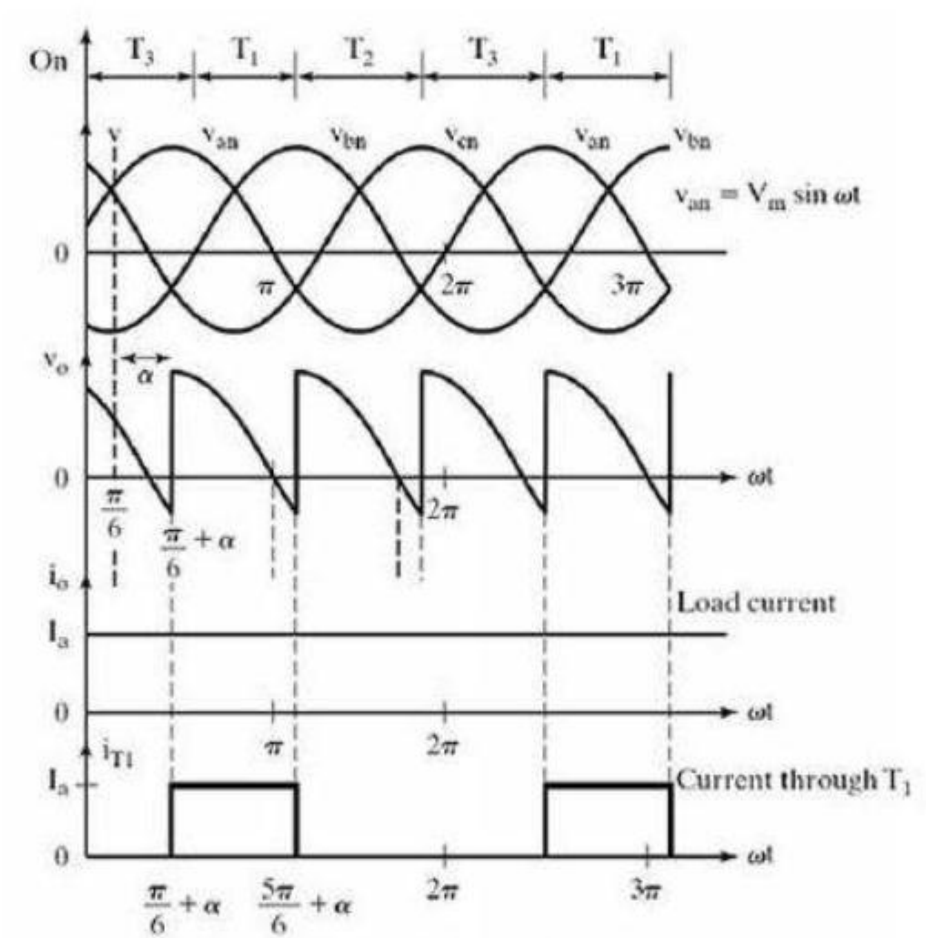


Figure: 2.17 input and output waveforms of three phase half wave rectifier

Three phase supply voltage equations

We define three line neutral voltages (3 phase voltages) as follows

$V_{RN} = V_{an} = V_m$ sinwt where V_m is the maximum voltage

$V_{YN} = V_{bn} = V_m \sin(wt-\frac{2\pi}{3})$

$V_{BN} = V_{cn} = V_m \sin(wt-\frac{4\pi}{3})$

The **3-phase half wave converter** combines three **single phase half wave controlled rectifiers in one** single circuit feeding a common load. The thyristor T_1 in series with one of the supply phase windings *'a-n'* acts as one half wave controlled rectifier The second thyristor T_2 in series with the supply phase winding *'b-n'* acts
as the second half wave controlled rectifier. The third thyristor T_3in series with the supply phase winding acts as the third half wave controlled rectifier. cathode point.

When the thyristor T_1 is triggered at $\omega t=(\prod/6 + \alpha)=(30° + \alpha)$, the phase voltage V_{an} appears across the load when T_1 conducts. The load current flows through the supply phase winding *'a- n'* and through thyristor T_1 as long as T_1 conducts.

When thyristor T_2 is triggered at $\omega t=(5\prod/6\alpha)$, T_1 becomes reverse biased and turns-off. The load current flows through the thyristor and through the supply phase winding *'b-n'* . When T_2 conducts the phase voltage v_{bn} appears across the load until the thyristor T_3 is triggered .

When the thyristor T_3 is triggered at $\omega t=(3\prod/2 + \alpha)=(270°+\alpha)$, T_2 is reversed biased and hence T_2turns-off. The phase voltage V_{an} appears across the load when T_3 conducts.

The 3-phase input supply is applied through the star connected supply transformer as shown in the figure. The common neutral point of the supply is connected to one end of the load while the other end of the load connected to the common

When T_1 is triggered again at the beginning of the next input cycle the thyristor T_3 turns off as it is reverse biased naturally as soon as T_1 is triggered. The figure shows the 3-phase input supply voltages, the output voltage which appears across the load, and the load current assuming a constant and ripple free load current for a highly inductive load and the current through the thyristor T_1.

For a purely resistive load where the load inductance 'L = 0' and the trigger angle $\alpha >(\prod/6)$, the load current appears as discontinuous load current and each thyristor is naturally commutated when the polarity of the corresponding phase supply voltage reverses. The frequency of output ripple frequency for a **3-phase half wave converter** is f_s, where f_s is the input supply frequency. 3

The **3-phase half wave converter** is not normally used in practical converter systems because of the disadvantage that the supply current waveforms contain dc components (i.e., the supply current waveforms have an average or dc value).

To derive an expression for the average output voltage of a 3-phase half wave converter for continuous load current

The reference phase voltage is $v_{RN}=v_{an}=V_m sin\omega t$. The trigger angle is measured from the cross over points of the 3-phase supply voltage waveforms. When the phase supply voltage V_{an} begins its positive half cycle at $\omega t=0$, the first cross over point appears at $\omega t=(\prod/6) radians$ 30°.

The trigger angle α for the thyristor T_1 is measured from the cross over point at . The thyristor T_1 is forward biased during the period $\omega t=30°$ to 150° , when the phase supply voltage v_{an} has higher amplitude than the other phase supply voltages. Hence T_1 can be triggered between 30° to 150°. When the thyristor T_1 is triggered at a trigger angle α, the average or dc output voltage for continuous load current is calculated using the equation

$$V_{avg} = \frac{3}{2\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{5\pi}{6}+\alpha} Vm\sin wt \, d(wt)$$

$$= \frac{3Vm}{2\pi}\left[(-cos\alpha)_{\frac{\pi}{6}+\alpha}^{\frac{5\pi}{6}+\alpha}\right]$$

$$= \frac{3\sqrt{3}Vm}{2\pi}\ cos\alpha$$

$$= \frac{3Vml}{2\pi}\ cos\alpha$$

Operation of three phase half controlled rectifier with R and RL loads

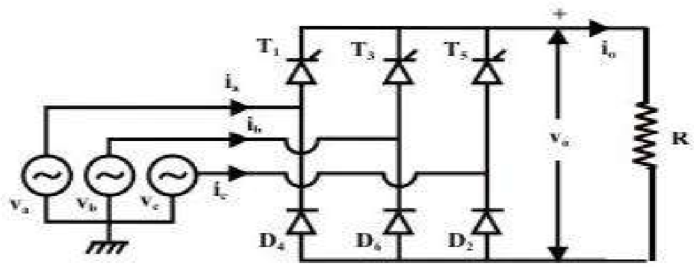


Figure: 2.18 circuit diagram three phase half controlled rectifier

Three phase half wave controlled rectifier output voltage waveforms for different trigger angles with R load

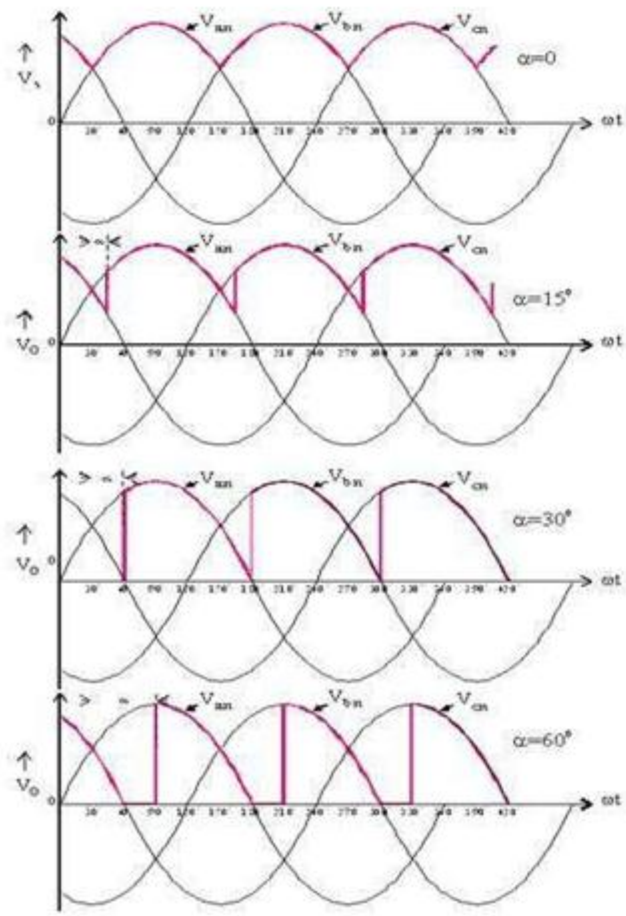


Figure: 2.19 input and output waveforms of three phase half controlled rectifier with R load

Three single phase half wave converters can be connected to form a three phase half wave converter. Similarly three phase semi converter uses 3 SCRs T1, T3 & T5 and 3 diodes D2, D4&D6 In the circuit shown above when any device conducts, line voltage is applied across load. so line voltage are necessary to draw Phase shift between two line voltages is 60 degree & between two phase voltages it is 120 degree Each phase & line voltage is sine wave with the frequency of 50 Hz. R,Y,B are phase voltages with respect to ‘N’.

In the case of a **three-phase half wave controlled** rectifier with resistive load, the thyristor T_1 is triggered at $\omega t=(30°+\alpha)$ and T_1 conducts up to $\omega t=180°=$&pron; radians. When the phase supply voltage decreases to zero at , the load current falls to zero and the thyristor T_1 turns off. Thus T_1 conducts from $\omega t=(30° + \alpha)$ to $(180°)$.

Hence the average dc output voltage for a 3-pulse converter (3-phase half wave controlled rectifier) is calculated by using the equation

$$\text{The average output voltage } V_{avg} = \frac{3}{2\pi}\int_{\frac{\pi}{3}+\alpha}^{\frac{2\pi}{3}} \text{Vmlsinwt d(wt)} + \int_{\frac{\pi}{3}}^{\frac{2\pi}{3}+\alpha} \text{Vmlsinwt d(wt)}$$

$$= \frac{3Vml}{2\pi}(1+\cos\alpha)$$

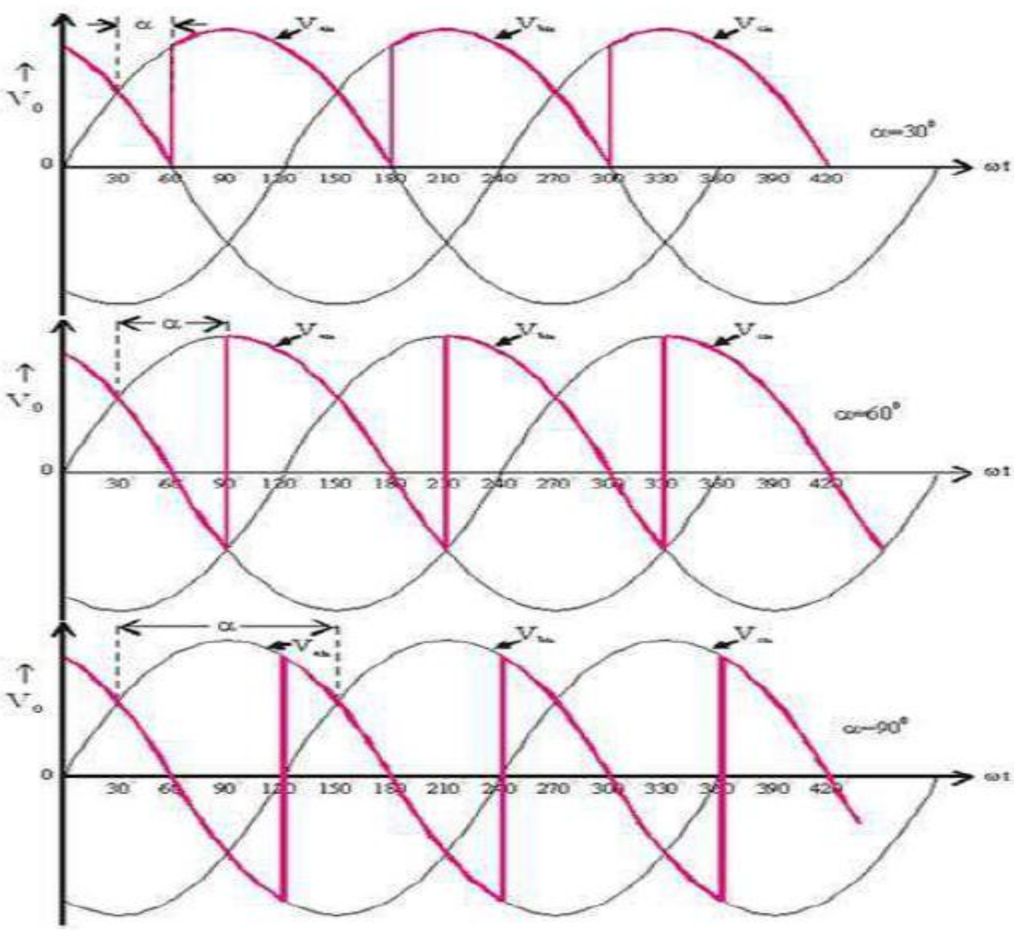


Figure: 2.19 Input and output waveforms of three phase half controlled rectifier with RL load

Operation of three phase fully controlled rectifier with R and RL loads

Three phase full converter is a fully controlled bridge controlled rectifier using six thyristors connected in the form of a full wave bridge configuration. All the six thyristors are controlled switches which are turned on at a appropriate times by applying suitable gate trigger signals.

The **three phase full converter** is extensively used in industrial power applications upto about 120kW output power level, where two quadrant operations is required. The figure shows a **three phase full converter** with highly inductive load. This circuit is also known as three phase full wave bridge or as a six pulse converter.

The thyristors are triggered at an interval of ($\prod/3$) radians (i.e. at an interval of 30°). The frequency of output ripple voltage is $6f_s$ and the filtering requirement is less than that of **three phase semi and half wave converters**.

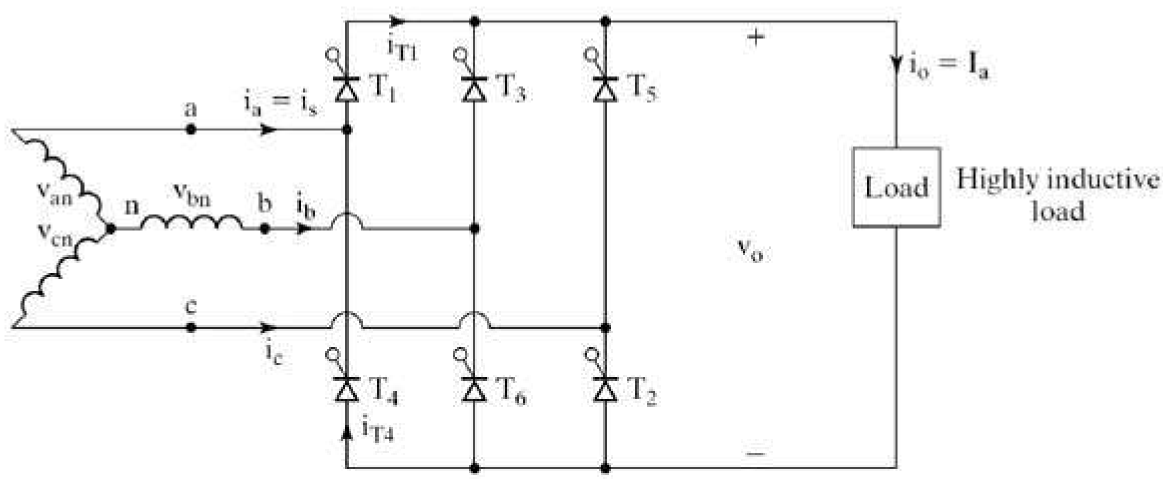


Figure: 2.20 circuit diagram three phase fully controlled rectifier with R and RL load

At $\omega t=(\prod/6 +\alpha)$, thyristor is already conducting when the thyristor is turned on by applying the gating signal to the gate of . During the time period $\omega t=(\prod/6 +\alpha)$ to ($\prod/2 +\alpha$), thyristors and conduct together and the line to line supply voltage appears across the load.

At $\omega t=(\prod/2 +\alpha)$, the thyristor T_2 is triggered and T_6 is reverse biased immediately and T_6 turns off due to natural commutation. During the time period $\omega t=(\prod/ +\alpha)$ *to* $(5\prod/6 +\alpha)$, thyristor T_1 and T_2 conduct together and the line to line supply voltage appears across the load.

The thyristors are numbered in the circuit diagram corresponding to the order in which they are triggered. The trigger sequence (firing sequence) of the thyristors is 12, 23, 34, 45, 56, 61, 12, 23, and so on. The figure shows the waveforms of three phase input supply voltages, output voltage, the thyristor current through T_1 and T_4, the supply current through the line

PE

We define three line neutral voltages (3 phase voltages) as

follows V_{RN}

We define three line neutral voltages (3 phase voltages) as follo

$V_{RN} = V_{an} = V_m$ sinwt where V_m is the maximum voltage

$V_{YN} = V_{bn} = V_m \sin\left(wt-\frac{2\pi}{3}\right)$

$V_{BN} = V_{cn} = V_m \sin\left(wt-\frac{4\pi}{3}\right)$

The corresponding line to line voltages are

$$V_{RY} = V_{ab} = V_{an} - V_{bn} = \sqrt{3}\,\text{Vm}\sin\left(wt+\frac{\pi}{6}\right)$$

$$V_{YB} = V_{bc} = V_{bn} - V_{cn} = \sqrt{3}\,\text{Vm}\sin\left(wt-\frac{\pi}{2}\right)$$

$$V_{BR} = V_{ca} = V_{cn} - V_{an} = \sqrt{3}\,\text{Vm}\sin\left(wt+\frac{\pi}{2}\right)$$

To derive an expression for the average output voltage of **three phase full converter** with highly inductive load assuming continuous and constant load current

The output load voltage consists of 6 voltage pulses over a period of 2∏ radians, hence the average output voltage is calculated as

$$V_{avg} = \frac{6}{2\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{\pi}{2}+\alpha} Vod(wt)$$

$$V_o = V_{ab} = \sqrt{3}\ Vm \sin\left(wt + \frac{\pi}{6}\right)$$

$$V_{avg} = \frac{3}{\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{\pi}{2}+\alpha} \sqrt{3}\ Vm \sin\left(wt + \frac{\pi}{6}\right) d(wt)$$

$$= \frac{3\sqrt{3}Vm}{\pi}\ cos\alpha$$

$$= \frac{3Vml}{\pi}\ cos\alpha$$

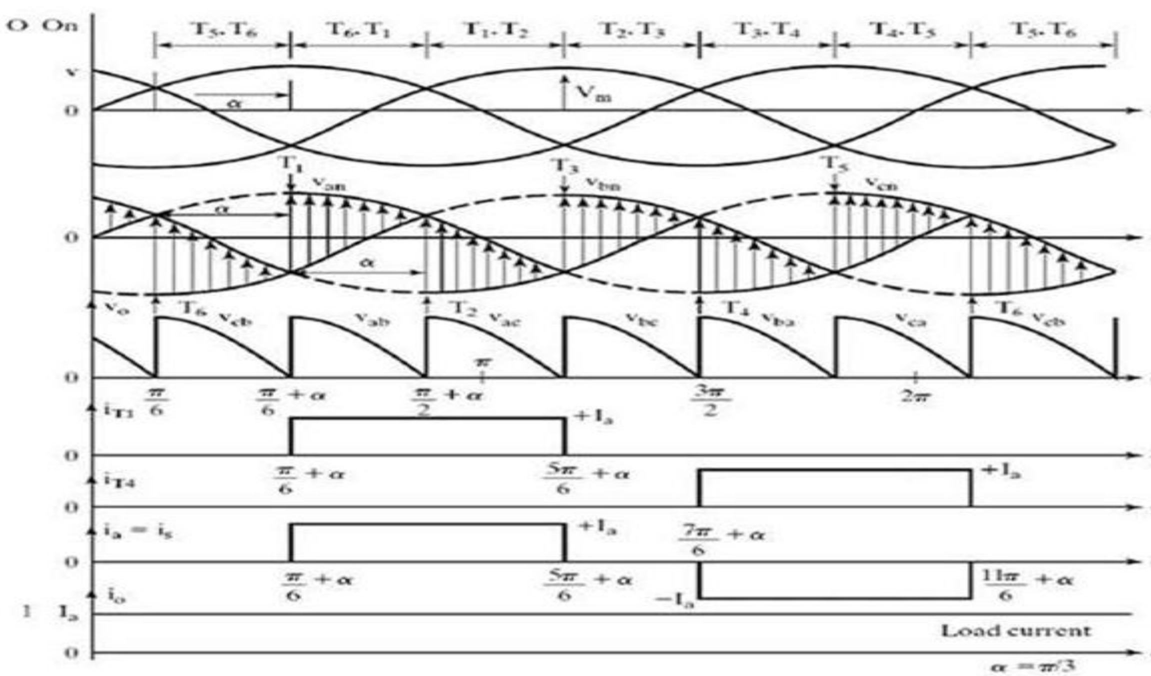
The RMS value of the output voltage is found from

$$V_{orms} = \left[\frac{6}{2\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{\pi}{2}+\alpha} Vo^2 d(wt)\right]^{1/2}$$

$$= \left[\frac{6}{2\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{\pi}{2}+\alpha} Vab^2 d(wt)\right]^{1/2}$$

$$= \left[\frac{3}{\pi}\int_{\frac{\pi}{6}+\alpha}^{\frac{\pi}{2}+\alpha} 3\ Vm^2 \sin^2\left(wt + \frac{\pi}{6}\right) d(wt)\right]^{1/2}$$

$$= \sqrt{3}Vm\left(\frac{1}{2} + \frac{3\sqrt{3}}{4\pi}cos2\alpha\right)^{1/2}$$



Operation of three phase half wave rectifier with RLE loads

A three phase fully controlled converter is obtained by replacing all the six diodes of an uncontrolled converter by six thyristors as shown in Figure

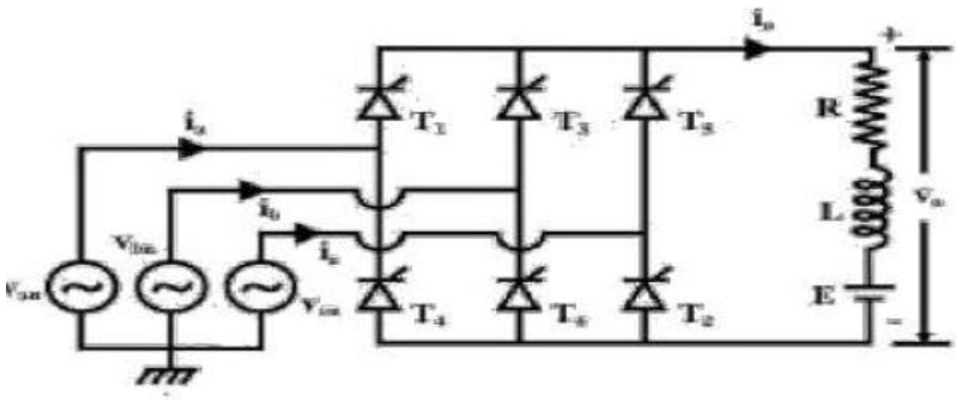


Figure: 2.22 circuit diagram of three phase fully controlled rectifier with RLE load

For any current to flow in the load at least one device from the top group (T1, T3, T5) and one from the bottom group (T2, T4, T6) must conduct. It can be argued as in the case of an uncontrolled converter only one device from these two groups will conduct.

Then from symmetry consideration it can be argued that each thyristor conducts for 120° of the input cycle. Now the thyristors are fired in the sequence T1 → T2 → T3 → T4 → T5 → T6 → T1 with 60° interval between each firing. Therefore thyristors on the same phase leg are fired at an interval of 180° and hence can not conduct simultaneously. This leaves only six possible conduction mode for the converter in the continuous conduction mode of operation. These are T1T2, T2T3, T3T4, T4T5, T5T6, T6T1. Each conduction mode is of 60° duration and appears in the sequence mentioned. Each of these line voltages can be associated with the firing of a thyristor with the help of the conduction table-1. For example the thyristor T1 is fired at the end of T5 T6 conduction interval. During this period the voltage across T1 was vac. Therefore T1 is fired α angle after the positive going zero crossing of vac. similar observation can be made about other thyristors.

Fig. 2.23 shows the waveforms of different variables. To arrive at the waveforms it is necessary to draw the conduction diagram which shows the interval of conduction for each thyristor and can be drawn with the help of the phasor diagram of fig. 2.22. If the converter firing angle is α each thyristor is fired "α"angle after the positive going zero crossing of the line voltage with which it's firing is associated. Once the conduction diagram is drawn all other voltage waveforms can be drawn from the line voltage waveforms and from the conduction table of fig. 2.22. Similarly line currents can be drawn from the output current and the conduction diagram. It is clear from the waveforms that output voltage and current waveforms are periodic over one sixth of the input cycle. Therefore this converter is also called

the "six pulse" converter. The input current on the other hand contains only odds harmonics of the input frequency other than the triplex (3rd, 9th etc.) harmonics. The next section will analyze the operation of this converter in more details.

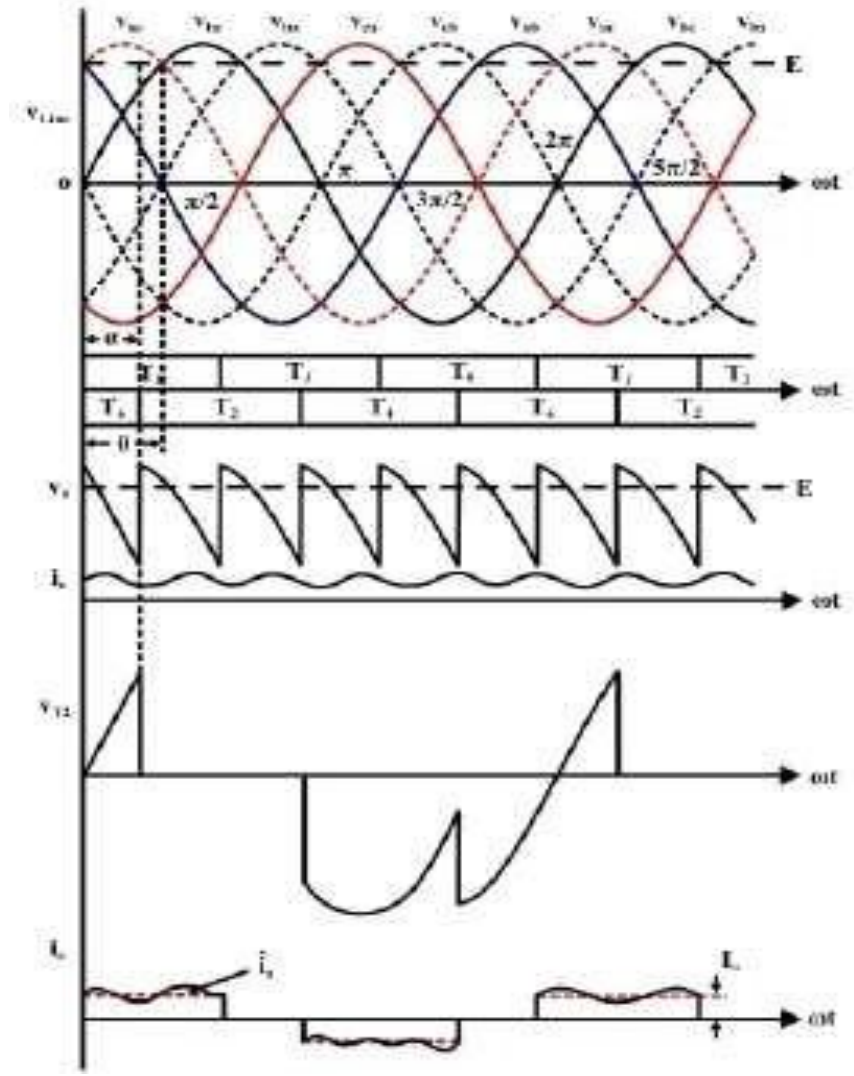


Figure: 2.23 Input and output waveforms of three phase fully controlled rectifier in rectifier mode

PE

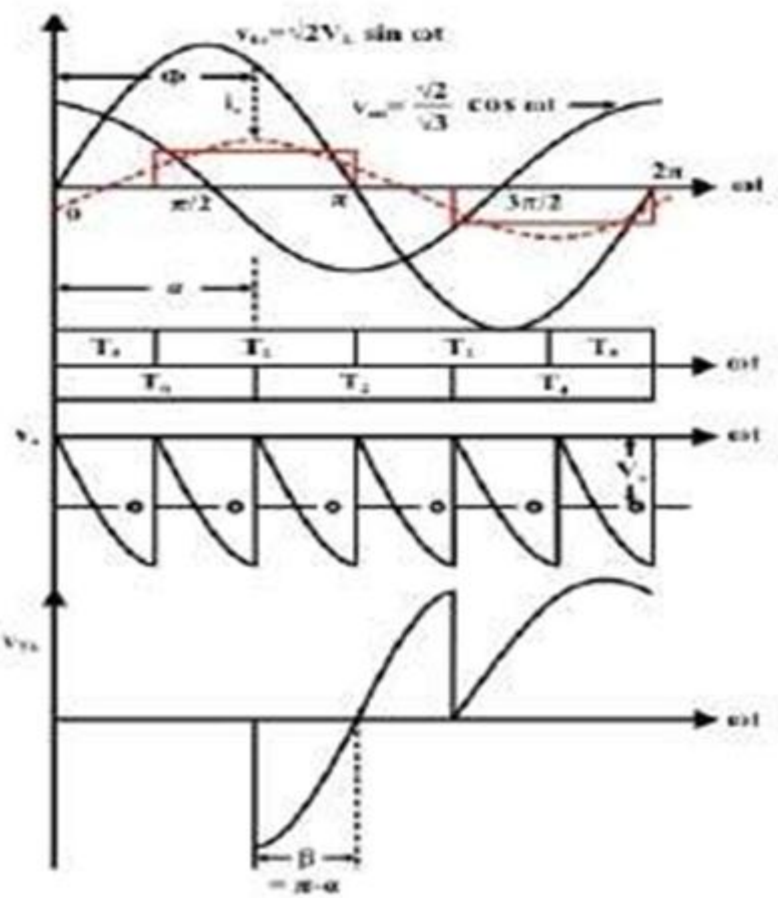


Figure: 2.24 Input and output waveforms of three phase fully controlled rectifier in inversion mode

Effect of source inductance in three phase rectifiers

The three phase fully controlled converter was analyzed with ideal source with no internal impedance. When the source inductance is taken into account, the qualitative effects on the performance of the converter is similar to that in the case of a single phase converter. Fig. 2.25 shows such a converter. As in the case of a single phase converter the load is assumed to be highly inductive such that the load can be replaced by a current source.

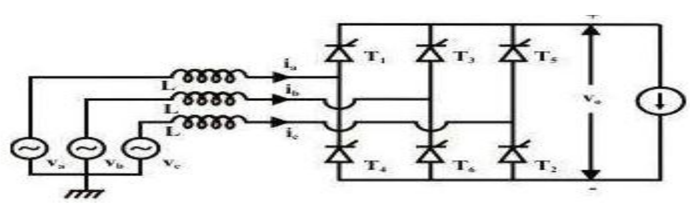


Figure: 2.25 circuit diagram for three phase rectifier with source inductance

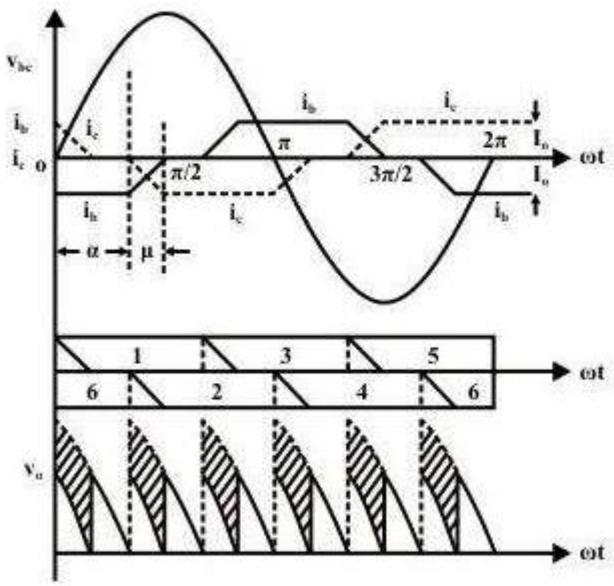


Figure: 2.26 waveforms for three phase rectifier with source inductance

As in the case of a single phase converter, commutations are not instantaneous due to the presence of source inductances. It takes place over an overlap period of "μ" instead. During the overlap period three thyristors instead of two conducts. Current in the outgoing thyristor gradually decreases to zero while the incoming thyristor current increases and equals the total load current at the end of the overlap period. If the duration of the overlap period is greater than 60º four thyristors may also conduct clamping the output voltage to zero for some time. However, this situation is not very common and will not be discussed any further in this lesson. Due to the conduction of two devices during commutation either from the top group or the bottom group the instantaneous output voltage during the overlap period drops (shown by the hatched portion of Fig. 2.26 resulting in reduced average voltage. The exact amount of this reduction can be calculated as follows.

In the time interval $\alpha < \omega t \le \alpha + \mu$, T_6 and T_2 from the bottom group and T_1 from the top group conducts.

The equivalent circuit of the converter during this period is given by the circuit diagram of Fig. 2.27

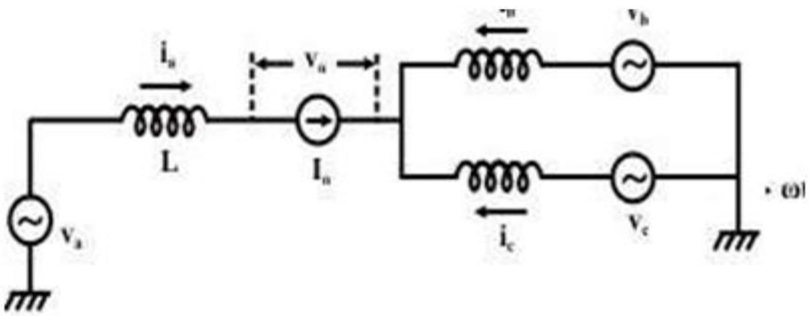


Figure: 2.27 Equivalent circuit of waveforms with source inductance

Therefore, in the interval $\alpha < \omega t \leq \alpha + \mu$

$$v_b = L\frac{di_b}{dt} - L\frac{di_c}{dt} + v_c$$

or, $$v_{bc} = L\frac{d}{dt}(i_b - i_c)$$

but $\quad i_b + i_c + I_o = 0 \qquad \therefore \quad \frac{di_b}{dt} = -\frac{di_c}{dt}$

$\therefore$ $$2L\frac{d}{dt}i_b = v_{bc} = \sqrt{2}V_L \sin\omega t$$

$\therefore$ $$i_b = C - \frac{\sqrt{2}V_L}{2\omega L}\cos\omega t$$

at $\omega t = \alpha, \quad i_b = -I_0 \qquad \therefore \quad C = \frac{\sqrt{2}V_L}{2\omega L}\cos\alpha - I_0$

$\therefore$ $$i_b = \frac{\sqrt{2}V_L}{2\omega L}(\cos\alpha - \cos\omega t) - I_0$$

at $\omega t = \alpha + \mu, \quad i_b = 0$

$\therefore$ $$\frac{\sqrt{2}V_L}{2\omega L}(\cos\alpha - \cos(\alpha + \mu)) = I_0$$

Or, $$\cos\alpha - \cos(\alpha + \mu) = \frac{\sqrt{2}\omega L}{V_L}I_0$$

DC – DC converters

A chopper uses high speed to connect and disconnect from a source load. A fixed DC voltage is applied intermittently to the source load by continuously triggering the power switch ON/OFF. The period of time for which the power switch stays ON or OFF is referred to as the chopper's ON and OFF state times, respectively.

Choppers are mostly applied in electric cars, conversion of wind and solar energy, and DC motor regulators.

Symbol of a Chopper

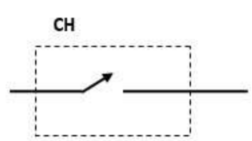


Figure: 3.1 symbol of chopper

Control strategies of Chopper

In DC-DC converters, the average output voltage is controlled by varying the alpha (α) value. This is
achieved by varying the Duty Cycle of the switching pulses. Duty cycle can be varied usually in 2 ways:

1. Time Ratio Control
2. Current Limit Control

In this post we shall look upon both the ways of varying the duty cycle. Duty Cycle is the ratio of 'On Time' to 'Time Period of a pulse'.

Time Ratio Control: As the name suggest, here the time ratio (i.e. the duty cycle ratio Ton/T) is varied. This kind of control can be achieved using 2 ways:

• Pulse Width Modulation (PWM) • Frequency Modulation Control (FMC)

Pulse Width Modulation (PWM)

In this technique, the time period is kept constant, but the 'On Time' or the 'OFF Time' is varied. Using this, the duty cycle ratio can be varied. Since the ON time or the 'pulse width' is getting changed in this method, so it is popularly known as Pulse width modulation.

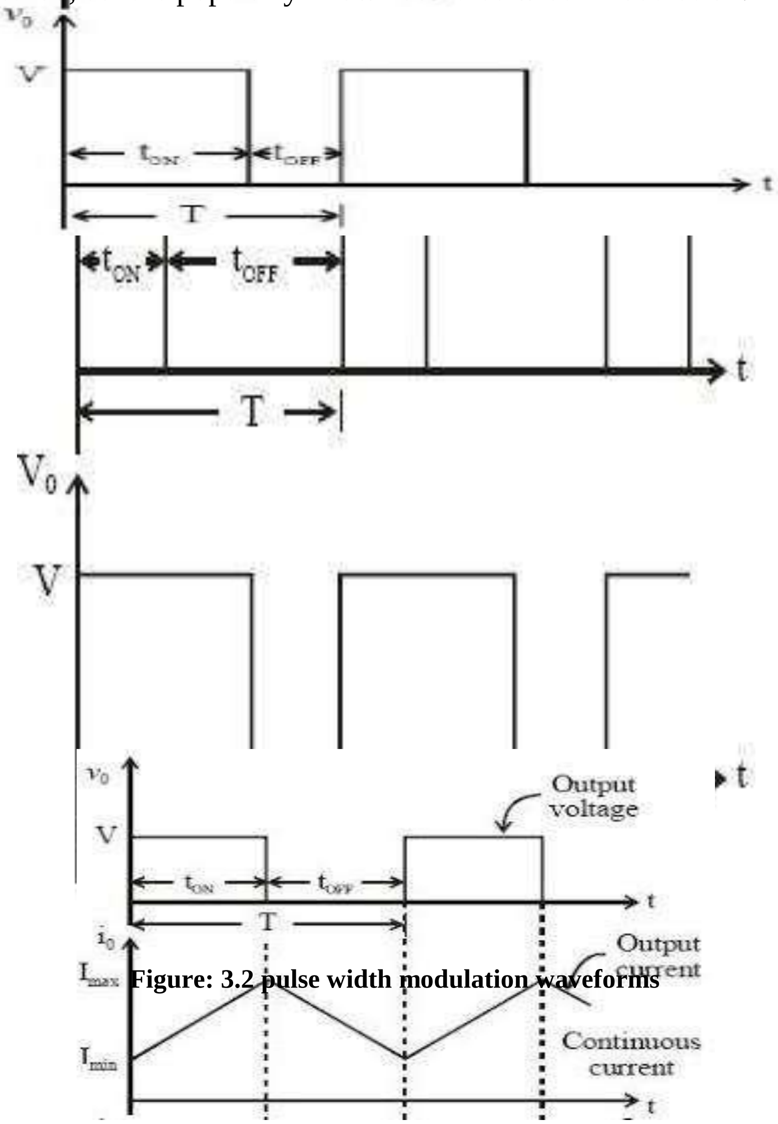


Figure: 3.2 pulse width modulation waveforms

Frequency Modulation Control (FMC)

In this control method, the 'Time Period' is varied while keeping either of 'On Time' or 'OFF time' as constant. In this method, since the time period gets changed, so the frequency also changes accordingly, so this method is known as frequency modulation control.

Figure: 3.3 Frequency modulation waveforms

Current Limit Control:

As is obvious from its name, in this control strategy, a specific limit is applied on the current variation.

In this method, current is allowed to fluctuate or change only between 2 values i.e. maximum current (I max) and minimum current (I min). When the current is at minimum value, the chopper is switched ON. After this instance, the current starts increasing, and when it reaches up to maximum value, the chopper is switched off allowing the current to fall back to minimum value. This cycle continues again and again.

Figure: 3.4 current limit control waveforms

Classification of Choppers

Depending on the voltage output, choppers are classified as −

1. Step Up chopper (boost converter)
2. Step Down Chopper(Buck converter)
3. Step Up/Down Chopper (Buck-boost converter)

Depending upon the direction of the output current and voltage, the converters can be classified into five classes namely

1. Class A [One-quadrant Operation]
2. Class B [One-quadrant Operation]
3. Class C [Two-quadrant Operation]
4. Class D Chopper [Two-quadrant Operation]
5. Class E Chopper [Four-quadrant Operation]

StepDownChopper

This is also known as a buck converter. In this chopper, the average voltage output V_O is less than the input voltage V_S. When the chopper is ON, $V_O = V_S$ and when the chopper is off, $V_O = 0$

When the chopper is ON −

$$VS=(VL+V0),\ VL=VS-V0,$$

$$Ldi/dt=VS-V0,$$

$$L\Delta i/TON=Vs+V0$$

$$VS=(VL+V0),$$

$$VL=VS-V0,$$

$$Ldi/dt=VS-V0,$$

$$L\Delta i/TON=Vs+V0$$

Thus, peak-to-peak current load is given by,

$$\Delta i=\frac{Vs-V0}{L}\ TON$$

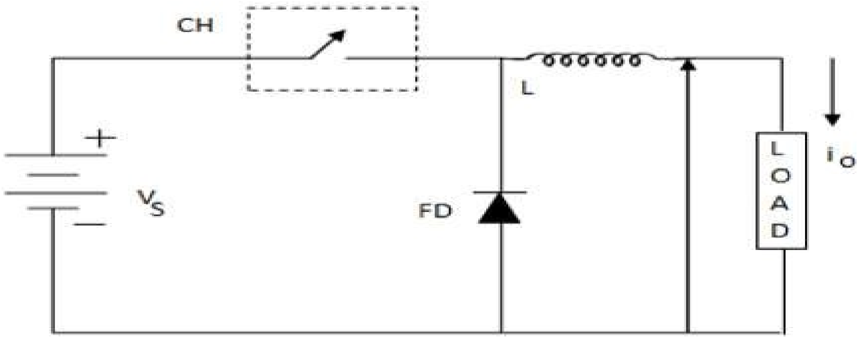


Figure: 3.5 Step down chopper

Where **FD** is free-wheel diode.

When the chopper is OFF, polarity reversal and discharging occurs at the inductor. The current passes through the free-wheel diode and the inductor to the load. This gives,

Ldi/dt=V0

Rewritten as

LΔi/TOFF=V0

LΔi/TOFF=V0

Δi=V0TOFF/L

From the above equations

$$\frac{V_S - V_0}{L} T_{ON} = \frac{V_0}{L} T_{OFF}$$
$$\frac{V_S - V_0}{V_0} = \frac{T_{OFF}}{T_{ON}}$$
$$\frac{V_S}{V_0} = \frac{T_{ON} - T_{OFF}}{T_{ON}}$$

$$V_0 = \frac{T_{ON}}{T} V_S = DV_S$$

$$\Delta i = \frac{V_S - DV_S}{L} DT, \text{ from } D = \frac{T_{ON}}{T}$$
$$= \frac{V_S - (1-D)D}{Lf}$$
$$f = \frac{1}{T} = \text{chopping frequency}$$

Current and Voltage Waveforms

The current and voltage waveforms are given below −

For a step down chopper the voltage output is always less than the voltage input. This is shown by the waveform below.

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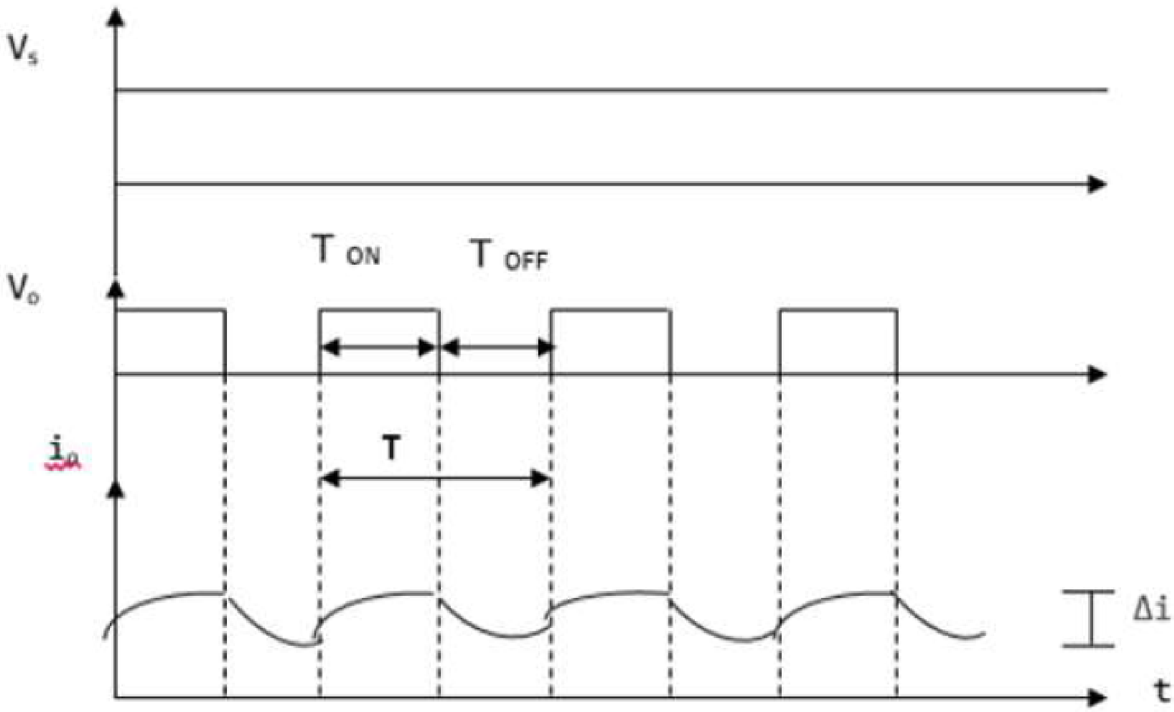

Figure: 3.6 Input and output waveforms

Step Up Chopper

The average voltage output (V_o) in a step up chopper is greater than the voltage input (V_s). The figure below shows a configuration of a step up chopper.

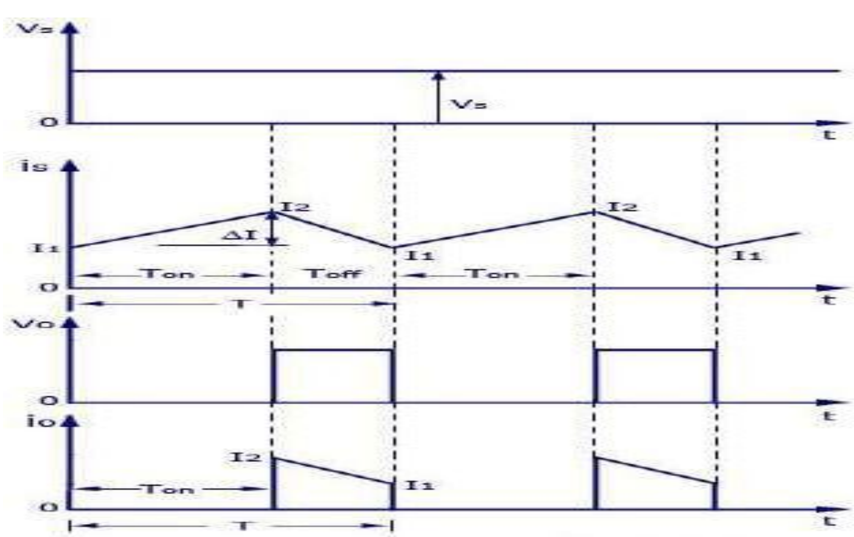

Figure: 3.7 circuit diagram of step up chopper

Current and Voltage Waveforms

V_0 (average voltage output) is positive when chopper is switched ON and negative when the chopper is OFF as shown in the waveform below.

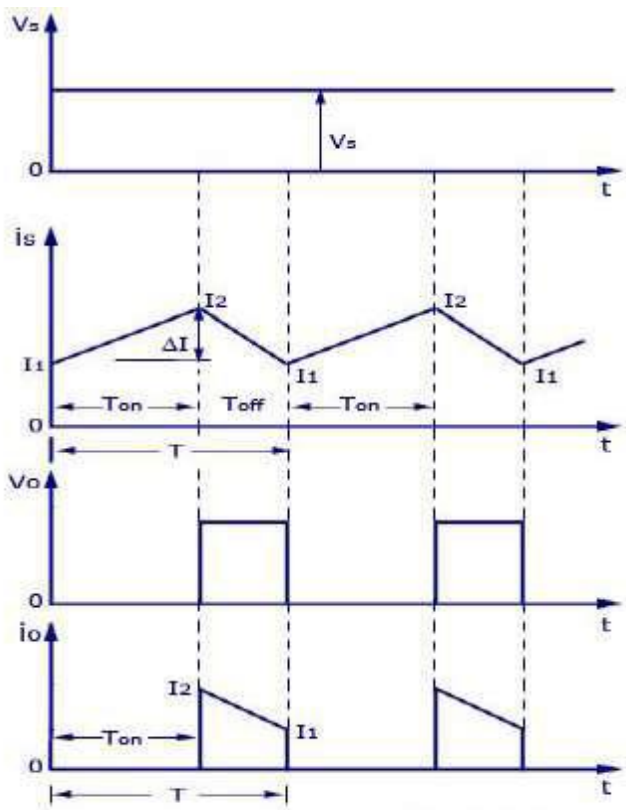

Figure: 3.8 Input and output waveforms of step up chopper

Where

T_{ON} – time interval when chopper is ON

T_{OFF} – time interval when chopper is OFF

V_L – Load voltage

V_s – Source voltage

T – Chopping time period = T_{ON} + T_{OFF}

V_o is given by –

$$V0 = \frac{1}{T}\int_0^{Ton} Vsdt$$

When the chopper (CH) is switched ON, the load is short circuited and, therefore, the voltage output for the period **T_{ON}** is zero. In addition, the inductor is charged during this time. This gives $V_S = V_L$

$$Vs = L\,\frac{di}{dt}, \frac{\Delta i}{Ton} = \frac{Vs}{L}$$

$$\Delta i = \frac{Vs}{L} \times Ton$$

Δi = is the inductor peak to peak current. When the chopper (CH) is OFF, discharge occurs through the inductor L. Therefore, the summation of the V_s and V_L is given as follows –

$$V0=VS+VL,\ VL=V0-VS$$

$$L\,\frac{di}{dt} = Vo - Vs$$

$$L\,\frac{\Delta i}{Toff} = Vo - Vs$$

$$\Delta i = \frac{Vo - Vs}{L} Toff$$

Equating Δi from on state to off state

$$\frac{Vs}{L} \times Ton = \frac{Vo - Vs}{L} Toff$$

,

$$Vo = \frac{TVs}{Toff}$$

$$Vo = \frac{Vs}{1 - D}$$

Step Up/ Step Down Chopper

his is also known as a buck-boost converter. It makes it possible to increase or reduce the voltage input level. The diagram below shows a buck-boost chopper

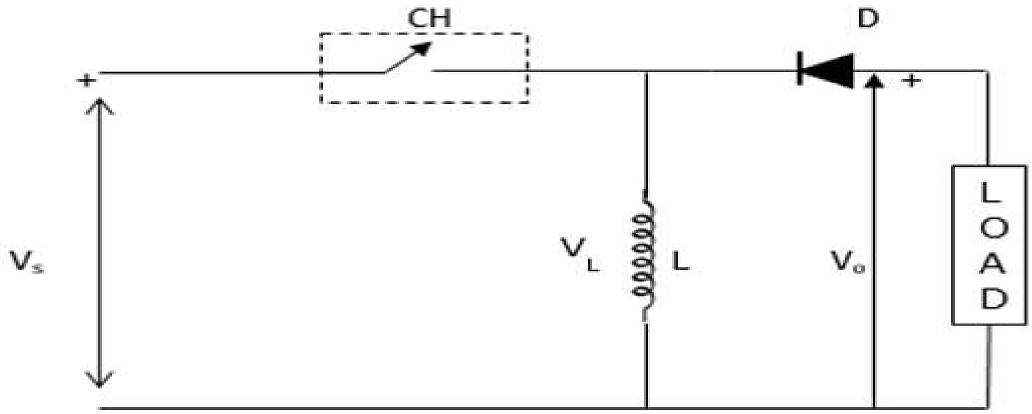

Figure: 3.9 circuit diagram of step up chopper

When the chopper is switched ON, the inductor L becomes charged by the source voltage V_s. Therefore, $V_s = V_L$.

$$Vs = L\,\frac{di}{dt}, \quad \frac{\Delta i}{Ton} = \frac{Vs}{L}$$

$$\Delta i = \frac{Vs}{L} Ton \times \frac{T}{T}$$

$$\Delta i = \frac{DVs}{Lf}$$

,

When the chopper is switched OFF, the inductor's polarity reverses and this causes it to discharge through the diode and the load.

Hence,

$$V0 = -VL$$

$$L\,\frac{di}{dt} = -VL$$

$$\frac{L\Delta i}{Toff} = -VL$$

$$\Delta i = -\frac{VL\,Toff}{L}$$

By comparing the above equations

$$\frac{DVs}{Lf} = -\frac{VL\,Toff}{L}$$

$$V_o = \frac{DVs}{1-D}$$

Principle of operation of class A chopper

Class A Chopper is a first quadrant chopper

• When chopper is ON, supply voltage V is connected across the load.

• When chopper is OFF, vO = 0 and the load current continues to flow in the same direction through the FWD.

• The average values of output voltage and current are always positive. Class A Chopper is a first quadrant chopper

• When chopper is ON, supply voltage V is connected across the load.

• When chopper is OFF, vO = 0 and the load current continues to flow in the same direction through the FWD.

• The average values of output voltage and current are always positive.

• Class A Chopper is a step-down chopper in which power always flows form source to load.

• It is used to control the speed of dc motor.

• The output current equations obtained in step down chopper with R-L load can be used to study the performance of Class A Chopper.

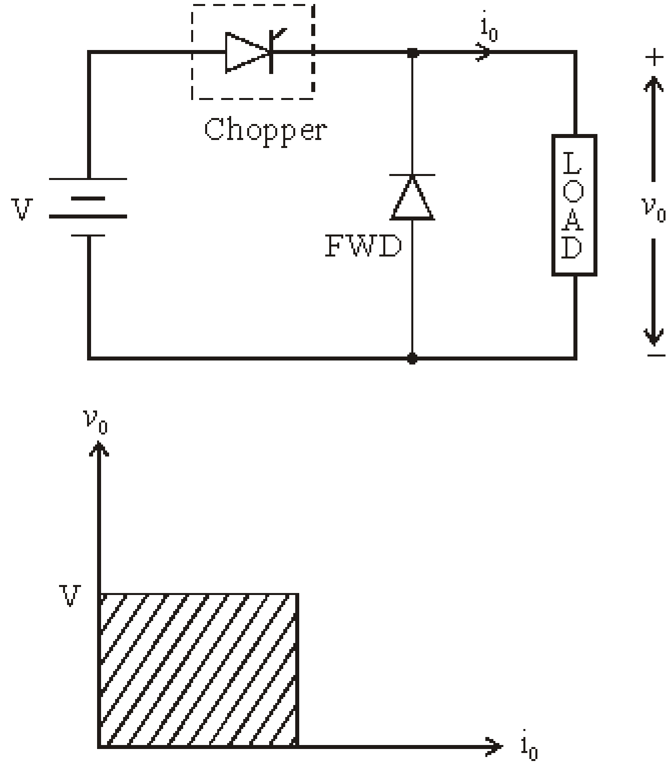

Figure: 3.10 circuit diagram and quadrant operation of Type A chopper

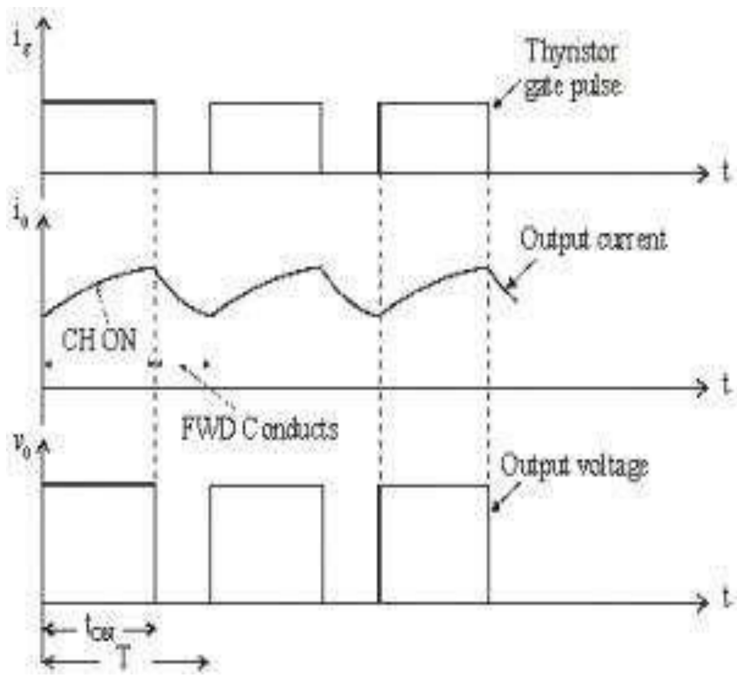

Figure: 3.11 Output voltage and current waveforms of type A chopper

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Voltage equation for the circuit shown in figure is

$$V = i_O R + L\frac{di_O}{dt} + E$$

Taking Laplace Transform

$$\frac{V}{S} = RI_O(S) + L\left[S I_O(S) - i_O(0^-)\right] + \frac{E}{S}$$

At $t = 0$, initial current $i_O(0^-) = I_{\min}$

$$I_O(S) = \frac{V-E}{LS\left(S+\frac{R}{L}\right)} + \frac{I_{\min}}{S+\frac{R}{L}}$$

Taking Inverse Laplace Transform

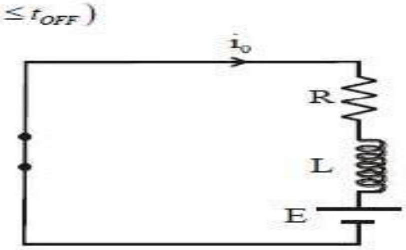
$$i_O(t) = \frac{V-E}{R}\left[1 - e^{-\left(\frac{R}{L}\right)t}\right] + I_{\min} e^{-\left(\frac{R}{L}\right)t}$$

This expression is valid for $0 \le t \le t_{ON}$. i.e., during the period chopper is ON.

At the instant the chopper is turned off, load current is

$$i_O(t_{ON}) = I_{\max}$$

When Chopper is OFF $(0 \le t \le t_{OFF})$



Voltage equation for the circuit shown in figure is

$$0 = Ri_O + L\frac{di_O}{dt} + E$$

Taking Laplace transform

$$0 = RI_O(S) + L\left[SI_O(S) - i_O(0^-)\right] + \frac{E}{S}$$

Redefining time origin we have at $t = 0$, initial current $i_O(0^-) = I_{\max}$

Therefore $$I_O(S) = \frac{I_{\max}}{S+\frac{R}{L}} - \frac{E}{LS\left(S+\frac{R}{L}\right)}$$

PF Taking Inverse Laplace Transform

$$i_O(t) = I_{\max} e^{-\frac{R}{L}t} - \frac{E}{R}\left[1 - e^{-\frac{R}{L}t}\right]$$

The expression is valid for $0 \le t \le t_{OFF}$, i.e., during the period chopper is OFF. At the instant the chopper is turned ON or at the end of the off period, the load current is

$$i_O(t_{OFF}) = I_{\min}$$

TO FIND $I_{\max}$ AND $I_{\min}$

At $\quad t = t_{ON} = dT, \quad i_O(t) = I_{\max}$

Class B Chopper

Class B Chopper is a step-up chopper

- When chopper is ON, E drives a current through L and R in a direction opposite to that shown in figure.
- During the ON period of the chopper, the inductance L stores energy.
- When Chopper is OFF, diode D conducts, and part of the energy stored in inductor L is returned to the supply.
- Average output voltage is positive. Average output current is negative.
- Therefore Class B Chopper operates in second quadrant.
- In this chopper, power flows from load to source.
- Class B Chopper is used for regenerative braking of dc motor.

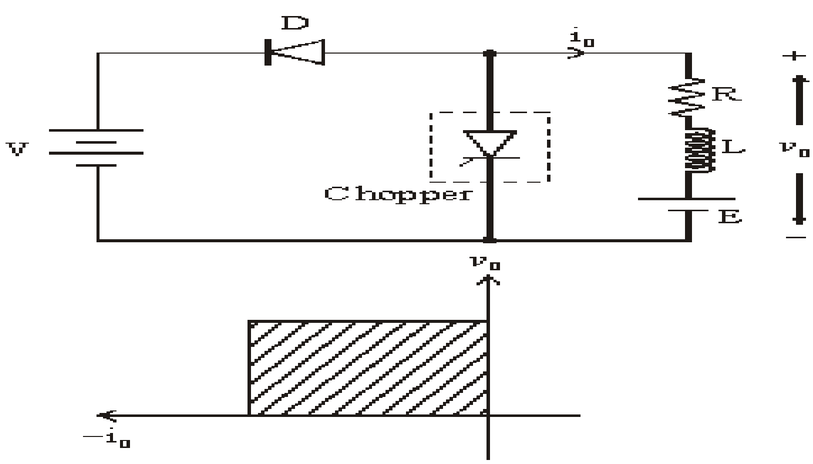


Figure: 3.12 circuit diagram and quadrant operation of Type B chopper

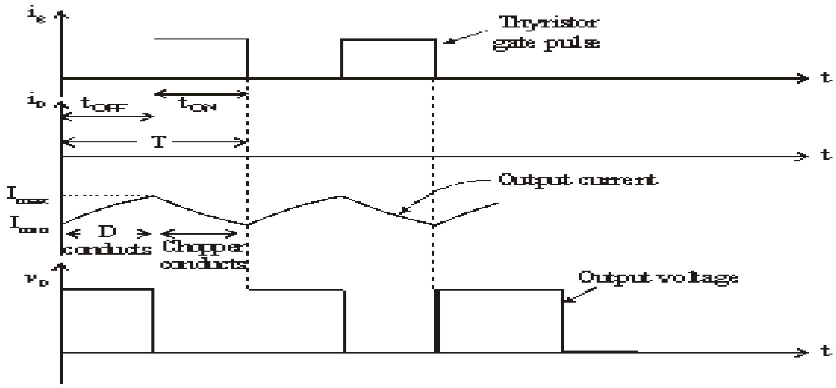


Figure: 3.13 Output voltage and current waveforms of type B chopper Class

C chopper

Class C Chopper can be used as a step-up or step-down chopper

• Class C Chopper is a combination of Class A and Class B Choppers.

• For first quadrant operation, CH1 is ON or D2 conducts.

• For second quadrant operation, CH2 is ON or D1 conducts.

• When CH1 is ON, the load current is positive.

• The output voltage is equal to 'V' & the load receives power from the source.

• When CH1 is turned OFF, energy stored in inductance L forces current to flow through the diode D2 and the output voltage is zero.

• Current continues to flow in positive direction.

• When CH2 is triggered, the voltage E forces current to flow in opposite direction through L and CH2 .

• The output voltage is zero.

• On turning OFF CH2 , the energy stored in the inductance drives current through diode D1 and the supply

• Output voltage is V, the input current becomes negative and power flows from load to source.

• Average output voltage is positive

• Average output current can take both positive and negative values.

• Choppers CH1 & CH2 should not be turned ON simultaneously as it would result in short circuiting the supply.

• Class C Chopper can be used both for dc motor control and regenerative braking of dc motor.

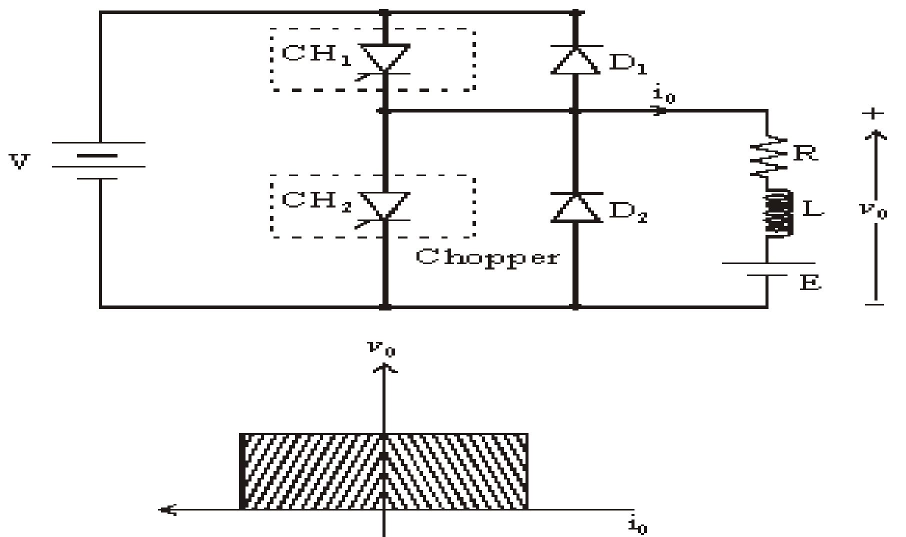


Figure: 3.14 circuit diagram and quadrant operation of Type C chopper

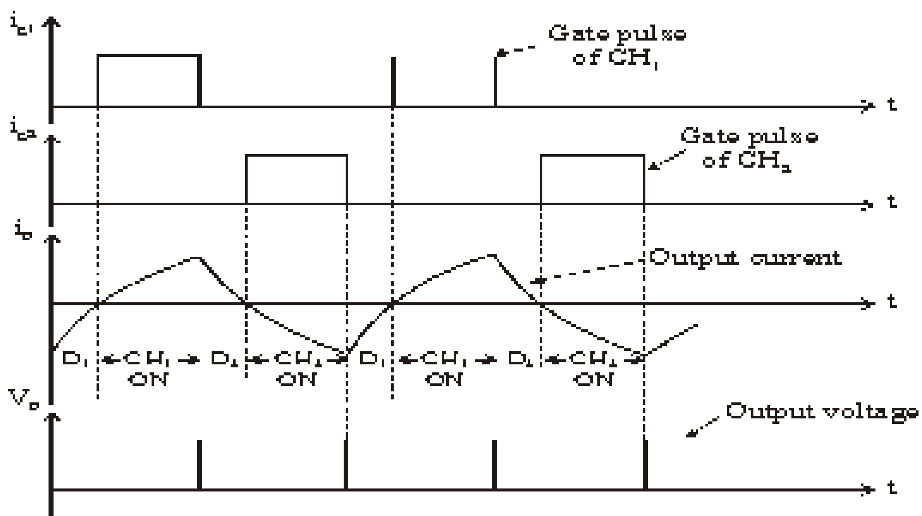


Figure: 3.15 Output voltage and current waveforms of type C chopper

Class D chopper

• Class D is a two quadrant chopper.

• When both CH1 and CH2 are triggered simultaneously, the output voltage vO = V and output current flows through the load.

• When CH1 and CH2 are turned OFF, the load current continues to flow in the same direction through load, D1 and D2 , due to the energy stored in the inductor L.

• Output voltage vO = – V.

• Average load voltage is positive if chopper ON time is more than the OFF time

• Average output voltage becomes negative if tON < tOFF .

• Hence the direction of load current is always positive but load voltage can be positive or negative.

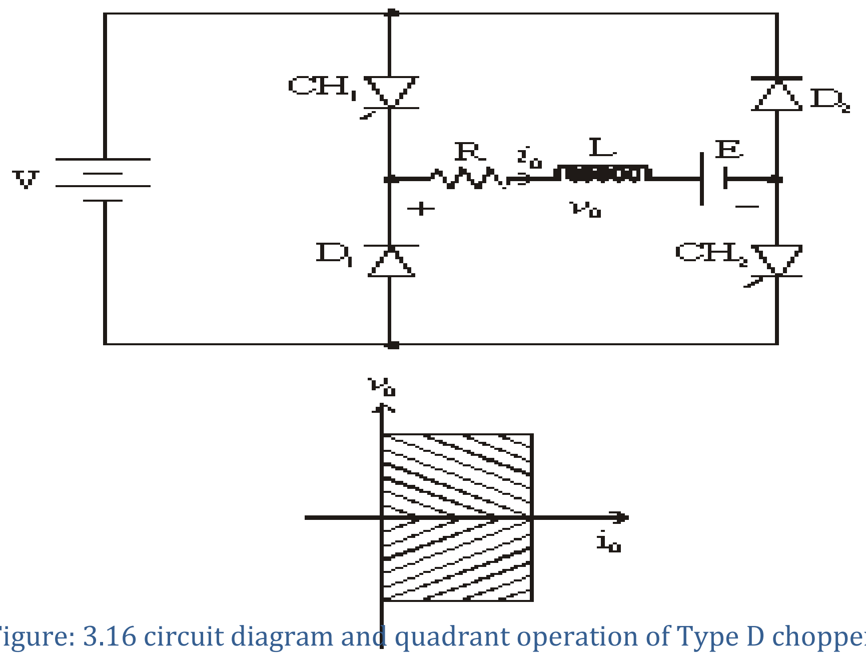


Figure: 3.16 circuit diagram and quadrant operation of Type D chopper

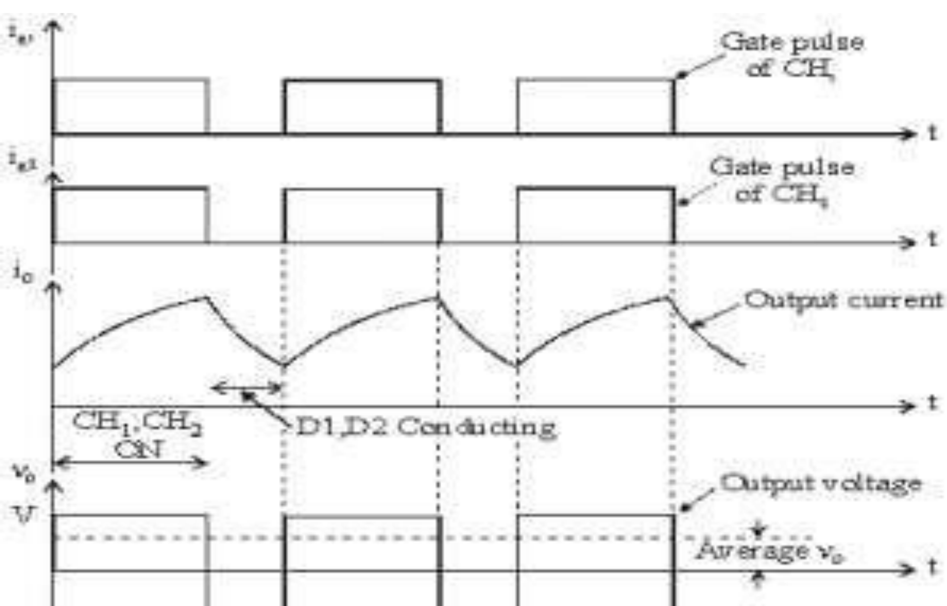


Figure: 3.17 Output voltage and current waveforms of type D chopper

Class E Chopper

- Class E is a four quadrant chopper
- When CH1 and CH4 are triggered, output current iO flows in positive direction through CH1 and CH4, and with output voltage vO = V.
- This gives the first quadrant operation.
- When both CH1 and CH4 are OFF, the energy stored in the inductor L drives iO through D2 and D3 in the same direction, but output voltage vO = -V.
- Therefore the chopper operates in the fourth quadrant.
- When CH2 and CH3 are triggered, the load current iO flows in opposite direction & output voltage vO = -V.
- Since both iO and vO are negative, the chopper operates in third quadrant.
- When both CH2 and CH3 are OFF, the load current iO continues to flow in the same direction D1 and D4 and the output voltage vO = V.
- Therefore the chopper operates in second quadrant as vO is positive but iO is negative.

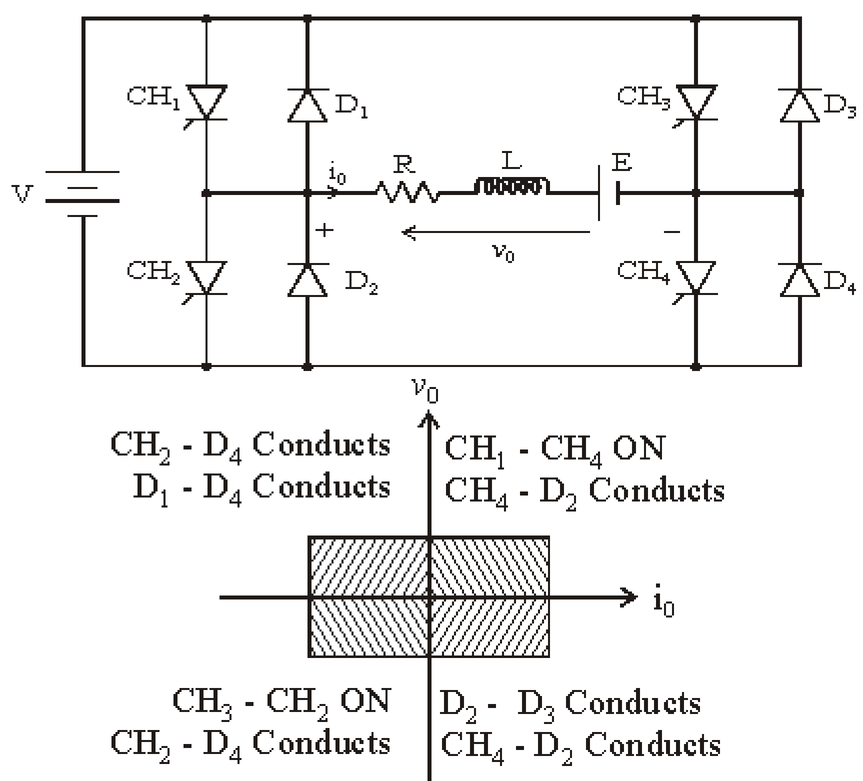


Figure: 3.18 circuit diagram and quadrant operation of Type E chopper

The word 'inverter' in the context of power-electronics denotes a class of power conversion (or power conditioning) circuits that operates from a dc voltage source or a dc current source and converts it into ac voltage or current. The inverter does reverse of what ac-to-dc converter does (refer to ac to dc converters). Even though input to an inverter circuit is a dc source, it is not uncommon to have this dc derived from an ac source such as utility ac supply. Thus, for example, the primary source of input power may be utility ac voltage supply that is converted to dc by an ac to dc converter and then 'inverted' back to ac using an inverter. Here, the final ac output may be of a different frequency and magnitude than the input ac of the utility supply

A single phase Half Bridge DC-AC inverter is shown in Figure below

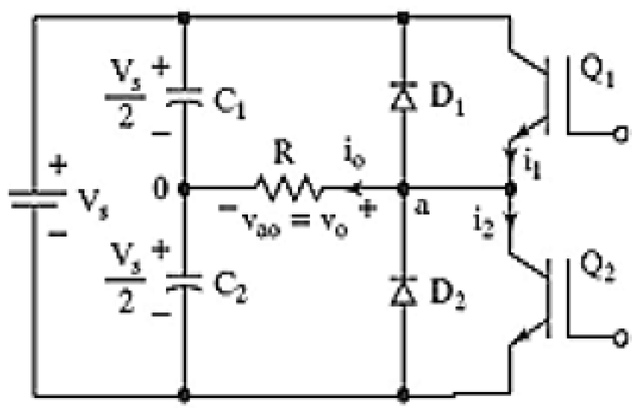


Figure: 5.1 Single phase Half Bridge DC-AC inverter with R load

The analysis of the DC-AC inverters is done taking into accounts the following assumptions and conventions.

1) The current entering node a is considered to be positive.

2) The switches S1 and S2 are unidirectional, i.e. they conduct current in one direction.

3) The current through S1 is denoted as i1 and the current through S2 is i2.

The switching sequence is so design is shown in Figure below. Here, switch S1 is on for the time duration $0 \leq t \leq T1$ and the switch S2 is on for the time duration $T1 \leq t \leq T2$. When switch S1 is turned on, the instantaneous voltage across the load is $v\ o = Vin/ 2$

When the switch S2 is only turned on, the voltage across the load is

$v\ o = Vin/ 2$.

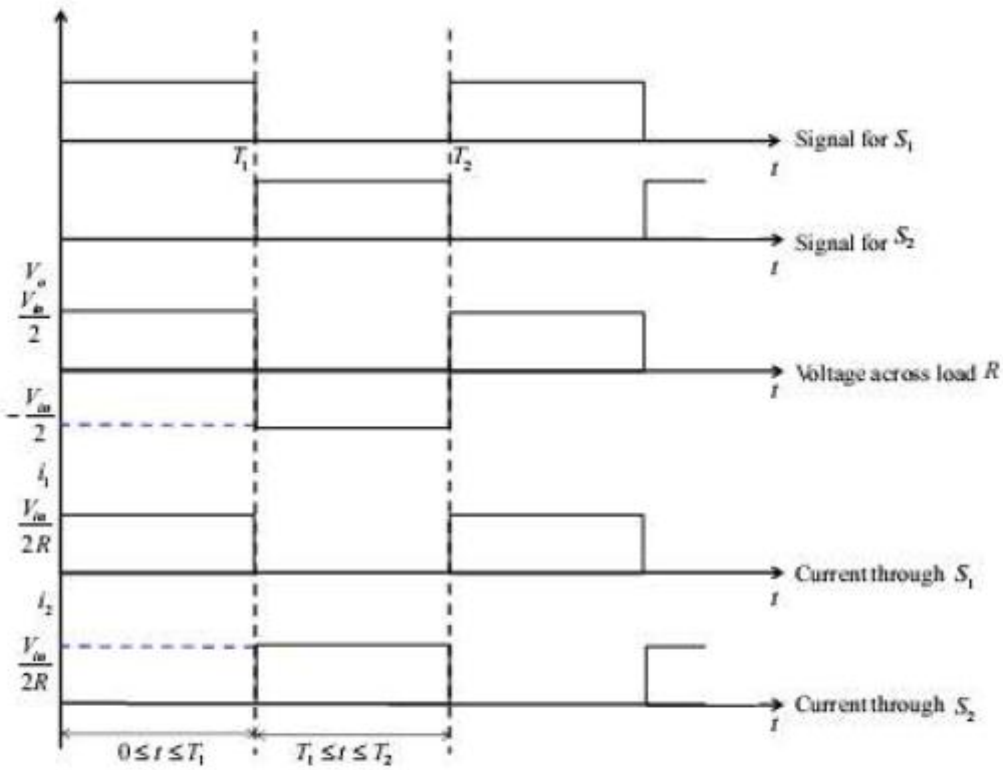


Figure: 5.2 Single phase Half Bridge DC-AC inverter output waveforms

The r.m.s value of output voltage ν o is given by,

$$V_{o,rms} = \left(\frac{1}{T_1} \int_0^{T_1} \frac{V_{in}^2}{4} dt \right) = \frac{V_{in}}{2}$$

The instantaneous output voltage ν o is rectangular in shape. The instantaneous value of ν o can be expressed in Fourier series as,

$$v_o = \frac{a_o}{2} + \sum_{n=1}^{\infty} a_n \cos(n\omega t) + b_n \sin(n\omega t)$$

Due to the quarter wave symmetry along the time axis , the values of a0 and an are zero. The

Value of bn is given by,

$$b_n = \frac{1}{\pi}\left[\int_{-\frac{\pi}{2}}^{0} \frac{-V_{in}}{2} d(\omega t) + \int_{0}^{\frac{\pi}{2}} \frac{V_{in}}{2} d(\omega t)\right] = \frac{2V_{in}}{n\pi}$$

Substituting the value of bn from above equation , we get

$$v_o = \sum_{n=1,3,5,...}^{\infty} \frac{2V_{in}}{n\pi} \sin(n\omega t)$$

The current through the resistor (iL) is given by,

$$i_L = \sum_{n=1,3,5,...}^{\infty} \frac{1}{R}\frac{2V_{in}}{n\pi} \sin(n\omega t)$$

Half Bridge DC-AC Inverter with L Load and R-L Load

The DC-AC converter with inductive load is shown in Figure below. For an inductive load, the load current cannot change immediately with the output voltage.

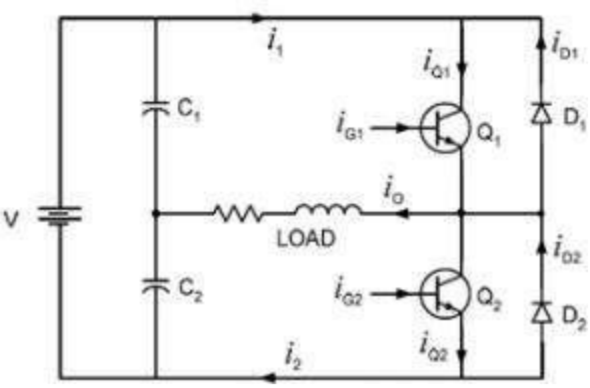
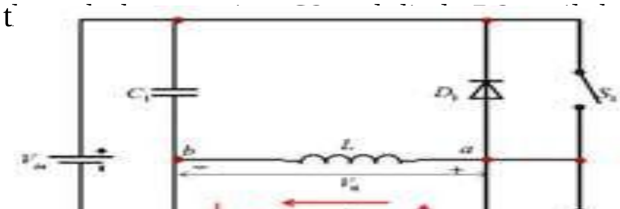


Figure: 5.3 Single phase Half Bridge DC-AC inverter with RL load

The working of the DC-AC inverter with inductive load is as follow is: Case 1: In the time interval 0<=t<= T1 the switch S1 is on and the current flows through the inductor from points a to b. When the switch S1 is turned off (case 1) at t-T1, the load current would continue to flow t … current falls to zero, as



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shown in Figure below.

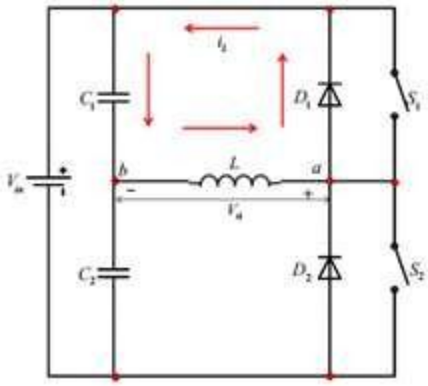


Figure: 5.4 Single phase Half Bridge DC-AC inverter with L load

$$i_L = \sum_{n=1,3,5,\ldots}^{\infty} \frac{1}{\omega nL} \frac{2V_{in}}{n\pi} \sin\left(n\omega t - \frac{\pi}{2}\right)$$

C... ed off at t = T1 , the load current flows through the diode D1 and capacitor C1until the current falls to zero, as shown in Figure below.

$$i_L = \sum_{n=1,3,5,\ldots}^{\infty} \frac{2V_{in}}{n\pi\sqrt{R^2 + (n\omega L)^2}} \sin(n\omega t - \theta_n)$$

$$\theta_n = \tan^{-1}\left(\frac{n\omega L}{R}\right)$$

Figure: 5.5 Single phase Half Bridge DC-AC inverter with L load

When the diodes D1 and D2 conduct, energy is feedback to the dc source and these diodes are known as feedback diodes. These diodes are also known as freewheeling diodes. The current for purely inductive load is given by,

Similarly, for the R – L load. The instantaneous load current is obtained as, Where,

Operation of single phase full bridge inverter

A single phase bridge DC-AC inverter is shown in Figure below. The analysis of the single phase DC-AC inverters is done taking into account following assumptions and conventions.

1) The current ... be positive.

2) The switches S1, S2, ... ion.

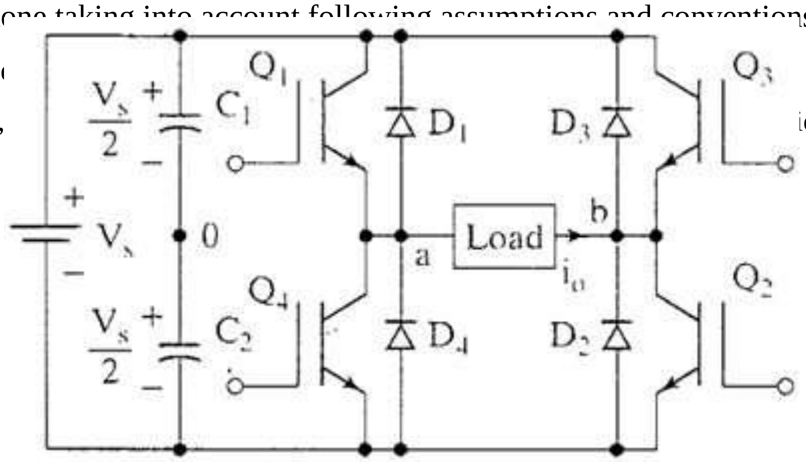


Figure: 5.6 Single phase Full Bridge DC-AC inverter with R load

When the switches S1 and S2 are turned on simultaneously for a duration $0 \leq t \leq T1$, the the input voltage Vin appears across the load and the current flows from point a to b.

Q1 – Q2 ON, Q3 – Q4 OFF =

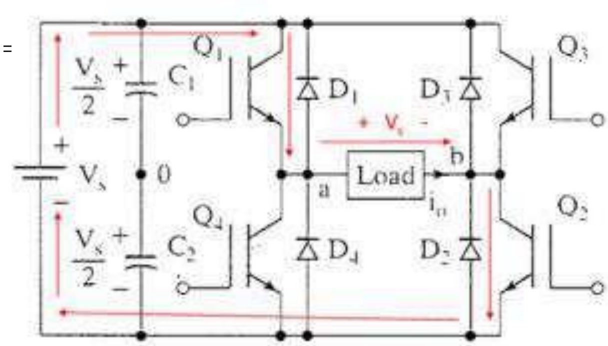


Figure: 5.7 Single phase Full Bridge DC-AC inverter with R load

If the switches S3 and S4 turned on duration T1 ≤ t ≤ T2, the voltage across the load the load is reversed

and the current through tł

to a. Q1 – Q2 OFF, Q3 –

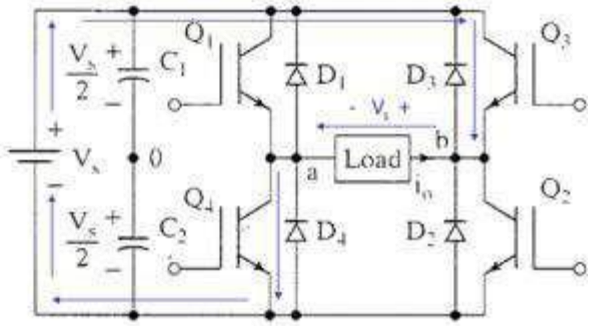


Figure: 5.8 Single phase Full Bridge DC-AC inverter with R load current directions

The voltage and current waveforms across the resistive load are shown in Figure below

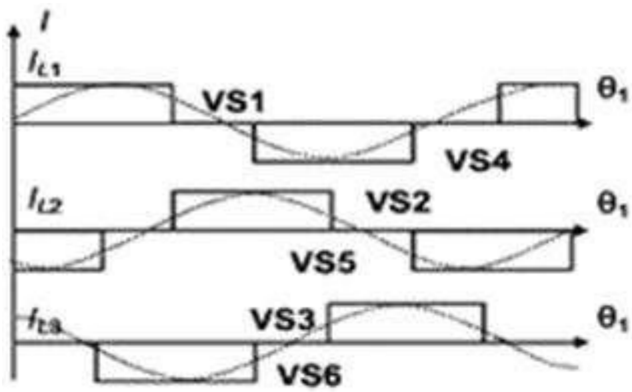


Figure: 5.9 Single phase Full Bridge DC-AC inverter waveforms

Single Phase Full Bridge Inverter for R-L load:

A single-phase square wave type voltage source inverter produces square shaped output voltage for a single-phase load. Such inverters have very simple control logic and the power switches need to operate at much lower frequencies compared to switches in some other types of inverters. The first generation inverters, using thyristor switches, were almost invariably square wave inverters because thyristor switches could be switched on and off only a few hundred times in a second. In contrast, the present day switches like IGBTs are much faster and used at switching frequencies of several kilohertz. Single-phase inverters mostly use half bridge or full

Bridge topologies. Power circuits of these topologies are shown in in Figure below.

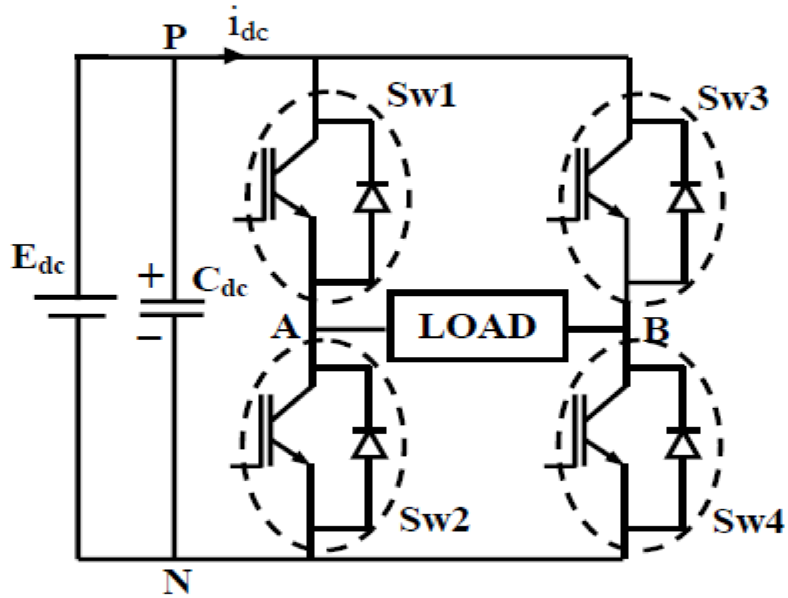
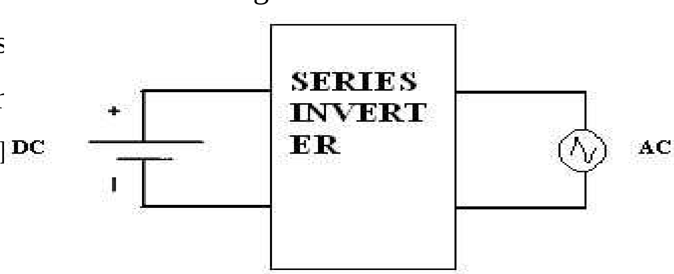


Figure: 5.10 Single phase Full Bridge DC-AC inverter with L load

The above topology is analyzed under the assumption of ideal circuit conditions. Accordingly, it is assumed that the input dc voltage (Edc) is constant and the switches are lossless. In full bridge topology has two such legs. Each leg of the inverter consists of two series connected electronic switches shown within dotted lines in the figures. Each of these switches consists of an IGBT type controlled switch across anner. These switches are capable of conductir ne polarity of voltage. The junction point of t e output point for the load.



Series inverter:

In **series inverter**, th … series with the load. This constitutes a ser … o produce the halves (positive and negativ …

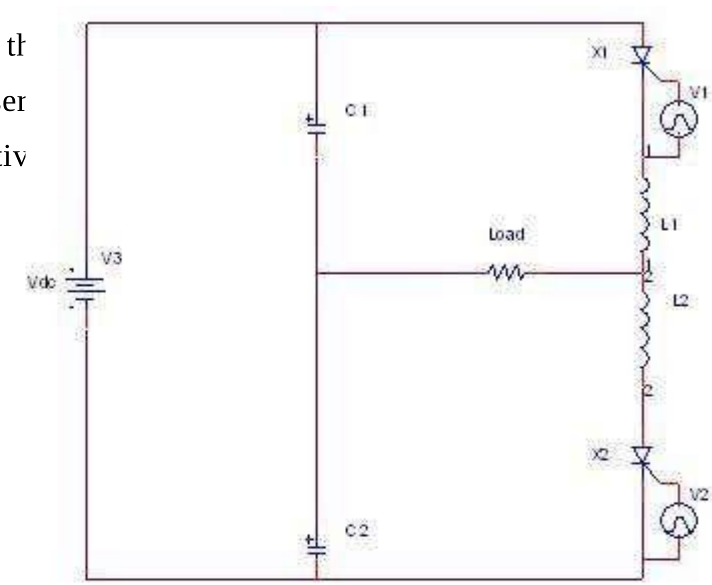


Figure: 5.12 Circuit diagram of series inverter

In the first half of the output currents when **SCR** T1 is triggered it will allow the current to flow through L1, and load, and C2 thus charging. The capacitor C1 which is already charged at these instant discharges through **SCR1**, L1 and the Load. Hence 50% of the current is drawn from the input source and 50% from the capacitor. Similarly in the second half of the output current C1 will be charged and C2 will discharge through the load, L2 and **SCR2**, Again 50% of the load current is obtained from the DC input source and rest from the capacitor. The **SCR**s T1 and T2 are alternatively fired to get AC voltage and current.

Operation of parallel inverter

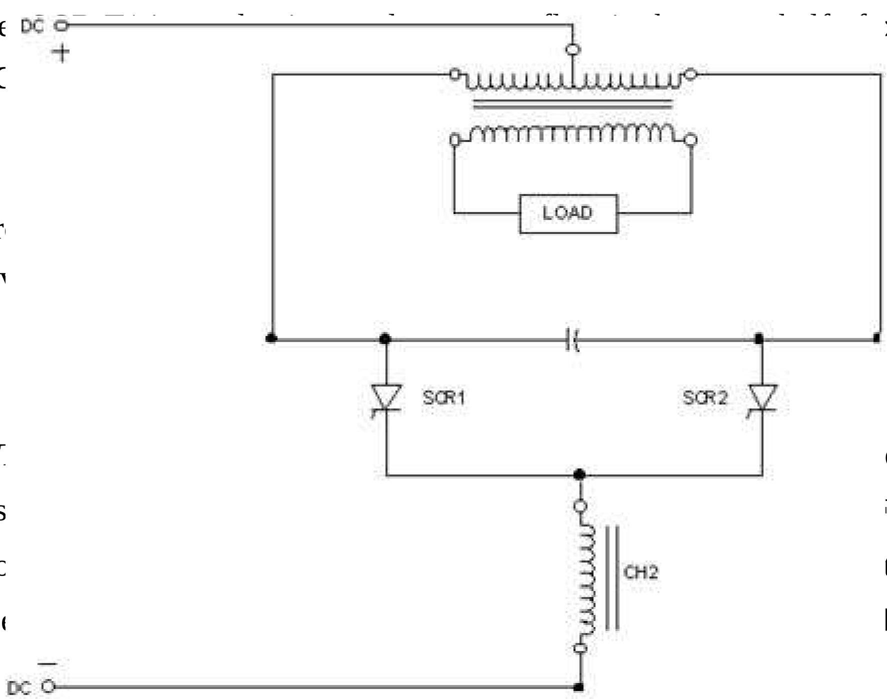
The **single phase parallel inverter circuit** consists of two **SCR**s T1 and T2, an inductor L, an output transformer and a commutating capacitor C. The output voltage and current are Vo and Io respectively. The function of L is to make the source current constant. During the working of this

inverter, capacitor C comes in **parallel** with the load via the transformer. So it is called a **parallel inverter**.

The operation of this inverter can be explained in the following

modes. Mode I

In this mode … primary
winding. **SC** … ell as lower
half of the



In other wor … charges to a
voltage of 2V

Mode II

At time to, … e t=0, capacitor
voltage 2Vs … . A current Io
begins to flo … tor has charged
(upper plat … hanges from Vs
at t=0 to –

Mode III

When capacitor has charged to –Vs, T1 may be tuned ON at any time When T1 is triggered, capacitor voltage 2Vs applies a reverse bias across T2, it is therefore turned OFF. After T2 is OFF, capacitor starts discharging, and charged to the opposite direction, the upper plate as positive.

Fig 1: is a schematic of the classical **parallel** commutated square wave inverter bridge. It is being included here for illustrative purposes since most other circuits utilize this circuit or a variation there of. The waveform generated and supplied to the load is basically a square wave having a peak to peak amplitude of twice the DC supply voltage and a period that is determined by the relate at which **SCR**s 1 through 4 are gated on. The **SCR**s are turned on in pairs by simultaneously applying signals to the gate terminals of **SCR**s 1 and 4 or **SCR**s 2 and 3. If **SCR**s 1 and 4 happen to be the first two switched on a current will flow from the positive terminal of the source through negative terminal of the source. This will establish a left to right, plus to minus voltage relationship on the load.

Simultaneously, the left terminal of capacitor C1 will be charged positively with respect to the right negative terminal. The steady-state load current through the various components is determined nearly completely by the impedance of the load. Chokes 1 and 2 and **SCR**s 1 and 4 present very low steady-state drops and therefore nearly all the source voltage appears across the load. Conduction of **SCR**s 1 and 4 will continue to the end of the half cycle, at which point the gates are removed from **SCR**s 1 and 4 remain in conduction along with **SCR**s 2 and 3 that have now been turned on. If it were not for chokes 1 and 2, the action of turning on the second set of **SCR**s would place very low impedance and therefore momentarily prevent the source from being short-circuited.Capacitor C1 now discharges with a current which flows into the cathode of **SCR** 1 through **SCR** 2 in a forward direction back to the negative terminal of the capacitor. This direction of current flow causes **SCR** 1 to become non-conductive provided that the reverse current through the **SCR** is of sufficient duration for the **SCR** to again become blocking. C1 simultaneously discharges through **SCR** 3 in a forward direction and through **SCR** 4 in a reverse direction. This will cause **SCR** 4 to become non-conductive just the same **SCR** 1. This entire sequence is referred to as commutation and typically in a modern inverter would occur in a period of time less than 50 microseconds. During this interval, chokes 1and 2 must have sufficient transient impedance to prevent a significant increase in current from the DC source.

Diodes 1, 2, 3 and 4 serve two functions. The first is to return any stored energy that may be "kicked back" from the load to the source. They also serve to prevent the choke from generating a high transient voltage immediately after commutation.

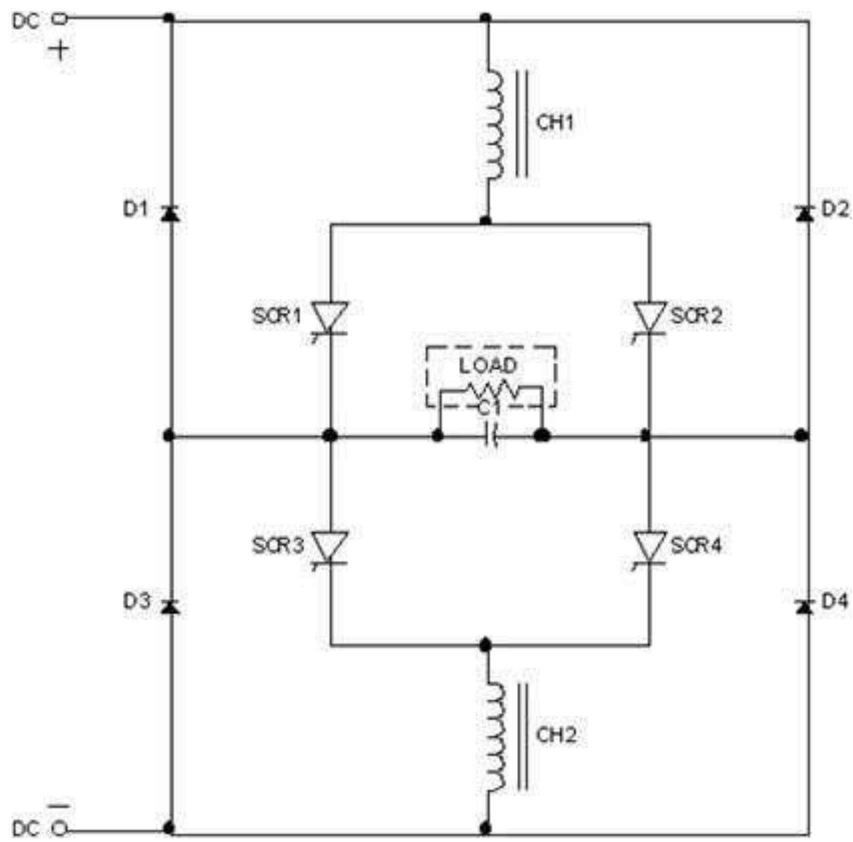
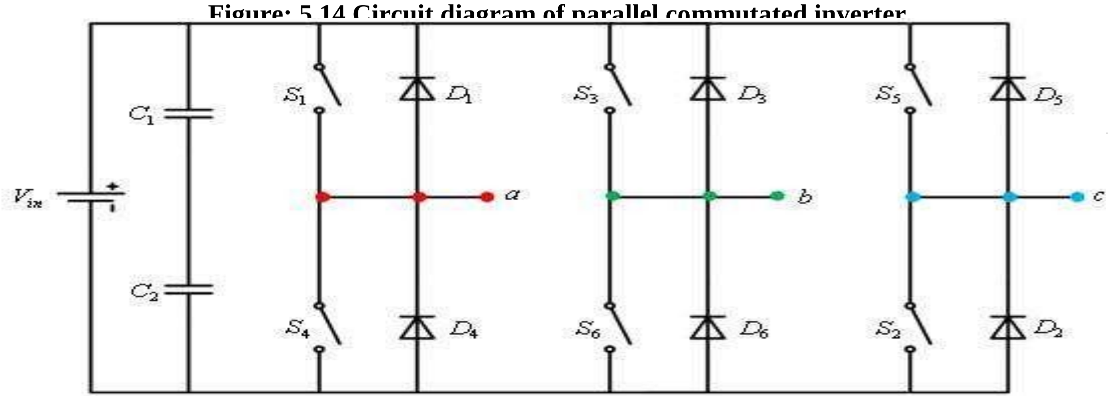


Figure: 5.14 Circuit diagram of parallel commutated inverter



Three phase inverters are normally used for high power applications. The advantages of a three phase inverter are:

- The frequency of the output voltage waveform depends on the switching rate of the swtiches and hence can be varied over a wide range.

- The direction of rotation of the motor can be reversed by changing the output phase sequence of the inverter.

- The ac output voltage can be controlled by varying the dc link voltage.

The general configuration of a three phase DC-AC inverter is shown in **Figure** Two types of control signals can be applied to the switches:

- 180° conduction

- 120° conduction

Figure: 5.15 Circuit diagram of three phase bridge inverter

180-Degree Conduction with Star Connected Resistive Load

The configuration of the three phase inverter with star connected resistive load is shown in **Figure.** The following convention is followed:

- A current leaving a node point ***a***, ***b*** or ***c*** and entering the neutral point ***n*** is assumed to be positive.

- All the three resistances are equal, $R_a = R_b = R_c = R$.

In this mode of operation each switch conducts for 180°. Hence, at any instant of time ***three switches*** remain ***on*** . When S_1 is ***on*** , the terminal ***a*** gets connected to the positive terminal of input DC source. Similarly, when S_4 is ***on*** , terminal ***a*** gets connected to the negative terminal of input DC source. There are six possible modes of operation in a cycle and each mode is of 60° duration and the explanation of each mode is as follows:

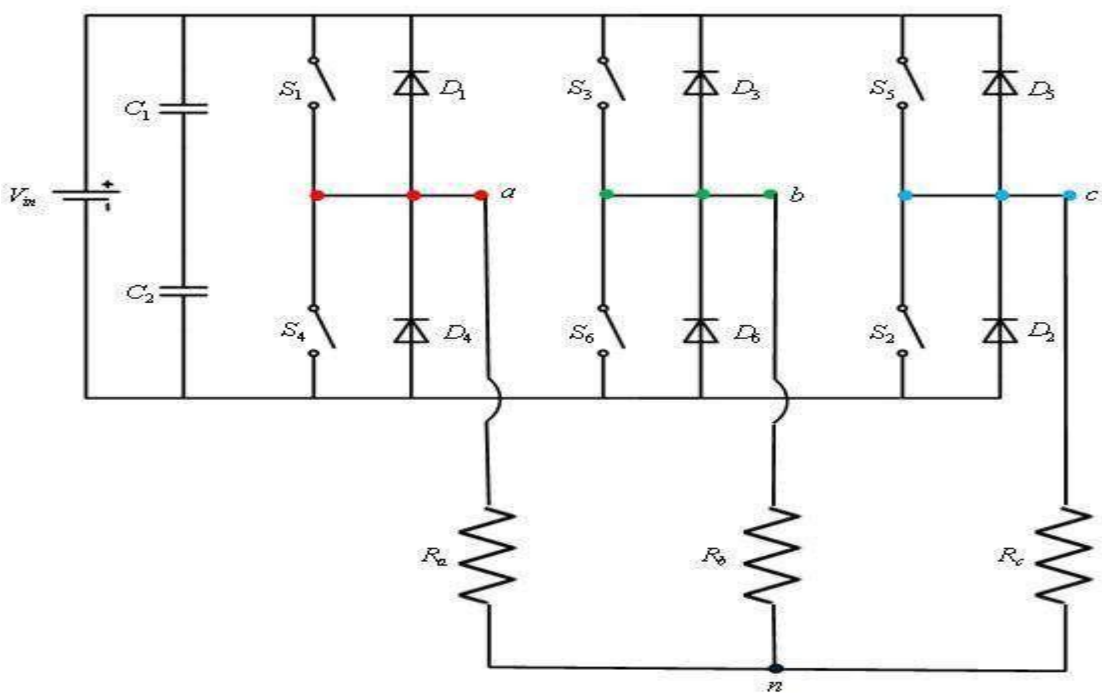


Figure: 5.16 Circuit diagram of three phase bridge inverter with star connected load

Mode 1 **:** In this mode the switches S_5 , S_6 and S_1 are turned ***on*** for time interval $0 \le \omega t \le \frac{\pi}{3}$. As a result of this the terminals ***a*** and ***c*** are connected to the positive terminal of the input DC source and the terminal ***b*** is connected to the negative terminal of the DC source. The current flow through R_a, R_b and R_c is shown in Figure and the equivalent circuit is shown in Figure**.** The equivalent resistance of the circuit shown in ***Figure*** is

$$R_{eq} = R + \frac{R}{2} = \frac{3R}{2} \tag{1}$$

The current *i* delivered by the DC input source is

$$i = \frac{V_{in}}{R_{eq}} = \frac{2}{3}\frac{V_{in}}{R} \tag{2}$$

The currents i_a and i_b are

$$i_a = i_c = \frac{1}{3}\frac{V_{in}}{R} \tag{3}$$

Keeping the current convention in mind, the current i_b is

$$i_b = -i = -\frac{2}{3}\frac{V_{in}}{R} \tag{4}$$

Having determined the currents through each branch, the voltage across each branch is

$$v_{an} = v_{cn} = i_a R = \frac{V_{in}}{3};\ v_{bn} = i_b R = -\frac{2V_{in}}{3} \tag{5}$$

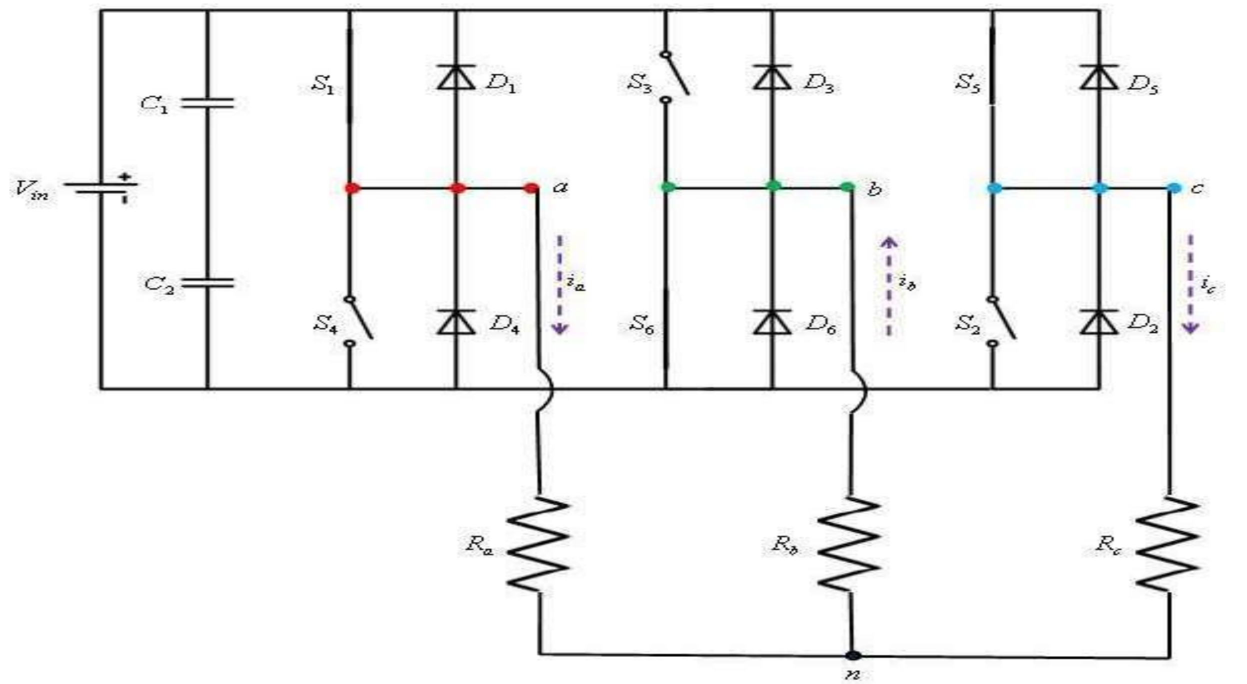


Figure: 5.17 Mode 1 operation of three phase bridge inverter with star connected load

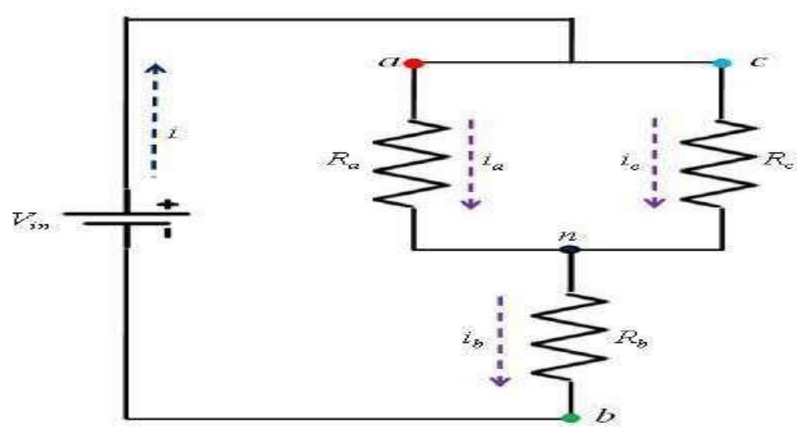


Figure: 5.18 Current flow in Mode 1 operation

Mode 2* :** In this mode the switches S_6 , S_1 and S_2 are turned ***on for time in $\frac{\pi}{3} \le \omega t \le \frac{2\pi}{3}$. The current flow and the equivalent circuits are shown in **Figure** and **Figure** respectively. Following the reasoning given for ***mode 1*** , the currents through each branch and the voltage drops are given by

$$i_b = i_c = \frac{1}{3}\frac{V_{in}}{R};\ i_a = -\frac{2}{3}\frac{V_{in}}{R} \qquad (6)$$

$$v_{bn} = v_{cn} = \frac{V_{in}}{3};\ v_{an} = -\frac{2V_{in}}{3} \qquad (7)$$

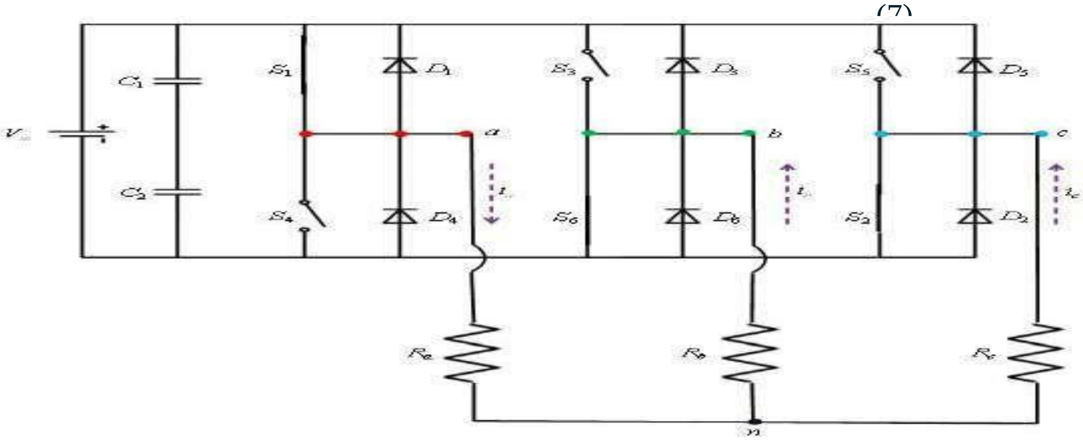


Figure [illegible] d load

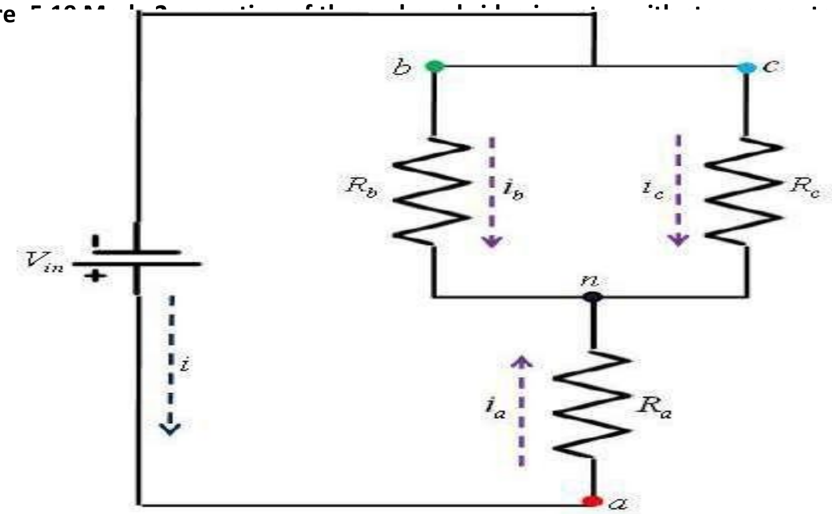


Figure: 5.20 Current flow in Mode 2 operation

Mode 3 **:** In this mode the switches S_1 , S_2 and S_3 are ***on*** for $\frac{2\pi}{3} \le \omega t \le \pi$. The current flow and the equivalent circuits are shown in **Figure** and **figure** respectively. The magnitudes of currents and voltages are:

$$i_a = i_b = \frac{1}{3}\frac{V_{in}}{R};\ i_c = -\frac{2}{3}\frac{V_{in}}{R} \qquad (8)$$

$$v_{an} = v_{bn} = \frac{V_{in}}{3};\ v_{cn} = -\frac{2V_{in}}{3} \qquad (9)$$

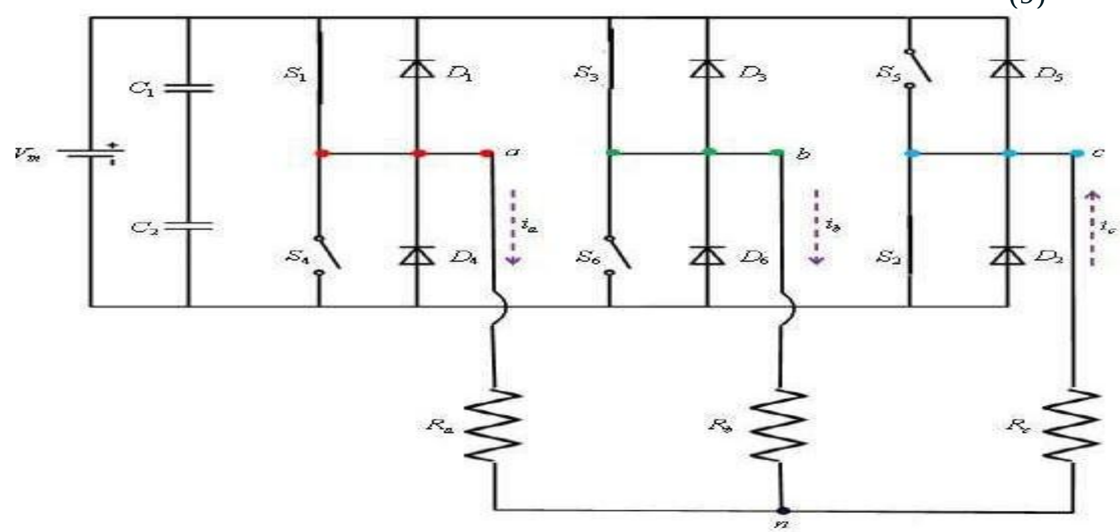


Figure: 5.21 Mode 3 operation of three phase bridge inverter with star connected load

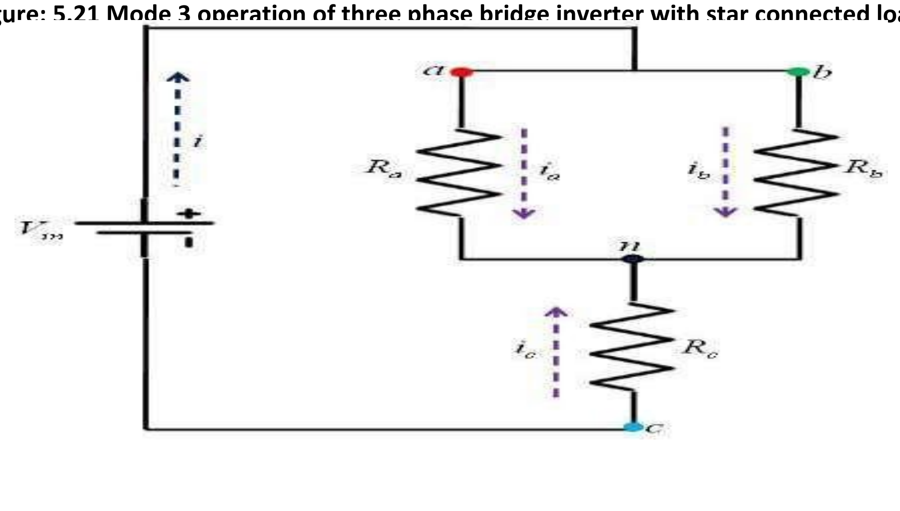


Figure: 5.23 Current flow in Mode 3 operation

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For ***modes 4, 5*** and ***6*** the equivalent circuits will be same as ***modes 1, 2*** and ***3*** respectively. The voltages and currents for each mode are:

$$\left.\begin{aligned} i_a = i_c = -\frac{1}{3}\frac{V_{in}}{R};\ i_b = \frac{2}{3}\frac{V_{in}}{R} \\ v_{an} = v_{cn} = -\frac{V_{in}}{3}; V_{bn} = \frac{2V_{in}}{3} \end{aligned}\right\} \text{ for } \textbf{\textit{mode 4}} \qquad (10)$$

$$\left.\begin{aligned} i_b = i_c = -\frac{1}{3}\frac{V_{in}}{R};\ i_a = \frac{2}{3}\frac{V_{in}}{R} \\ v_{bn} = v_{cn} = -\frac{V_{in}}{3}; V_{an} = \frac{2V_{in}}{3} \end{aligned}\right\} \text{ for } \textbf{\textit{mode5}} \qquad (11)$$

$$\left.\begin{aligned} i_a = i_b = -\frac{1}{3}\frac{V_{in}}{R};\ i_c = \frac{2}{3}\frac{V_{in}}{R} \\ v_{an} = v_{bn} = -\frac{V_{in}}{3}; V_{cn} = \frac{2V_{in}}{3} \end{aligned}\right\} \text{ for } \textbf{\textit{mode 6}} \qquad (12)$$

The plots of the phase voltages (v_{an}, v_{bn} and v_{cn}) and the currents (i_a, i_b and i_c) are shown in **Figure**
Having known the phase voltages, the line voltages can also be determined as:

$$\begin{aligned} v_{ab} &= v_{an} - v_{bn} \\ v_{bc} &= v_{bn} - v_{cn} \\ v_{ca} &= v_{cn} - v_{an} \end{aligned} \qquad (13)$$

The plots of line voltages are also shown in **Figure** and the phase and line voltages can be expressed in terms of Fourier series as:

$$v_{an} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{3n\pi}\left[1+\sin\frac{n\pi}{2}\sin\frac{n\pi}{6}\right]\sin\left(n\omega t\right)$$
$$v_{bn} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{3n\pi}\left[1+\sin\frac{n\pi}{2}\sin\frac{n\pi}{6}\right]\sin\left(n\omega t-\frac{2n\pi}{3}\right)$$
$$v_{cn} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{3n\pi}\left[1+\sin\frac{n\pi}{2}\sin\frac{n\pi}{6}\right]\sin\left(n\omega t-\frac{4n\pi}{3}\right)$$

(14)

$$v_{ab} = v_{an} - v_{bn} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{n\pi}\sin\frac{n\pi}{2}\sin\frac{n\pi}{3}\sin\left(n\omega t+\frac{n\pi}{6}\right)$$
$$v_{bc} = v_{bn} - v_{cn} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{n\pi}\sin\frac{n\pi}{2}\sin\frac{n\pi}{3}\sin\left(n\omega t-\frac{n\pi}{2}\right)$$
$$v_{ca} = v_{cn} - v_{an} = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4V_{in}}{n\pi}\sin\frac{n\pi}{2}\sin\frac{n\pi}{3}\sin\left(n\omega t-\frac{7n\pi}{6}\right)$$

(15)

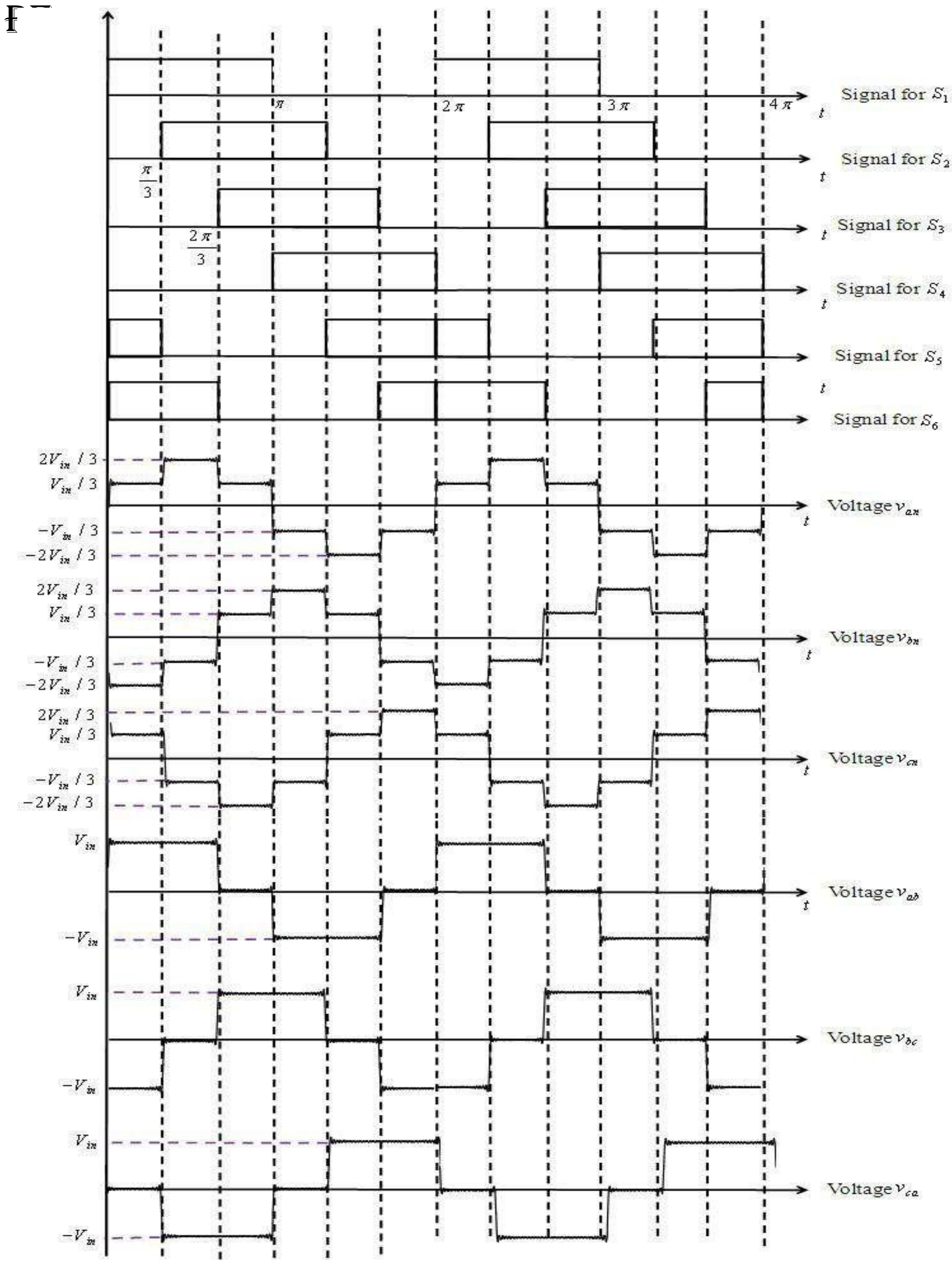

Figure: 5.24 Line and phase voltages of three phase bridge inverter

Three Phase DC-AC Converters with 120 degree conduction mode

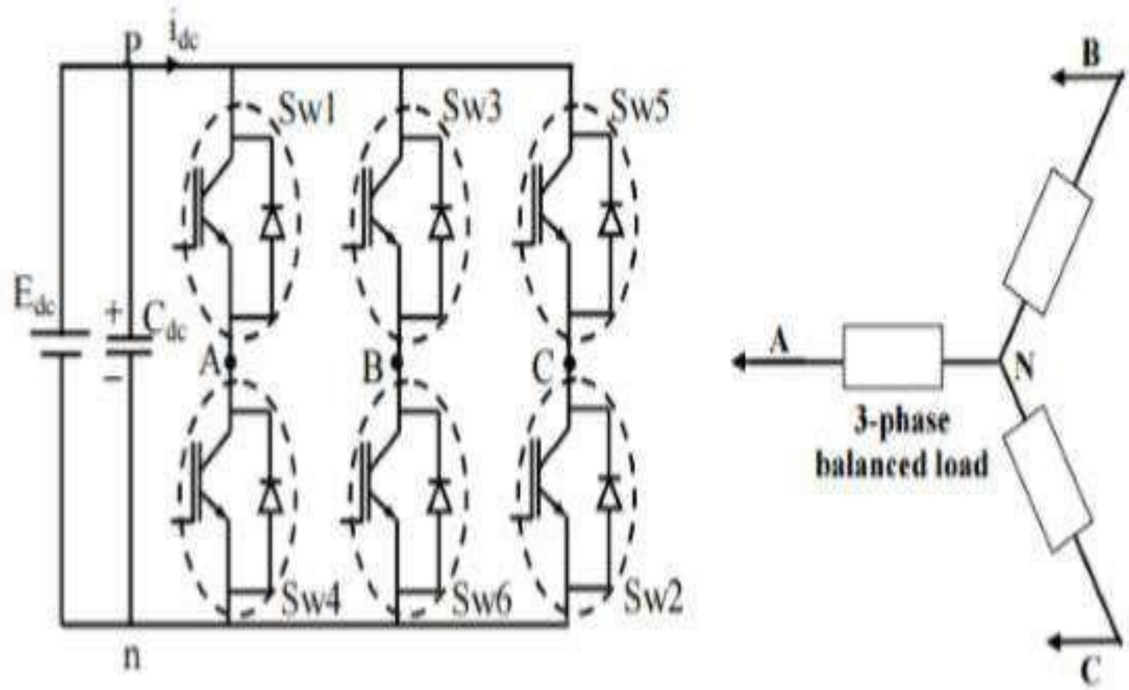

Figure: 5.25 Circuit diagram of three phase bridge inverter

120° mode of conduction

In this mode of conduction, each electronic device is in a conduction state for 120°. It is most suitable for a delta connection in a load because it results in a six-step type of waveform across any of its phases. Therefore, at any instant only two devices are conducting because each device conducts at only 120°.

The terminal A on the load is connected to the positive end while the terminal B is connected to the negative end of the source. The terminal C on the load is in a condition called floating state. Furthermore, the phase voltages are equal to the load voltages as shown below.

Phase voltages = Line

voltages $V_{AB} = V$

$V_{BC} = -V/2$

$V_{CA} = -V/2$

PE

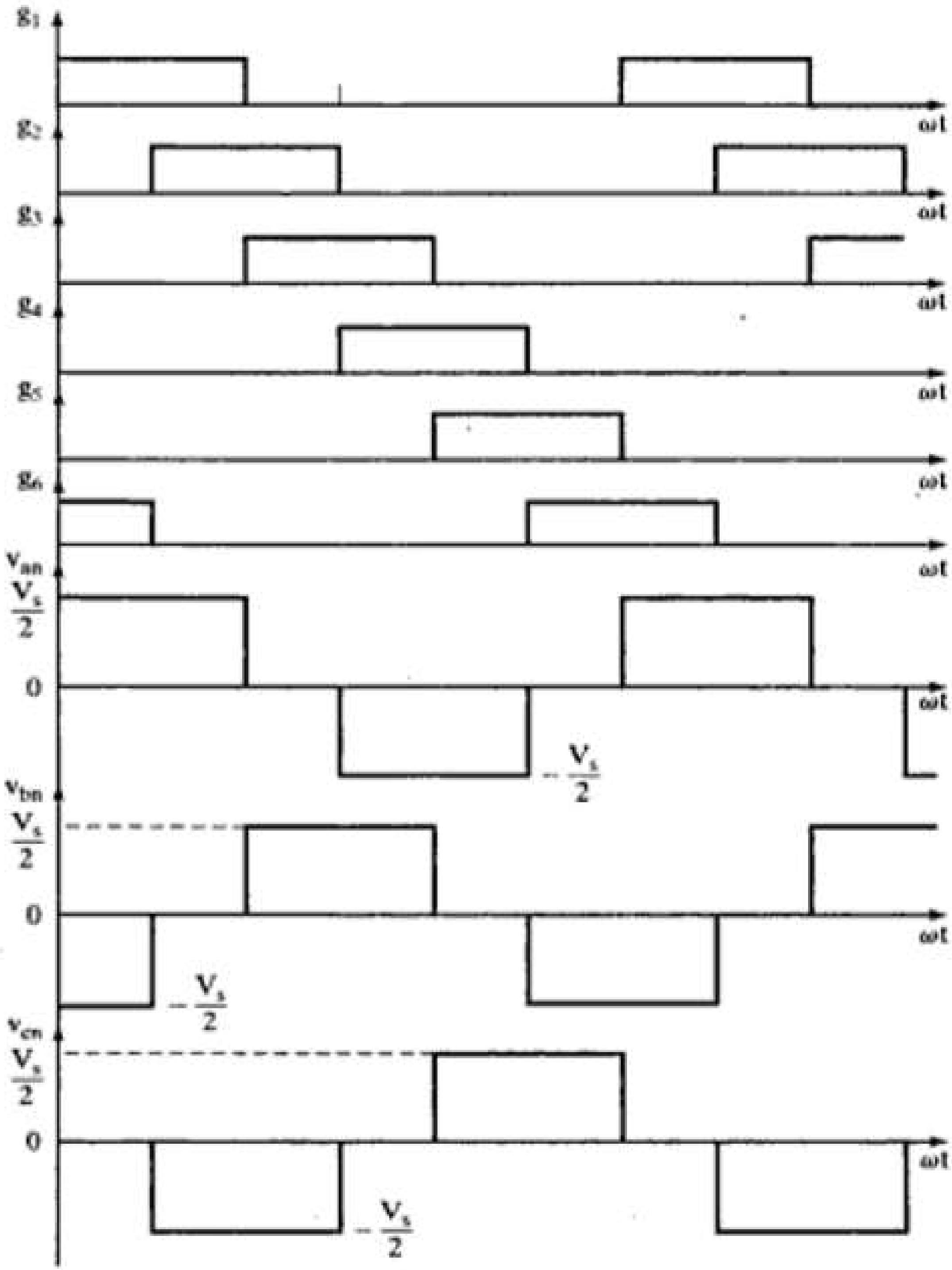

Figure: 5.26 Line and phase voltages of three phase bridge inverter

Voltage control techniques for inverters

Pulse width modulation techniques

PWM is a technique that is used to reduce the overall harmonic distortion (THD) in a load current. It uses a pulse wave in rectangular/square form that results in a variable average waveform value f(t), after its pulse width has been modulated. The time period for modulation is given by T. Therefore, waveform average value is given by

–

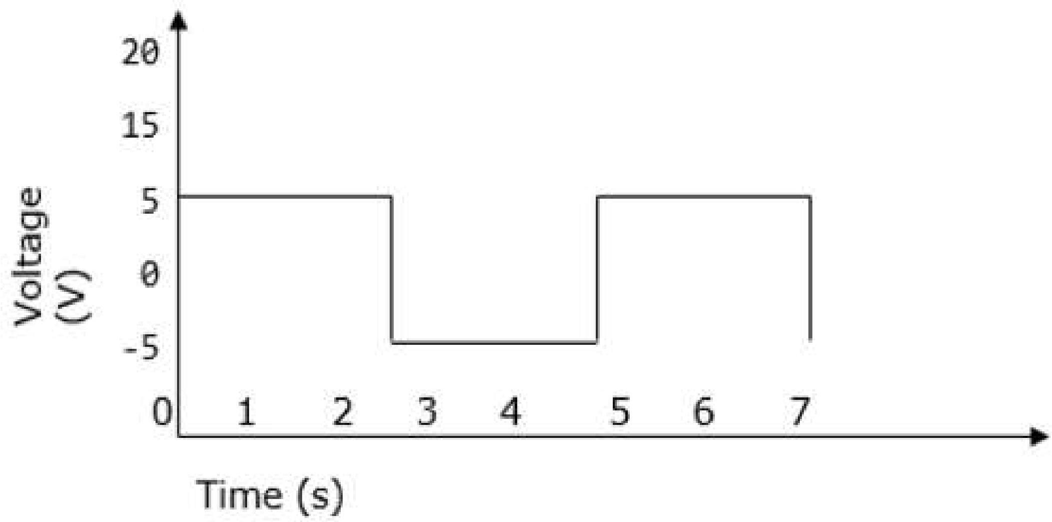

Figure: 5.27 Square waveform used for PWM technique

Sinusoidal Pulse Width Modulation

In a simple source voltage inverter, the switches can be turned ON and OFF as needed. During each cycle, the switch is turned on or off once. This results in a square waveform. However, if the switch is turned on for a number of times, a harmonic profile that is improved waveform is obtained.

The sinusoidal PWM waveform is obtained by comparing the desired modulated waveform with a triangular waveform of high frequency. Regardless of whether the voltage of the signal is smaller or larger than that of the carrier waveform, the resulting output voltage of the DC bus is either negative or positive.

PE

PF

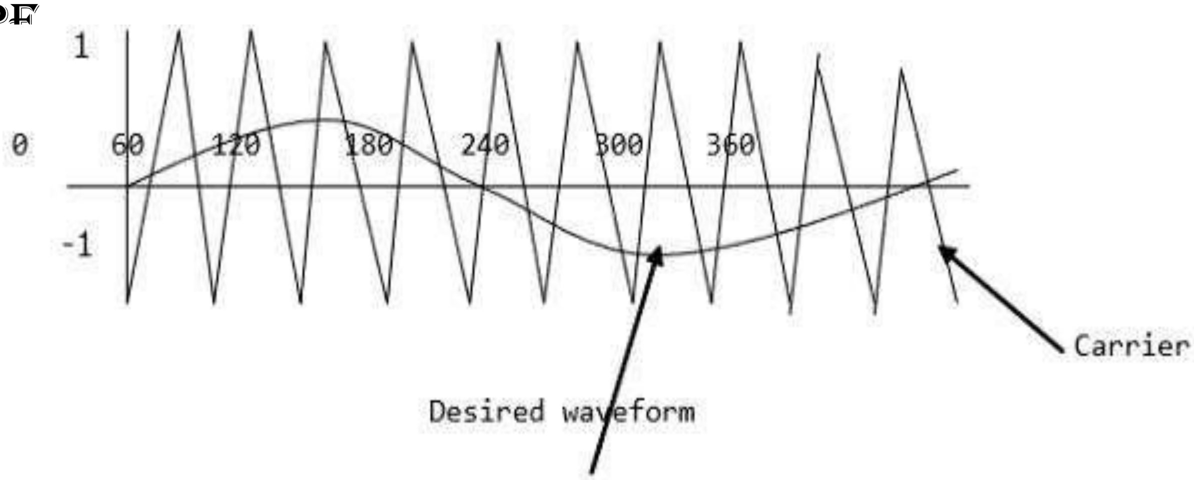

Figure: 5.28 Sinusoidal PWM waveform

The sinusoidal amplitude is given as A_m and that of the carrier triangle is give as A_c. For sinusoidal PWM, the modulating index m is given by A_m/A_c.

Modified Sinusoidal Waveform PWM

A modified sinusoidal PWM waveform is used for power control and optimization of the power factor. The main concept is to shift current delayed on the grid to the voltage grid by modifying the PWM converter. Consequently, there is an improvement in the efficiency of power as well as optimization in power factor.

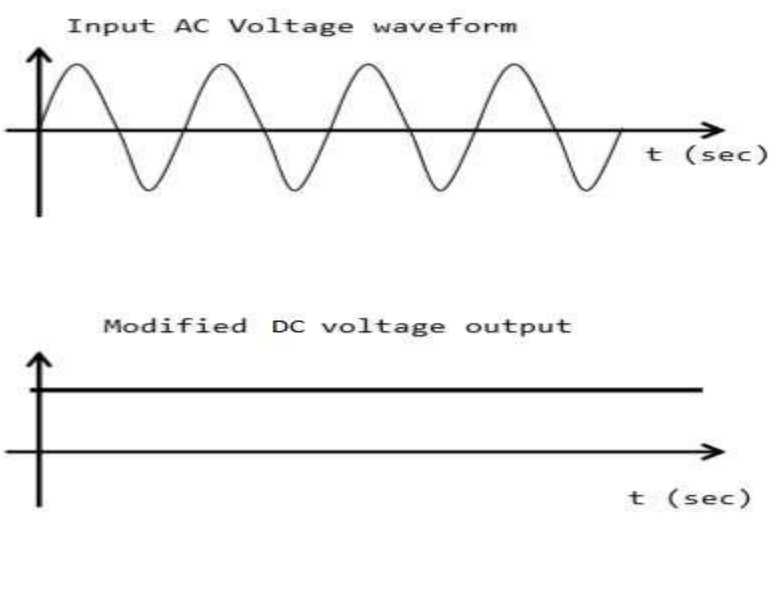

Figure: 5.29 Modified sinusoidal PWM waveform

Multiple PWM

The multiple PWM has numerous outputs that are not the same in value but the time period over which they are produced is constant for all outputs. Inverters with PWM are able to operate at high voltage output.

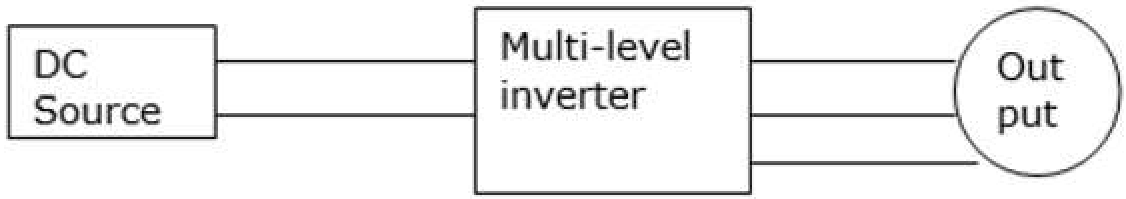

Figure: 5.30 Block diagram of multiple PWM technique

The waveform below is a sinusoidal wave produced by a multiple PWM

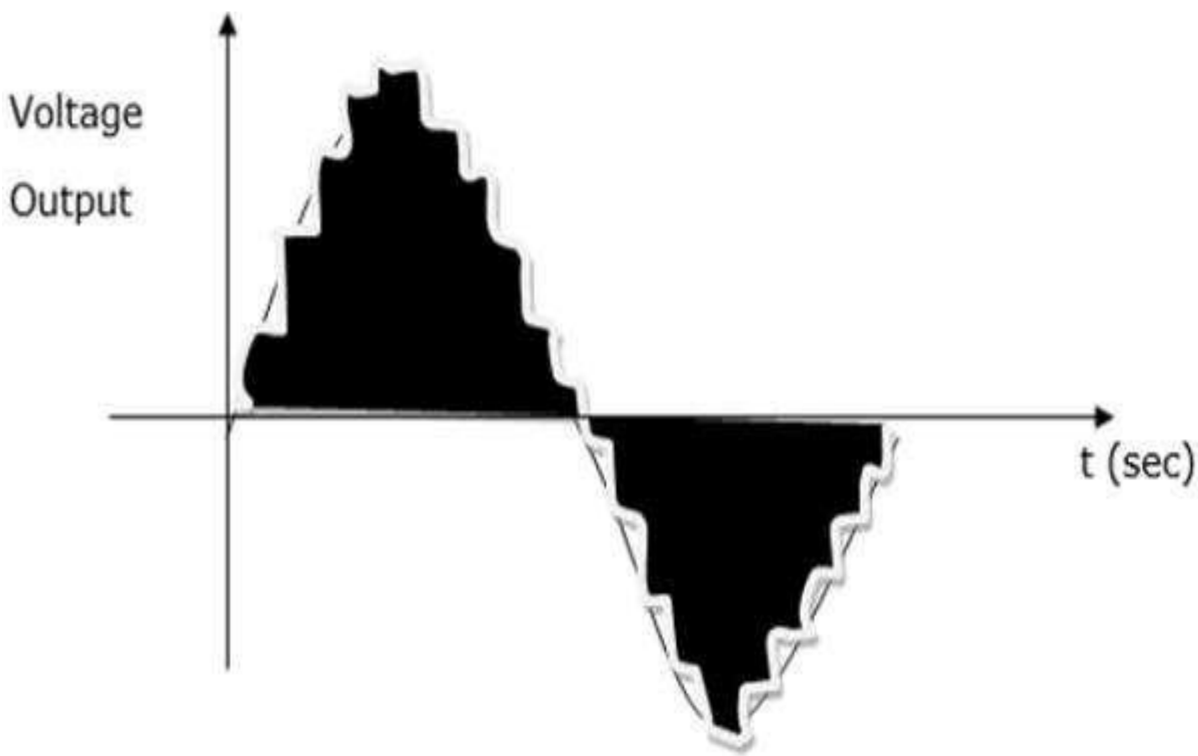

Figure: 5.31 Waveform of multiple PWM technique

A periodic waveform that has frequency, which is a multiple integral of the fundamental power with frequency of 60Hz is known as a harmonic. Total harmonic distortion (THD) on the other hand refers to the total contribution of all the harmonic current frequencies.

Harmonics are characterized by the pulse that represents the number of rectifiers used in a given circuit.

It is calculated as follows $h=(n\times P)+1 \text{ or } -1$

Where **n** − is an integer 1, 2, 3, 4….n

P − Number of rectifiers

Harmonics have an impact on the voltage and current output and can be reduced using isolation transformers, line reactors, redesign of power systems and harmonic filters.

Operation of sinusoidal pulse width modulation

The sinusoidal PWM (SPWM) method also known as the triangulation, sub harmonic, or sub oscillation method, is very popular in industrial applications. The SPWM is explained with reference to Figure, which is the half-bridge circuit topology for a single-phase inverter.

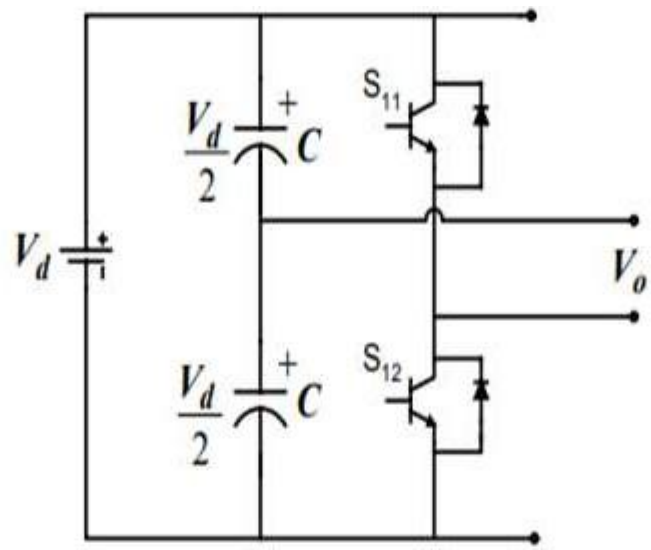

Figure: 5.32 schematic diagram of Half bridge PWM inverter

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For realizing SPWM, a high-frequency triangular carrier wave is compared with a sinusoidal reference of the desired frequency. The intersection of and waves determines the switching instants and commutation of the modulated pulse. The PWM scheme is illustrated in Figure, in which v_c the peak value of triangular carrier wave and v_r is that of the reference, or modulating signal. The figure shows the triangle and modulation signal with some arbitrary frequency and magnitude. In the inverter of Figure the switches and are controlled based on the comparison of control signal and the triangular wave which are mixed in a comparator. When sinusoidal wave has magnitude higher than the triangular wave the comparator output is high, otherwise it is low.

$$v_r > v_c \qquad S_{11} \text{ is on },\ V_{out} = \frac{V_d}{2}$$

and

$$v_r < v_c \qquad S_{12} \text{ is on },\ V_{out} = -\frac{V_d}{2}$$

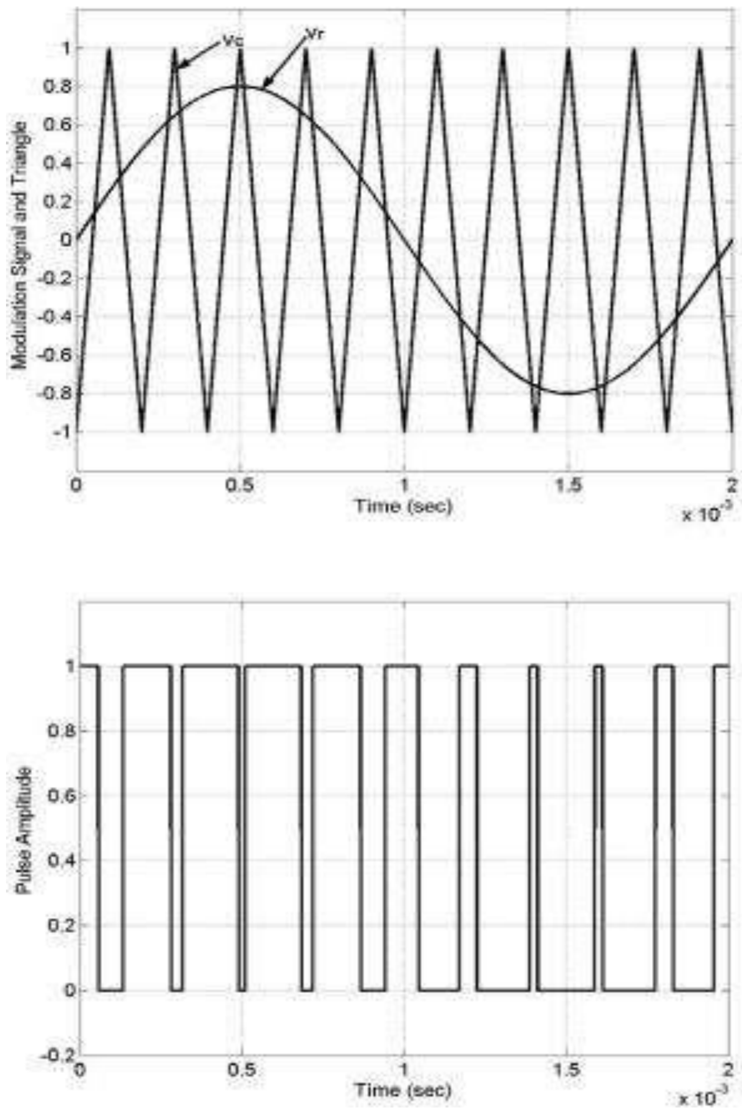

Figure: 5.33 Sine-Triangle Comparison and switching pulses of half bridge PWM inverter

The comparator output is processes 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. The magnitude 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 . The amplitude of the triangular wave is generally kept constant. The frequency modulation ratio is defined as

$M_F =$

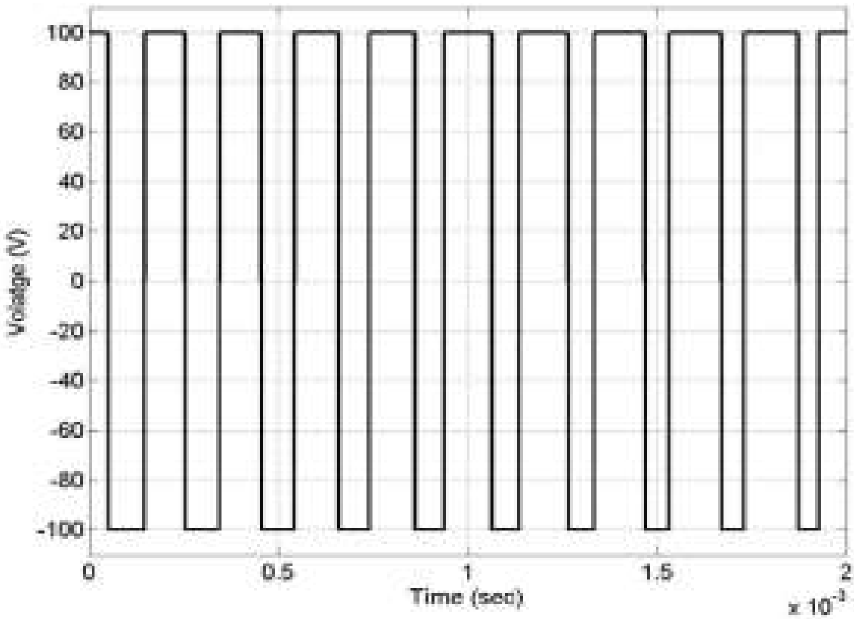

Figure: 5.34 Output voltage of the Half-Bridge inverter

Operation of current source inverter with ideal switches

Single-phase Current Source Inverter

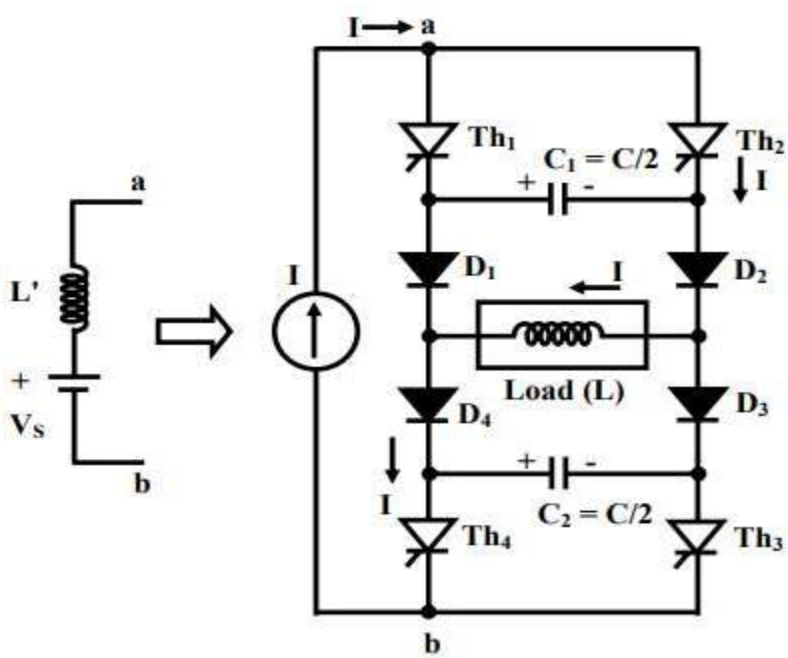

Figure: 5.35 Single phase current source inverter (CSI) of ASCI type

The circuit of a Single-phase Current Source Inverter (CSI) is shown in Fig. 5.35. The type of operation is termed as Auto-Sequential Commutated Inverter (ASCI). A constant current source is assumed here, which may be realized by using an inductance of suitable value, which must be high, in series with the current limited dc voltage source. The thyristor pairs, Th1 & Th3, and Th2 & Th4, are alternatively turned ON to obtain a nearly square wave current waveform. Two commutating capacitors − C1 in the upper half, and C2 in the lower half, are used. Four diodes, D1–D4 are connected in series with each thyristor to prevent the commutating capacitors from discharging into the load. The output frequency of the inverter is controlled in the usual way, i.e., by varying the half time period, (T/2), at which the thyristors in pair are triggered by pulses being fed to the respective gates by the control circuit, to turn them ON, as can be observed from the waveforms (Fig. 5.36). The inductance (L) is taken as the load in this case, the reason(s) for which need not be stated, being v y two modes.

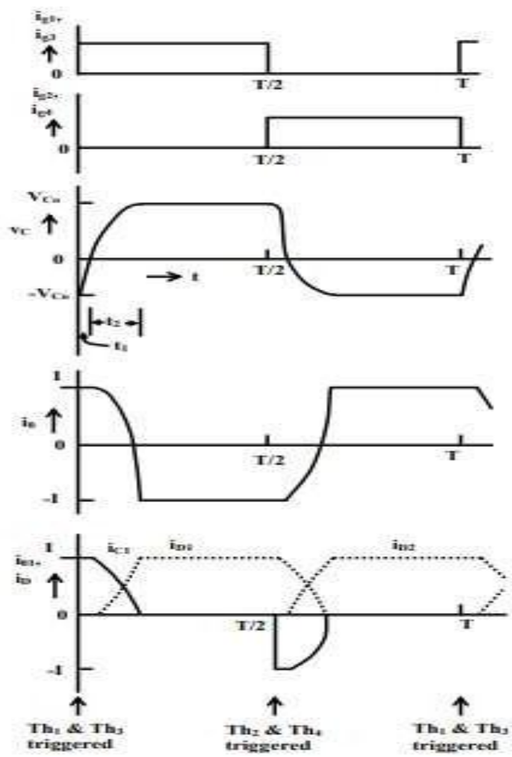

Figure: 5.36 output waveforms of Single phase current source inverter

Mode I: The circuit for this mode is shown in Fig. 5.37. The following are the assumptions. Starting from the instant, , the thyristor pair, Th − t = 0 2 & Th4, is conducting (ON), and the current (I) flows through the path, Th2, D2, load (L), D4, Th4, and source, I. The commutating capacitors are initially charged equally with the polarity as given, i.e., . This mans that both capacitors have right hand plate positive and left hand plate negative. If two capacitors are not charged initially, they have to pre-charge.

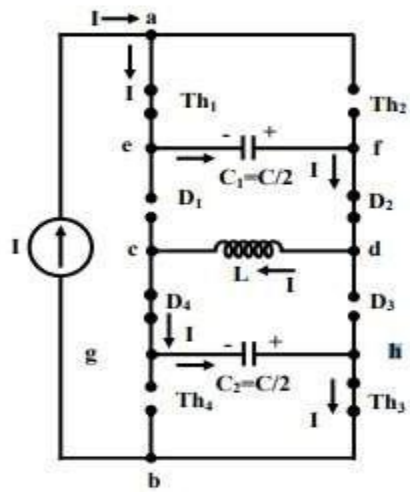

Figure: 5.37 Mode I operation of CSI

Mode II: The circuit for this mode is shown in Fig. 5.38. Diodes, D2 & D4, are already conducting, but at

= tt 1 , diodes, D1 & D3, get forward biased, and start conducting. Thus, at the end of time t1, all four diodes, D1–D4 conduct. As a result, the commutating capacitors now get connected in parallel with the load (L).

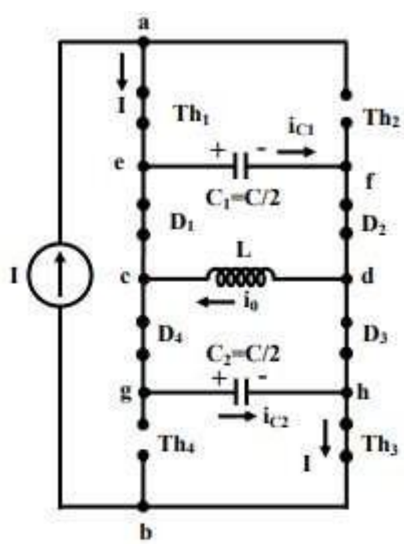

Figure: 5.38 Mode II operation of CSI

Load Commutated CSI

Two commutating capacitors, along with four diodes, are used in the circuit for commutation from one pair of thyristors to the second pair. Earlier, also in VSI, if the load is capacitive, it was shown that forced commutation may not be needed. The operation of a single-phase CSI with capacitive load (Fig. 5.39) is discussed here. It may be noted that the capacitor, C is assumed to be in parallel with resistive load (R). The capacitor, C is used for storing the charge, or voltage, to be used to force-commutate the conducting thyristor pair as will be shown. As was the case in the last lesson, a constant current source, or a voltage source with large inductance, is used as the input to the circuit.

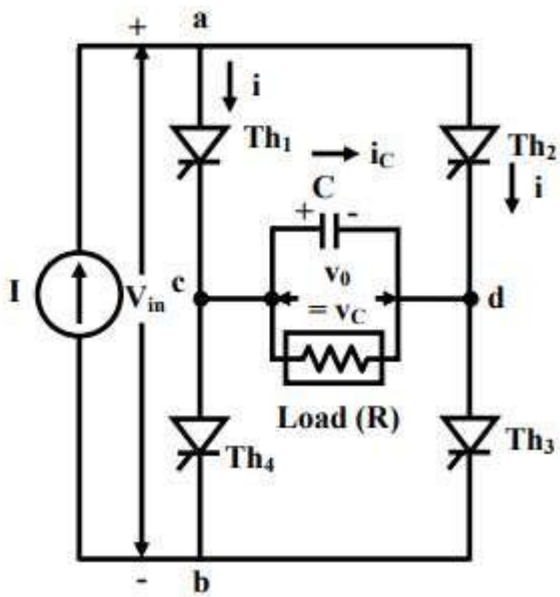

Figure: 5.39 Circuit diagram of load commutated CSI

The power switching devices used here is the same, i.e. four Thyristors only in a full- bridge configuration. The positive direction for load current and voltage is shown in Fig. 5.40 Before t = 0, the capacitor voltage is , i.e. the capacitor has left plate negative and right plate positive. At that time, the thyristor pair, Th2 & Th4 was conducting. When (at t = 0), the thyristor pair, Th1 & Th3 is triggered by the pulses fed at the gates, the conducting thyristor pair, Th2 & Th4 is reverse biased by the capacitor voltage C = −Vv 1 , and turns off immediately. The current path is through Th1, load (parallel combination of R & C), Th3, and the source. The current in the thyristors is I_{Ti} , the output current is

Iac= I

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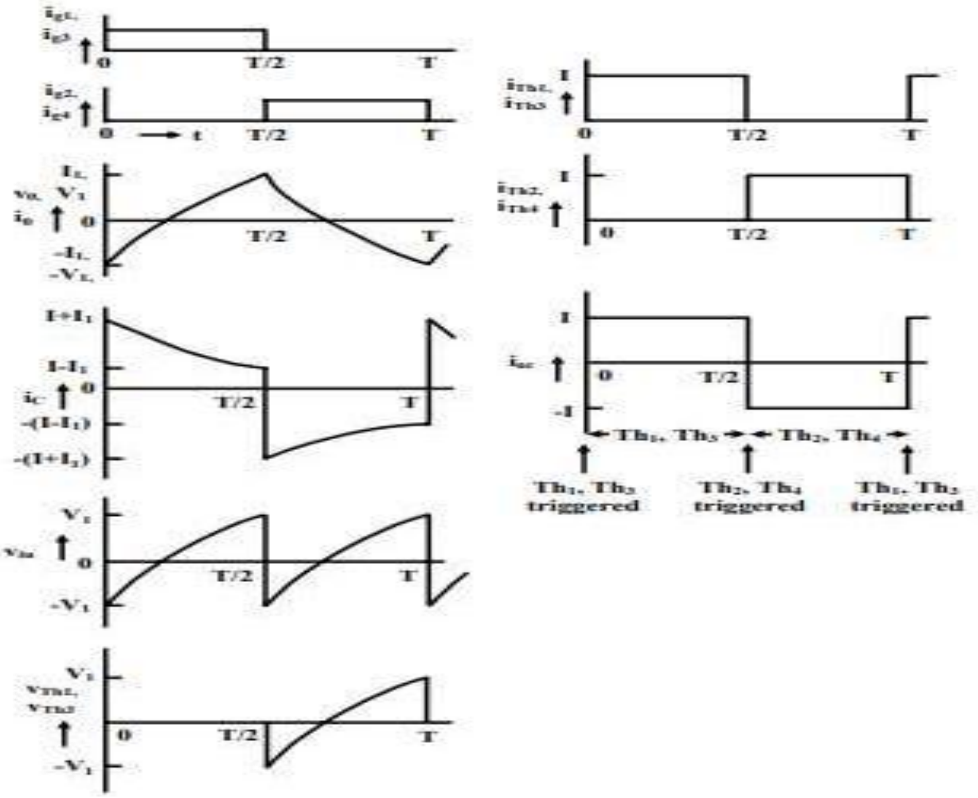

F :

Figure: 5.40 Voltage and current waveforms of load commutated CSI

Introduction to Cyclo converters

The **Cycloconverter** has been traditionally used only in very high power drives, usually above one megawatt, where no other type of drive can be used. Examples are cement tube mill drives above 5 MW, the 13 MW German-Dutch wind tunnel fan drive, reversible rolling mill drives and ship propulsion drives. The reasons for this are that the traditional **Cycloconverter** requires a large number of thyristors, at least 36 and usually more for good motor performance, together with a very complex control circuit, and it has some performance limitations, the worst of which is an output frequency limited to about one third the input frequency .

Figure 3.11 Block diagram of cycloconverters

The **Cycloconverter** has four thyristors divided into a positive and negative bank of two thyristors each. When positive current flows in the load, the output voltage is controlled by phase control of the two positive bank thyristors whilst the negative bank thyristors are kept off and vice versa when negative current flows in the load. An idealized output waveform for a sinusoidal load current and a 45 degrees load phase angle is shown in Figure 3.11. It is important to keep the non conducting thyristor bank off at all times, otherwise the mains could be shorted via the two thyristor banks, resulting in waveform distortion and possible device failure from the shorting current. A major control problem of the **Cycloconverter** is how to swap between banks in the shortest possible time to avoid distortion whilst ensuring the two banks do not conduct at the same time. A common addition to the power circuit that removes the requirement to keep one bank off is to place a centre tapped inductor called a circulating current inductor between the outputs of the two banks. Both banks can now conduct together without shorting the mains. Also, the circulating current in the inductor keeps both banks operating all the time, resulting in improved output waveforms. This technique is not often used, though, because the circulating current inductor tends to be expensive and bulky and the circulating current reduces the power factor on the input

In a **1-φ Cycloconverter,** the output frequency is less than the supply frequency. These converters require natural commutation which is provided by AC supply. During positive half cycle of supply, Thyristors P1 and N2 are forward biased. First triggering pulse is applied to P1 and hence it starts conducting.

As the supply goes negative,P1 gets off and in negative half cycle of supply, P2 and N1 are forward biased. P2 is triggered and hence it conducts. In the next cycle of supply,N2 in positive half cycle andN1 in negative half cycle are triggered. Thus, we can observe that here the output frequency is 1/2 times the supply frequency.

Operation Principles

The following sections will describe the operation principles of the **Cycloconverter** starting from the simplest one, **single-phase to single-phase (1f-1f) Cycloconverter.**

Single-phase to Single-phase (1 Φ-1 Φ) Cycloconverter

To understand the operation principles of **[Cycloconverters,](#)** the single-phase to single-phase **Cycloconverter** (Fig. 3.12) should be studied first. This converter consists of back-to-back connection of two full-wave rectifier circuits. Fig 3.13 shows the operating waveforms for this converter with a resistive load.

Zero Firing angle, i.e. thyristors act like diodes. Note that the firing angles are named as αP for the positive converter and αN for the negative converter. The input voltage, vs is an ac voltage at a frequency, fi as shown in Fig. 3.13. For easy understanding assume that all the thyristors are fired at α=0°

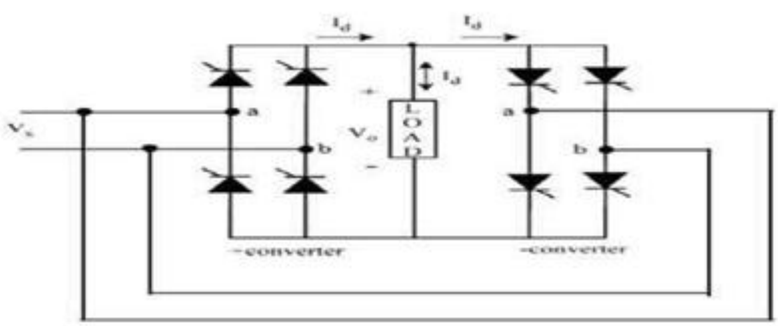

Figure 3.12 circuit diagram of cycloconverter

Consider the operation of the **Cycloconverter** to get one-fourth of the input frequency at the output. For the first two cycles of vs, the positive converter operates supplying current to the load. It rectifies the input voltage; therefore, the load sees 4 positive half cycles as seen in Fig. In the next two cycles, the negative converter operates supplying current to the load in the reverse direction. The current waveforms are not shown in the figures because the resistive load current will have the same waveform as the voltage but only scaled by t he resistance. Note that when one of the converters operates the other one is disabled, so that there is no current circulating between the two rectifiers.

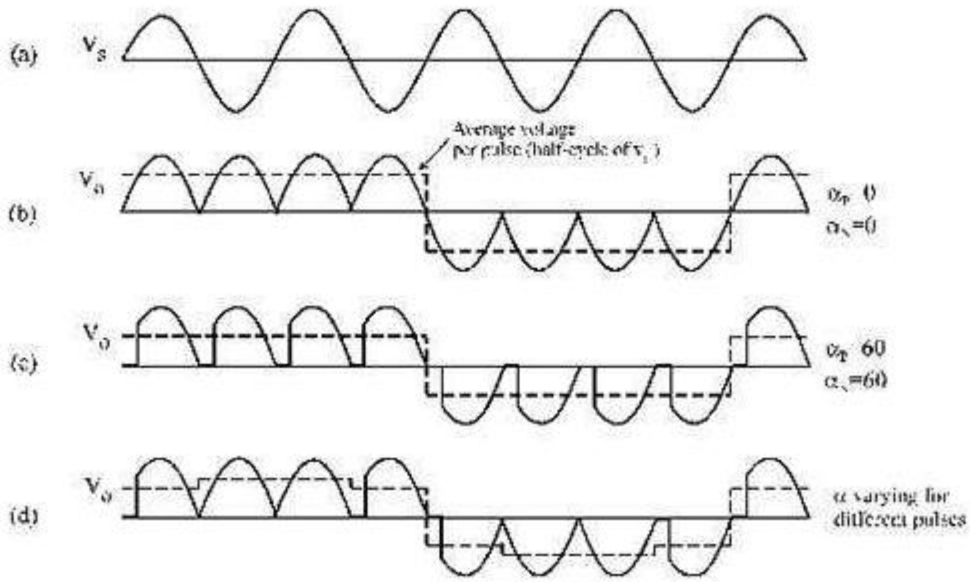

Figure 3.13 Input and output waveforms of cycloconverter

Single phase midpoint Cyclo converters

Basically, these are divided into two main types, and are given below

Step-down cyclo-converter

It acts like a step-down transformer that provides the output frequency less than that of input, fo < fi.

Step-up cyclo-converter

It provides the output frequency more than that of input, fo > fi.

In case of step-down cyclo-converter, the output frequency is limited to a fraction of input frequency, typically it is below 20Hz in case 50Hz supply frequency. In this case, no separate commutation circuits are needed as SCRs are line commutated devices.

But in case of step-up cyclo-converter, forced commutation circuits are needed to turn OFF SCRs at desired frequency. Such circuits are relatively very complex. Therefore, majority of cyclo-converters are of step-down type that lowers the frequency than input frequency.

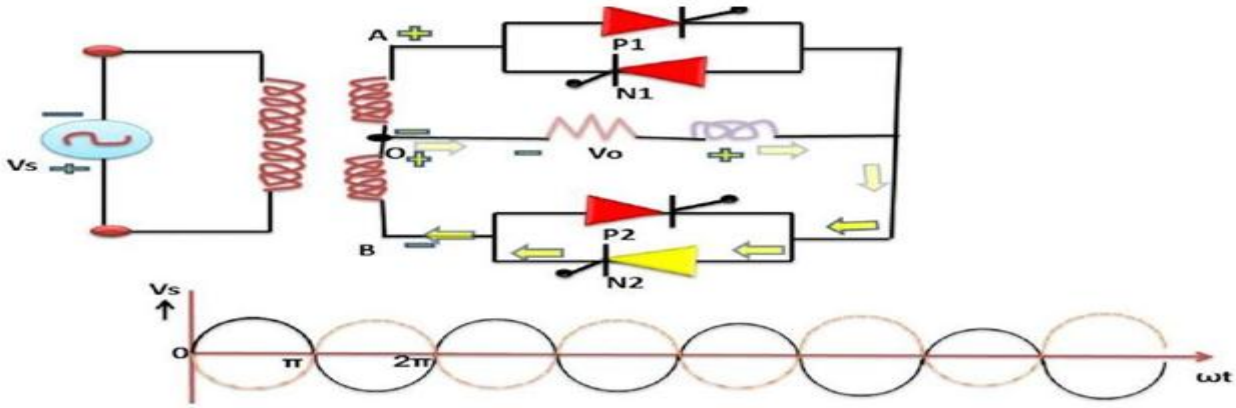

Figure 3.14 circuit diagram of midpoint cycloconverter

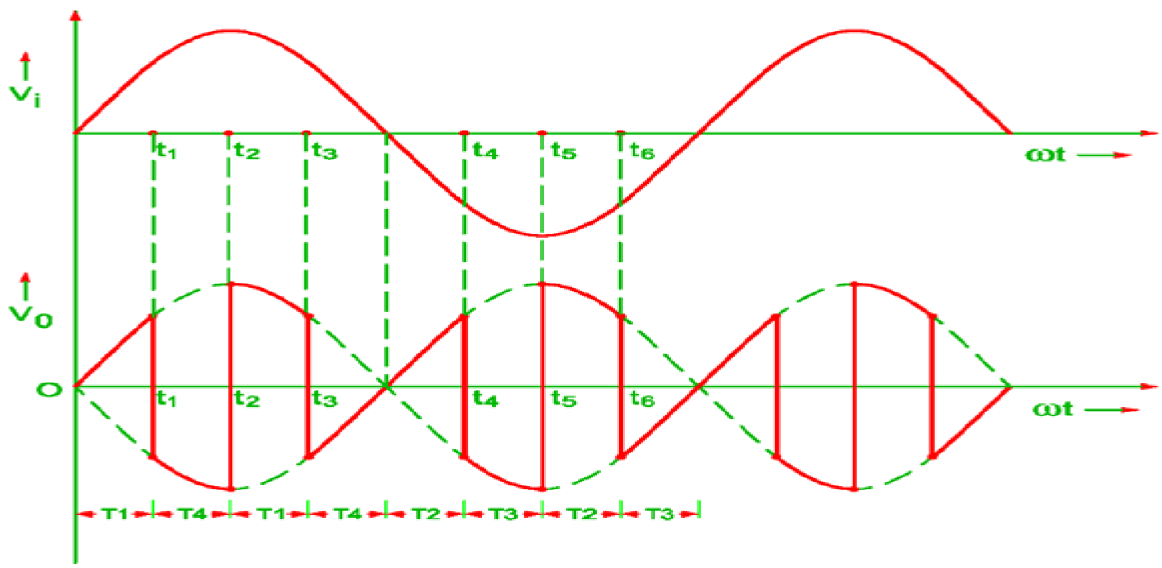

Figure 3.15 Input and output waveforms of midpoint cycloconverter

It consists of single phase transformer with mid tap on the secondary winding and four thyristors. Two of these thyristors P1, P2 are for positive group and the other two N1, N2 are for the negative group. Load is connected between secondary winding midpoint 0 and the load terminal. Positive directions for output voltage and output current are marked in figure 3.14

In figure 3.14 during the positive half cycle of supply voltage terminal a is positive with respect to terminal b. therefore in this positive half cycle, both p1 and N2 are forward biased from wt= 0 to Π. As such SCR P1 is turned on at wt = 0 so that load voltage is positive with terminal A and 0 negative. Now the load voltage is positive. At instant t1 P1 is force commutated and forward biased thyristor N2 is turned on so that load voltage is negative with terminal 0 and A negative. Now the load voltage is negative. Now N2 is force commutated and P1 is turned on the load voltage is positive this is a continuous process and will get step up cyclo converter output

Bridge configuration of single phase Cyclo converter

The equivalent circuit of a cyclo-converter is shown in figure below. Here each two quadrant phase controlled converter is represented by a voltage source of desired frequency and consider that the output power is generated by the alternating current and voltage at desired frequency.

The diodes connected in series with each voltage source represent the unidirectional conduction of each two quadrant converter. If the output voltage ripples of each converter are neglected, then it becomes ideal and represents the desired output voltage.

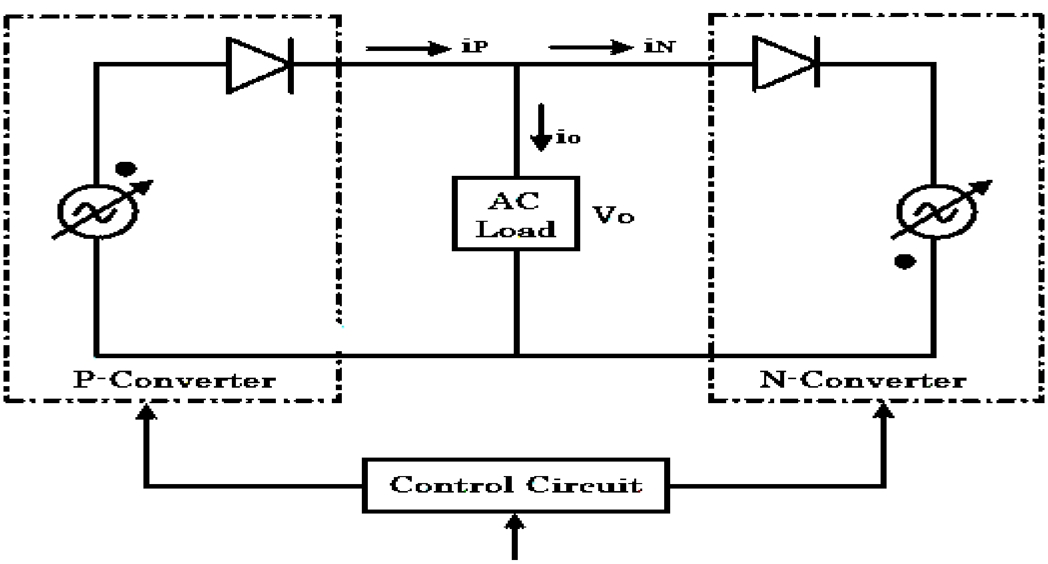

Figure 3.16 Block diagram of bridge type cycloconverter

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If the firing angles of individual converters are modulated continuously, each converter produces same sinusoidal voltages at its output terminals.

So the voltages produced by these two converters have same phase, voltage and frequency. The average power produced by the cyclo-converter can flow either to or from the output terminals as the load current can flow freely to and from the load through the positive and negative converters.

Therefore, it is possible to operate the loads of any phase angle (or power factor), inductive or capacitive through the cyclo-converter circuit.

Due to the unidirectional property of load current for each converter, it is obvious that positive converter carries positive half-cycle of load current with negative converter remaining in idle during this period.

Similarly, negative converter carries negative half cycle of the load current with positive converter remaining in idle during this period, regardless of the phase of current with respect to voltage.

This means that each converter operates both in rectifying and inverting regions during the period of its associated half cycles.

The figure below shows ideal output current and voltage waveforms of a cyclo-converter for lagging and leading power factor loads. The conduction periods of positive and negative converters are also illustrated in the figure.

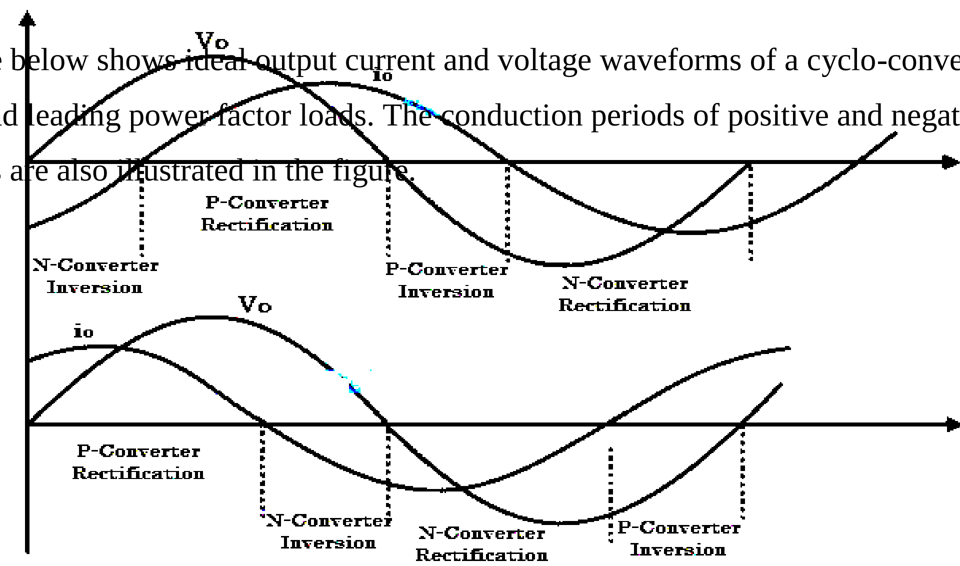


Figure 3.17 cycloconverter waveforms

The positive converter operates whenever the load current is positive with negative converter remaining in idle. In the same manner negative converter operates for negative half cycle of load current.

Both rectification and inversion modes of each converter are shown in figure. This desired output voltage is produced by regulating the firing angle to individual converters.

Single-phase to single-phase cyclo-converters

These are rarely used in practice; however, these are required to understand fundamental principle of cyclo-converters.

It consists of two full-wave, fully controlled bridge thyristors, where each bridge has 4 thyristors, and each bridge is connected in opposite direction (back to back) such that both positive and negative voltages can be obtained as shown in figure below. Both these bridges are excited by single phase, 50 Hz AC supply.

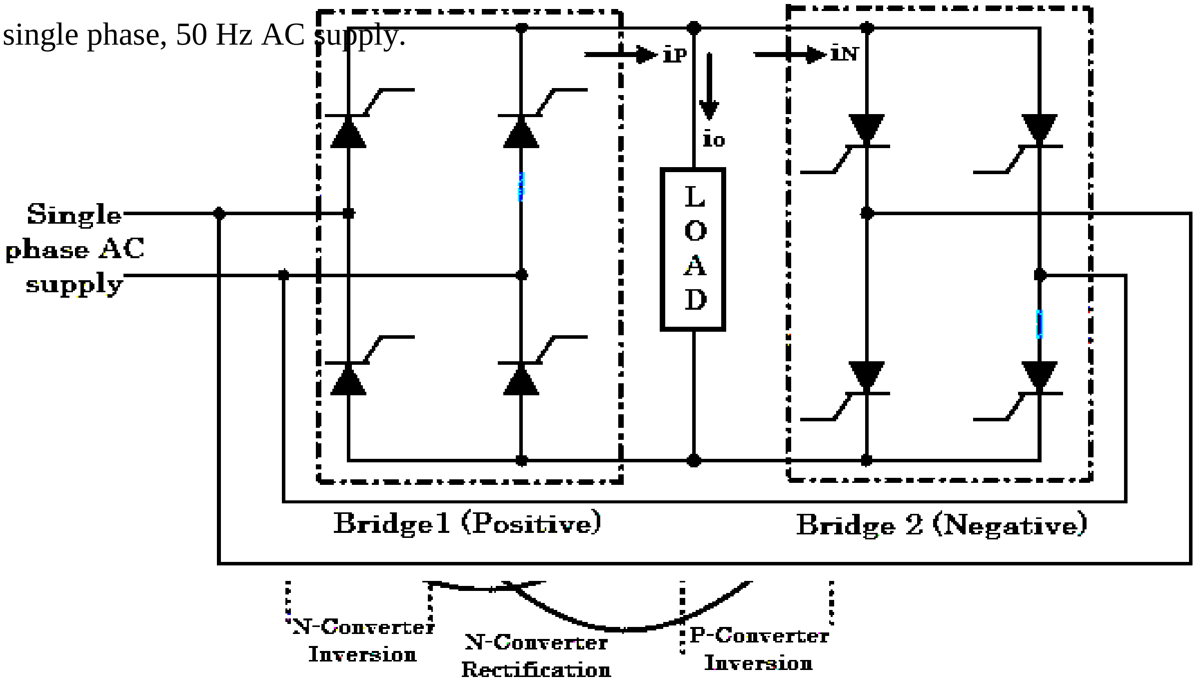


Figure 3.18 Circuit diagram of bridge type cycloconverter

During positive half cycle of the input voltage, positive converter (bridge-1) is turned ON and it supplies the load current. During negative half cycle of the input, negative bridge is turned ON and it supplies load current. Both converters should not conduct together that cause short circuit at the input.

To avoid this, triggering to thyristors of bridge-2 is inhibited during positive half cycle of load current, while triggering is applied to the thyristors of bridge-1 at their gates. During negative half cycle of load current, triggering to positive bridge is inhibited while applying triggering to negative bridge.

By controlling the switching period of thyristors, time periods of both positive and negative half cycles are changed and hence the frequency. This frequency of fundamental output voltage can be easily reduced in steps, i.e., 1/2, 1/3, 1/4 and so on.

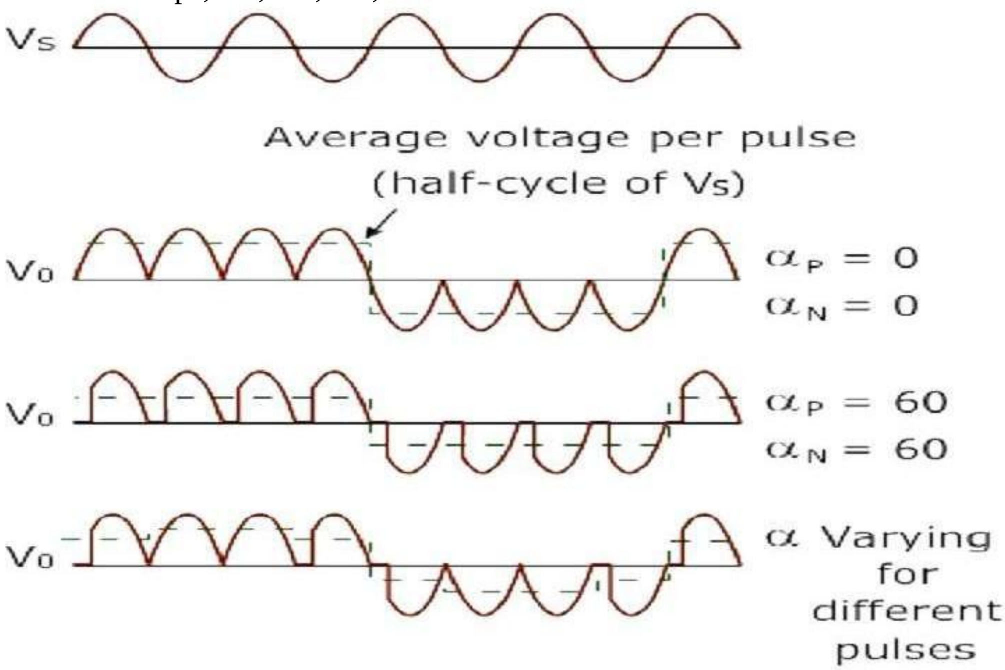


Figure 3.19 Input and output waveforms of bridge type cycloconverter

The above figure shows output waveforms of a cyclo-converter that produces one-fourth of the input frequency. Here, for the first two cycles, the positive converter operates and supplies current to the load.

It rectifies the input voltage and produce unidirectional output voltage as we can observe four positive half cycles in the figure. And during next two cycles, the negative converter operates and supplies load current.

Here current waveforms are not shown because it is a resistive load in where current (with less magnitude) exactly follows the voltage.

Here one converter is disabled if another one operates, so there is no circulating current between two converters. Since the discontinuous mode of control scheme is complicated, most cyclo-converters are operates on circulating current mode where continuous current is allowed to flow between the converters with a reactor.

This circulating current type cyclo-converter can be operated on with both purely resistive (R) and inductive (R-L) loads.

Introduction to dual converters

Dual converter, the name itself says two converters. It is really an electronic converter or circuit which comprises of two converters. One will perform as rectifier and the other will perform as inverter. Therefore, we can say that double processes will occur at a moment. Here, two full converters are arranged in anti-parallel pattern and linked to the same dc load. These converters can provide four quadrant operations. The basic block diagram is shown below

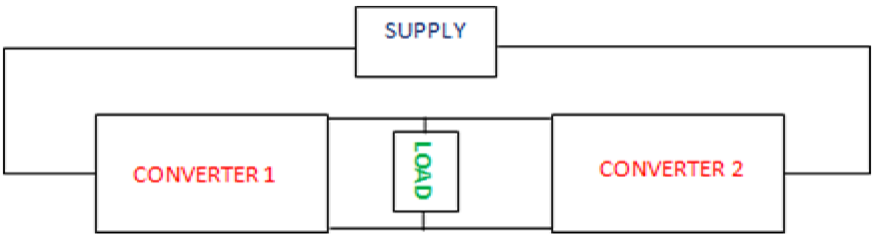


Figure: 2.28 Block diagram of dual converter

Modes of Operation of Dual Converter

There are two functional modes: Non-circulating current mode and circulating mode.

Non Circulating Current Mode

- One converter will perform at a time. So there is no circulating current between the converters.
- During the converter 1 operation, firing angle (α_1) will be $0<\alpha_1< 90^o$; V_{dc} and I_{dc} are positive.
- During the converter 2 operation, firing angle (α_2) will be $0<\alpha_2< 90^o$; V_{dc} and I_{dc} are negative.

Circulating Current Mode

- Two converters will be in the ON condition at the same time. So circulating current is present.
- The firing angles are adjusted such that firing angle of converter 1 (α_1) + firing angle of converter 2 (α_2) = 180^o.
- Converter 1 performs as a controlled rectifier when firing angle be $0<\alpha_1< 90^o$ and Converter 2 performs as an inverter when the firing angle be $90^o<\alpha_2< 180^o$. In this condition, V_{dc} and I_{dc} are positive.
- Converter 1 performs as an inverter when firing angle be $90^o<\alpha_1< 180^o$ and Converter 2 performs as a

controlled rectifier when the firing angle be $0<\alpha_2< 90^o$ In this condition, V_{dc} and I_{dc} are negative.

- The four quadrant operation is shown below

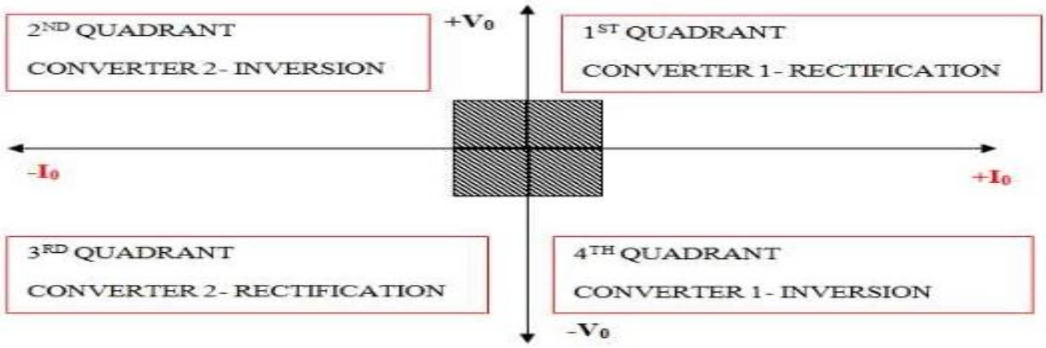


Figure: 2.29 Four quadrant operations of dual converter

Ideal Dual Converter

The term ‘ideal’ refers to the ripple free output voltage. For the purpose of unidirectional flow of DC current, two diodes (D_1 and D_2) are incorporated between the converters. However, the direction of current can be in any way. The average output voltage of the converter 1 is V_{01} and converter 2 is V_{02}. To make the output voltage of the two converters in same polarity and magnitude, the firing angles of the Thyristors have to be controlled.

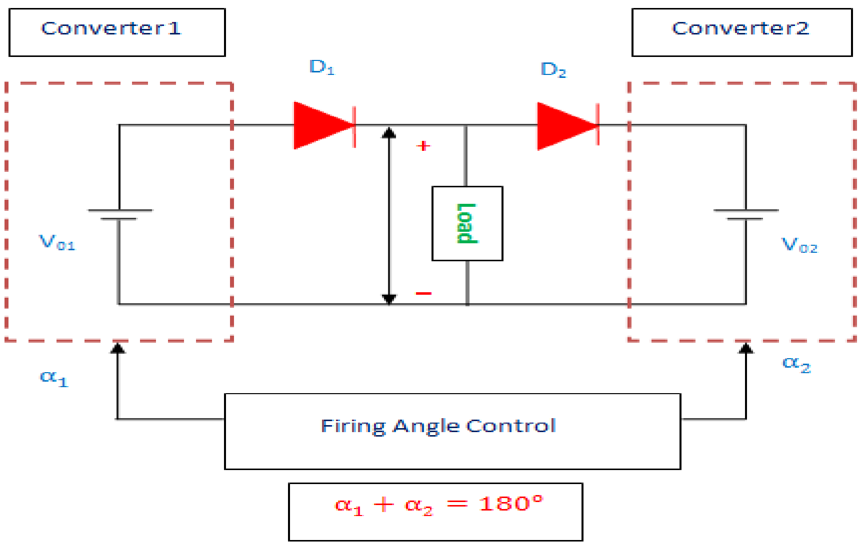


Figure: 2.30 Ideal dual converter

Single Phase Dual Converter

The source of this type of converter will be single-phase supply. Consider, the converter is in non-circulating mode of operation. The input is given to the converter 1 which converts the AC to DC by the method of rectification. It is then given to the load after filtering. Then, this DC is provided to the converter 2 as input. This converter performs as inverter and converts this DC to AC. Thus, we get AC as output. The circuit diagram is shown below.

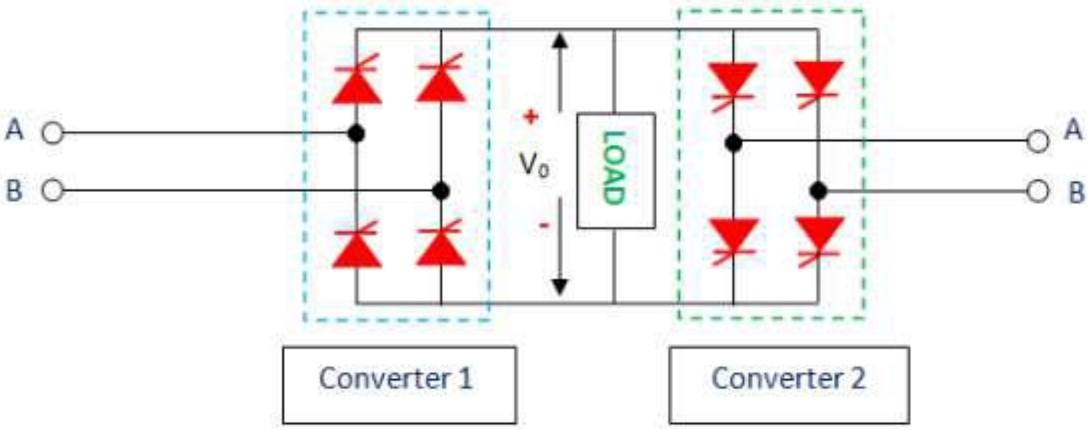


Figure: 2.31 Single phase Dual converter

Average output voltage of Single-phase converter= $\frac{2V_m Cos\ \alpha}{\pi}$

Average output voltage of Three-phase converter = $\frac{3V_{ml} Cos\ \alpha}{\pi}$

For converter 1, the average output voltage, $V_{01} = V_{max} Cos\alpha_1$

For converter 2, the average output voltage, $V_{02} = V_{max} Cos\alpha_2$

$$V_0 = V_{01} = -V_{02}$$
$$V_{max} Cos\alpha_1 = -V_{max} Cos\alpha_2$$
$$Cos\alpha_1 = Cos(180^o - \alpha_2)\ or\ Cos\alpha_2 = Cos(180^o + \alpha_2)$$

Output voltage, $\alpha_1 + \alpha_2 = 180^o\ And\ \alpha_1 - \alpha_2 = 180^o$

The firing angle can never be greater than 180°. So, $\alpha_1 + \alpha_2 = 180^o$

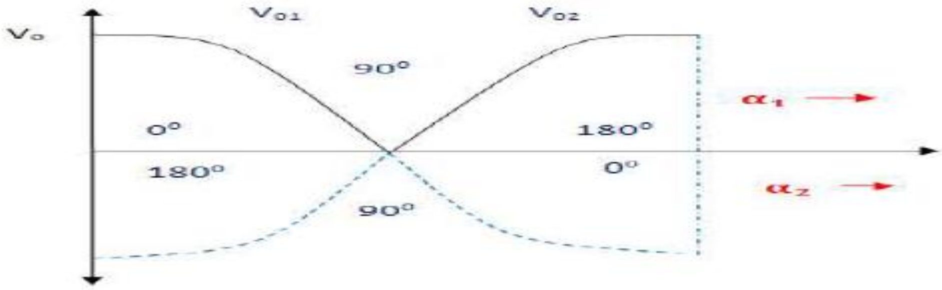


Figure: 2.32 output voltage variation with firing angle

Three Phase Dual Converter

Here, three-phase rectifier and three-phase inverter are used. The processes are similar to single-phase dual converter. The three-phase rectifier will do the conversion of the three-phase AC supply to the DC. This DC is filtered and given to the input of the second converter. It will do the DC to AC conversion and the output that we get is the three-phase AC. Applications where the output is up to 2 megawatts. The circuit is shown below.

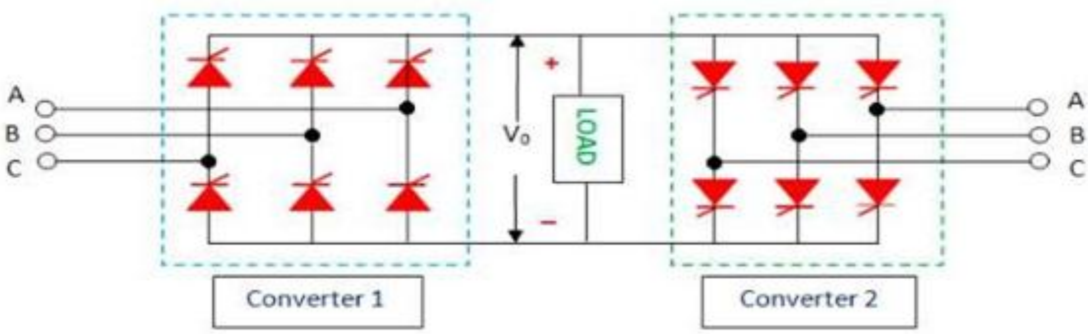


Figure: 2.33 Three phase dual converter

Application of Dual Converter

- Direction and Speed control of DC motors.
- Applicable wherever the reversible DC is required.
- Industrial variable speed DC drives.

INTRODUCTION

The speed of d.c. motor can be controlled very easily by means of regulating its supply voltage by the use of phase controlled rectifiers. This control can be applied to either the field or the armature circuit. The motor response with armature control is faster than that with field control since the time constant of the field is very much larger than that of the armature. Generally, field control is used for speeds above rated value and armature control for speeds below rated value.

Controlled rectifiers either single-phase or multi-phase are widely used in d.c. drive applications where a.c. source is available. The single-phase or multi-phase a.c. is converted to d.c. by a controlled rectifier or converter to give a variable d.c. source, by varying the triggering angle of the thyristor or any other power semiconductor device, that could be supplied to a d.c. motor and thus the speed of the motor can be controlled. Controlled rectifiers used in d.c. drives can be classified as follows:

1- Single-phase controlled rectifiers

(i) Single-phase half-wave converter drives.
(ii) Single-phase full-wave half-controlled converter drives.
(iii) Single-phase full-wave fully-controlled converter drives.
(iv) Single-phase dual converter drives.

2. Multi-phase controlled rectifiers

(i) Three-phase half-wave converter drives.
(ii) Three-phase full-wave fully-controlled converter drives.
(iii) Three-phase full-wave half-controlled converter drives.

PE (iv) Three-phase full-wave dual converter drives.

SINGLE-PHASE CONVERTER DRIVES

These are used for small and medium power motors up to 75kW (100*hp*) ratings. In the following subsections, the various types of these converters will be discussed.

Single-Phase Half-Wave Converter Drives

Fig.13.1 shows a single-phase half-wave converter drive used to control the speed of separately-excited motor. This d.c. drive is very simple, needs only one power switch and one freewheeling diode connected across the motor terminals for the purpose of dissipation of energy stored in the inductance of the motor and to provide an alternative path for the motor current to allow the power switch to commutate easily.

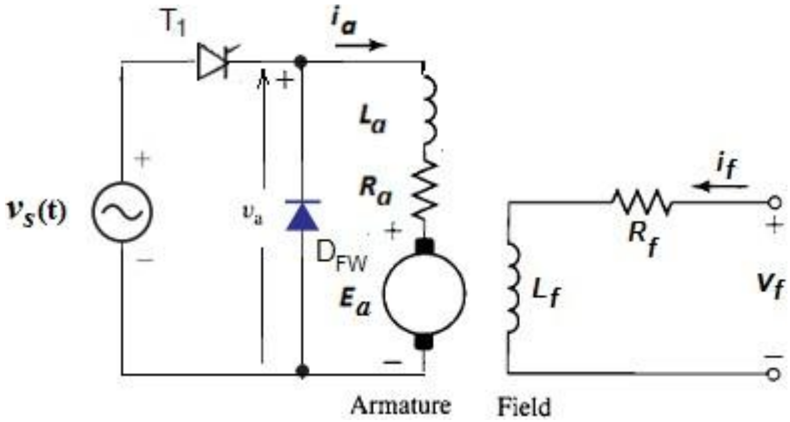


Fig.13.1 Single-phase half-wave converter drive.

Waveforms for steady-state operation of the converter with motor load is shown in Fig.13.2 for the case $\alpha \equiv 80^{\circ}$. It is clear that during the interval $\beta < \omega t < 2\pi$, the armature current is zero, hence the torque developed by the motor is zero, the speed of the motor will be reduced. Since the mech- anical time constant of the motor is larger than its electrical time constant, the inertia of the motor will maintain the speed, but its value will fluctuate resulting in poor motor performance. Therefore, this type of drive is rarely used; it is only used for small d.c. motors below 500 W ratings.

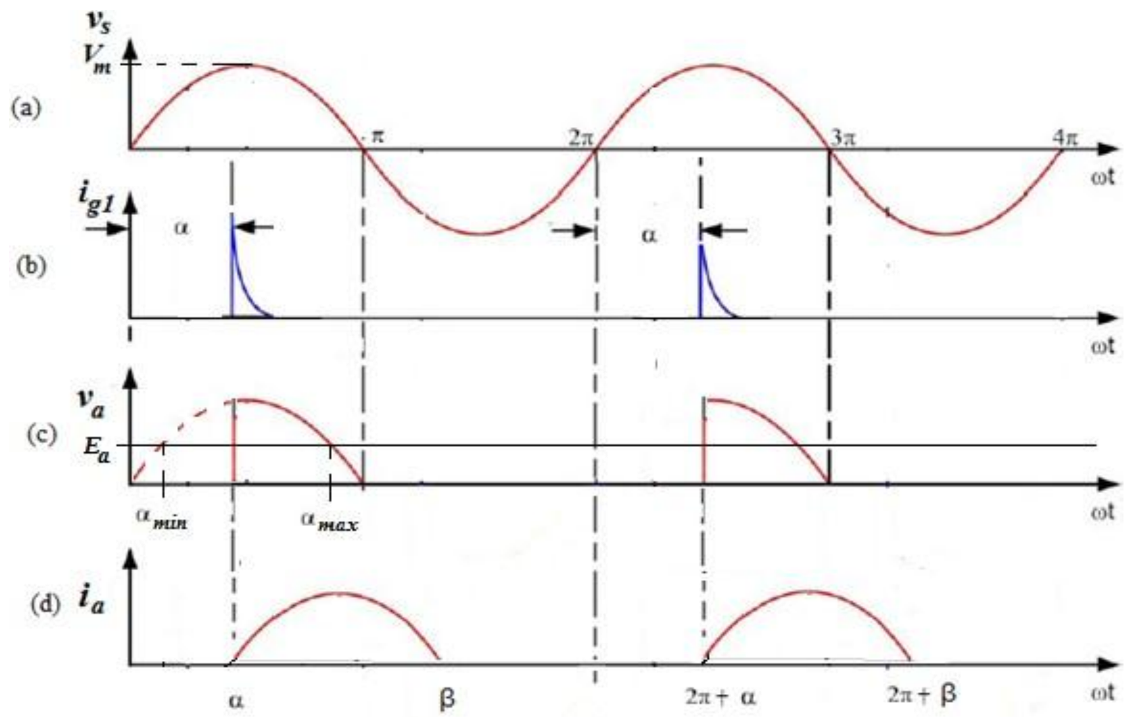


Fig.13.2 Waveforms for steady-state operation of the single-phase half- wave rectifier with motor load.

The average value of the armature voltage can be evaluated as follows:

Assuming the supply voltage $v_s(\omega t) = V_m \sin \omega t$, thus in the positive half-cycle, T_1 will conduct from α to π , where α is the firing angle, and D_{FW} will conduct from π to β, where β is the extinction angle of the current. Hence the average value of the armature current will be,

$$V_{a(av)} = \frac{1}{2\pi}\int_{\alpha}^{\pi} V_s(\omega t)\ d\omega t\ = \frac{1}{2\pi}\int_{\alpha}^{\pi} V_m \sin \omega t\ d\omega t$$

$$= \frac{V_m}{2\pi}\ (1 + \cos \alpha) \qquad (13.1)$$

It is to be noted that the thyristor T_1 is only conducts when supply voltage exceeds back *emf* E_a , therefore, referring to Fig.13.2(c), we define two triggering angles α_{min} and α_{max} as,

α_{min} is the minimum firing angle below which the thyristor cannot be triggered. i.e. when the supply voltage $V_m\ sin\alpha\ > \ E_a$. This angle can be calculated as,

$$V_m\ sin\ \alpha_{min} = E_a \qquad \rightarrow \qquad \alpha_{min} = sin^{-1}\left(\frac{E_a}{V_m}\right) \qquad (13.2)$$

Similarly α_{max} is the maximum firing angle above which the thyristor cannot be triggered. Its value is given by

$$\alpha_{max} = \pi - \alpha_{min} = \ \pi - sin^{-1}\left(\frac{E_a}{V_m}\right) \qquad (13.3)$$

The speed of the motor can be calculated from the general equation of the speed of d.c. motor as,

$$n = \frac{V_a}{K_e \emptyset} - \frac{R_a}{K_T K_e \emptyset^2} T_L$$

$$n = \frac{V_m}{2\pi\, K_e\, \emptyset}(1 + \cos\alpha) - \frac{R_a}{K_T K_e \emptyset^2} T_L \qquad (13.4)$$

or in terms of the angular velocity ω using Eq.(11.25),

$$\omega = \frac{V_{a(av)}}{K\emptyset} - \frac{R_a}{K^2\emptyset^2} T_L$$

Substituting for $V_{a(av)}$ we get,

$$\omega = \frac{V_m}{2\pi K\emptyset}(1 + \cos\alpha) - \frac{R_a}{K^2\emptyset^2} T_L \qquad (13.5)$$

The starting torque can also be calculated from Eq.(13.4) or Eq.(13.5) by setting n or ω equal zero and calculate the torque, (using Eq.(13.5) for example), as

$$0 = \frac{V_m}{2\pi K\emptyset}(1 + \cos\alpha) - \frac{R_a}{K^2\emptyset^2} T_{st}$$

From which,

$$T_{st} = \frac{K\emptyset V_m}{2\pi R_a}(1 + \cos\alpha) \qquad (13.6)$$

And the no load speed is calculated from Eq.(13.5) by setting $T_L = 0$ to give,

$$\omega_o = \frac{V_m}{2\pi K\emptyset}(1 + \cos\alpha) \qquad (13.7)$$

Single-Phase Semiconverter with Separately-Excited d.c. Motor Load

The circuit diagram of a single-phase full-wave half-controlled (semiconverter) drive for controlling a separately-excited d.c. motor is depicted in Fig.13.4. Here a full wave rectifier bridge is supplies the field circuit, and a half-controlled bridge supplies the armature circuit. The vast majority of shunt motors are controlled in this manner.

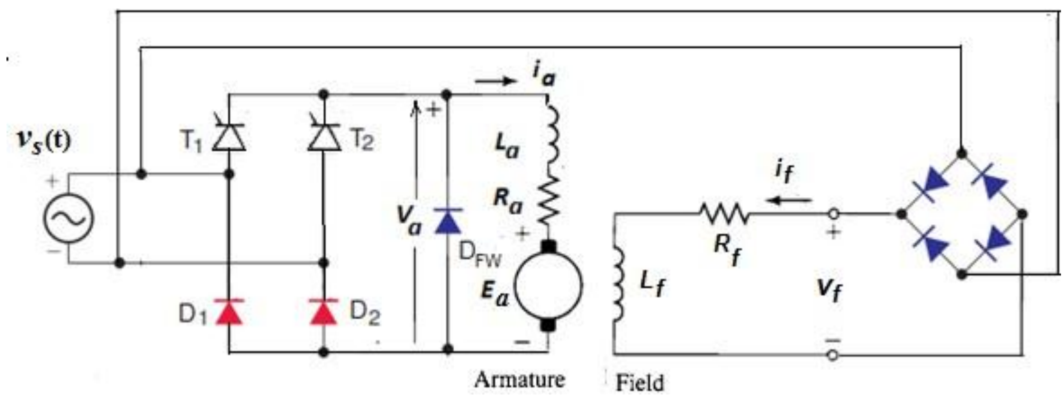


Fig.13.4 Single-phase semiconverter with d.c. motor load.

Assuming the supply voltage , in the positive half- cycle, T_1 and D_2 will conduct from α to (α + δ), where α is the firing angle and δ is the conduction angle. Generally, for medium and large motors the inductance of the armature is small and hence, for the separately-excited motor, the armature current falls to zero at the instant when the back *emf* E_a is equal to the supply voltage . i.e.

The waveforms of the voltage v_a across the armature and the current i_a through the armature are shown in Fig.13.2. It is obvious that the armature current is discontinuous.

(A) Discontinuous armature current operation

The differential equations describing the motor system, during the period the thyristors conduct, are

$$\alpha + \delta = \pi - \sin^{-1}\frac{E_a}{V_m} \qquad (13.8)$$

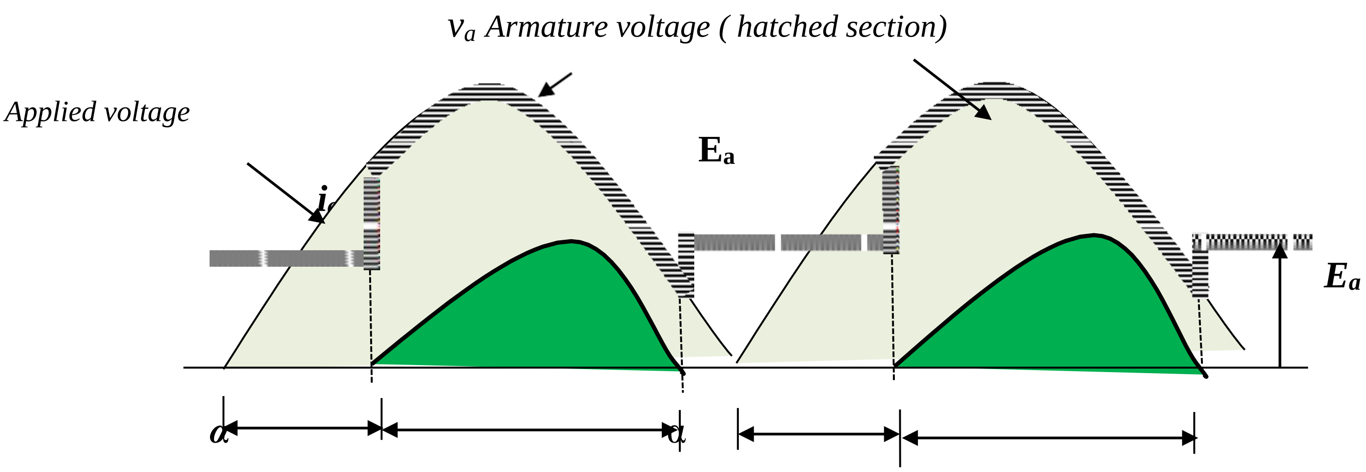


Fig.13.2 Waveforms for single-phase semiconverter operation with discontinuous armature current.

If we assume that the inertia of the rotating system is large then speed fluctuations will be negligible. If each term of is integrated from α to (α + δ) and then divided by π , the instantaneous voltage, current and speed will be converted to their respective average values,

where

$$v_a = i_a R_a + L_a \frac{di_a}{dt} + e_a = i_a R_a + L_a \frac{di_a}{dt} + K\emptyset \omega_m \qquad (13.9)$$

$$T_m = K\emptyset\, I_a = J \frac{d\omega_m}{dt} + B.\omega_m + T_L \qquad (13.10)$$

and the average voltage across is zero.

(B) Analysis with Continuous Armature Current Operation

If the armature inductance is large then conduction will continue, even after the supply voltage has reversed, for which typical waveforms are shown in Fig.13.6. Hence assume continuous current operation, the average value of the armature voltage is:

$$V_{a(av)} = \frac{1}{\pi}\int_{0}^{2\pi} V_a(\omega t)d\omega t = \frac{1}{\pi}\int_{\alpha}^{\pi} V_m \sin(\omega t)\, d\omega t$$

$$V_{a(av)} = \frac{V_m}{\pi}(1 + cos\alpha) \qquad (13.14)$$

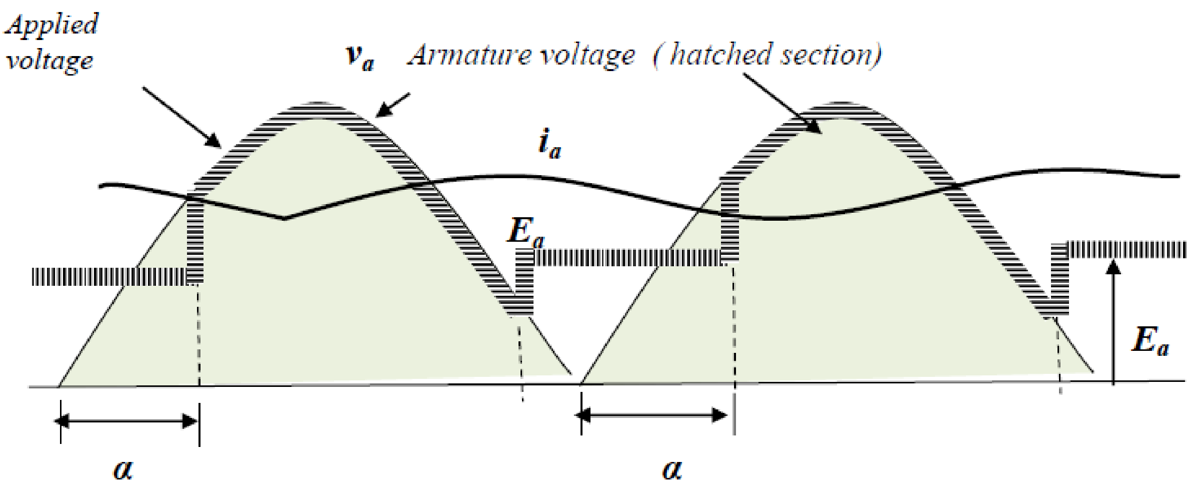


Fig.13.6 Waveforms for single-phase semiconverter operation with continuous armature current.

The average load current is:

$$I_{a(av)} = \frac{V_{a(av)} - E_a}{R_a}$$

$$I_{a(av)} = \frac{V_m}{\pi R_a}(1 + cos\alpha) - \frac{E_a}{R_a} \qquad (13.15)$$

The drive circuit shown in Fig.13.4 can also be used for the variable speed operation of a d.c. series motor. Here the motor field winding is in series with the armature and hence the armature current becomes continuous, using D_1 or D_2 as freewheeling diodes whenever the supply voltage reverses. Two modes of operation are possible. In one mode current flows through T_1 and D_2 (or T_2 and D_1) and the supply voltage appears across the motor. In the second mode T_1 and D_1 (or T_2 and D_2) conduct and the motor voltage is zero. The motor equations are:

$$v_a = V_m \sin(\omega t) \; = i_a R_T + L_T \frac{di_a}{dt} + e_a \quad \text{for } \alpha \le \omega t \le \pi \quad (13.16)$$

and

$$v_a = 0 \; = i_a R_T + L_T \frac{di_a}{dt} + e_a \quad \text{for } \pi \le \omega t \le (\pi + \alpha) \quad (13.17)$$

where R_T and L_T are the total resistance and inductance in the series circuit respectively.

$$T_m = K\emptyset \, I_a = J \frac{d\omega_m}{dt} + B.\omega_m + T_L \qquad (13.18)$$

Integrating as before gives

$$V_{a(av)} = \frac{V_m}{\pi}(1 + cos\alpha) = R_T \, I_{a(av)} + K\emptyset\omega_{m(av)} \qquad (13.19)$$

$$T_{av} = K\emptyset {I_{a(av)}}^2 = B.\omega_{m(av)} + T_L \qquad (13.20)$$

For series motor, it is known that, $\emptyset = k_f \, I_{a(av)}$, hence , $K\emptyset = Kk_f I_{a(av)}$.

Now, let $Kk_f = K_{af}$ $\rightarrow$ new constant (henry), thus the above Eq. (13.19) and Eq. (13.20) can be re-written as,

$$V_{a(av)} = \frac{V_m}{\pi}(1 + cos\alpha) = R_T \, I_{a(av)} + K_{af} I_{a(av)} \, \omega_{m(av)} \qquad (13.21)$$

$$T_{av} = K_{af} {I_{a(av)}}^2 = B.\omega_{m(av)} + T_L \qquad (13.22)$$

<u>Single-Phase Full-Wave Fully-Controlled Rectifier Drives</u>

The circuit diagram of a single-phase full-wave fully-controlled converter drive for controlling a separately-excited d.c. motor is depicted in Fig.13.8.

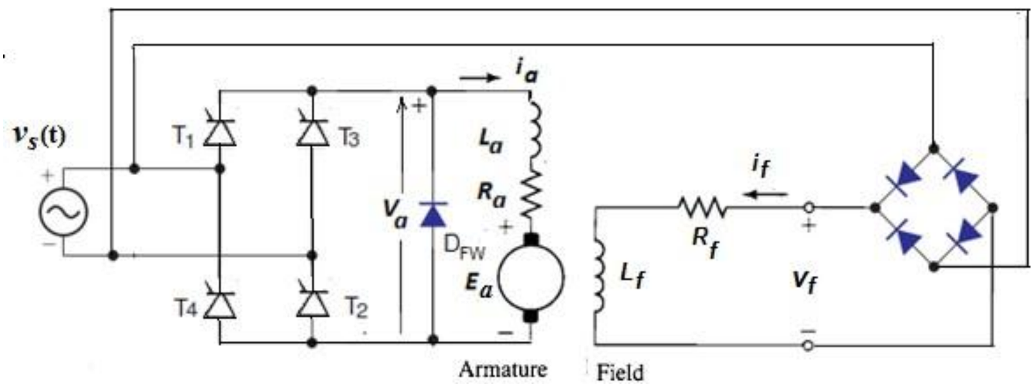


Fig. 13.8 Single-phase full-converter with d.c. motor load.

Here a full-wave rectifier bridge is supplies the field circuit, while the full-converter supplies the armature circuit. The converter has four thyristors that need alternate switching of the pairs of these thyristors T_1 , T_2 or T_3 , T_4 . The converter provides + V_a or – V_a depending on the value of the triggering angle α of the thyristors, thus two quadrant operation is possible. Armature current remains unidirectional due to the converter configuration. The vast majority of shunt motors are also controlled in this manner.

(C) Single-phase full-converter operation with continuous motor current

The waveforms of voltage v_a across the armature and the current i_a through the armature are shown in Fig. 13.9 for continuous current mode of operation.

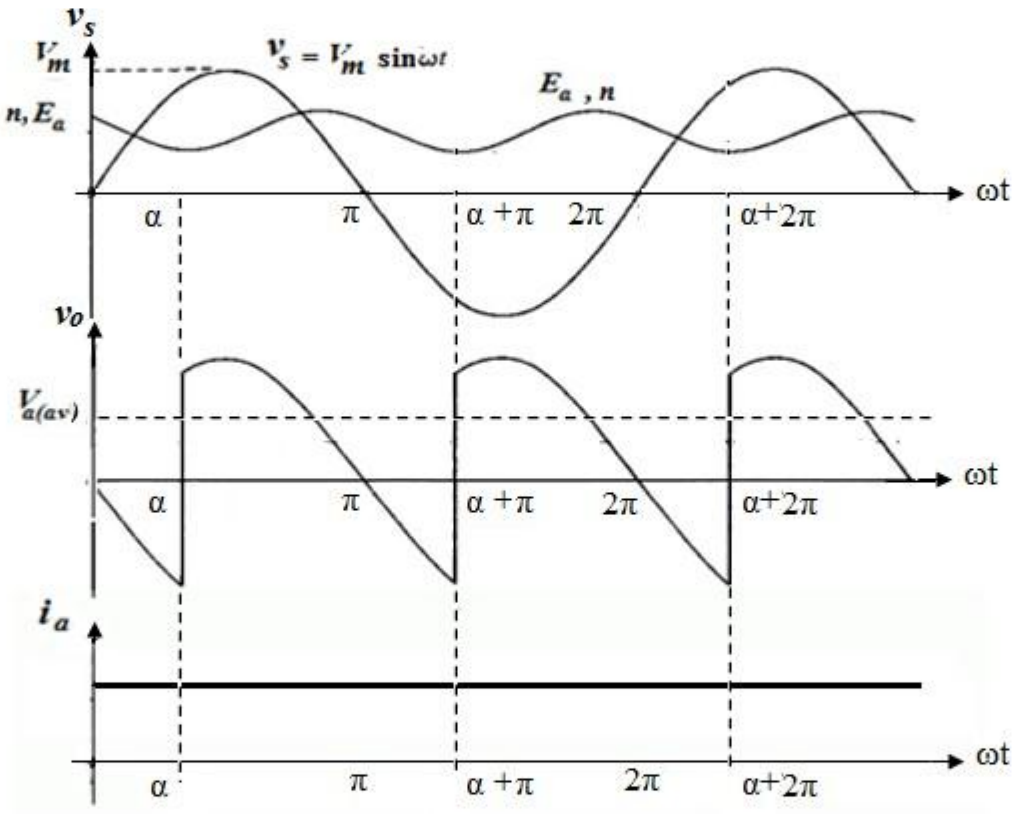


Fig.13.9 Waveforms of the armature voltage and the current for contin- uous current operating mode.

In any case it is obvious that the thyristor only conducts when supply voltage exceeds the back *emf* , i.e. .

When thyristors T_1 and T_2 triggered at $\omega t = \alpha$, T_3 and T_4 must be turned off. When thyristors T_3 and T_4 are triggered at $\omega t = \alpha + \pi$ negative voltage is applied across T_1 and T_2 causes them to commutate naturally. The average value of the armature voltage, as can be deduced from Fig.13.9 is

$$V_{a(av)} = \frac{1}{2\pi}\int_{0}^{2\pi} v_a(\omega t)d\omega t = \frac{2}{2\pi}\int_{\alpha}^{\pi+\alpha} V_m \sin(\omega t)\, d\omega t$$

$$V_{a(av)} = \frac{2V_m}{\pi}\cos\alpha \qquad (13.23)$$

The average current

$$I_{a(va)} = \frac{V_{a(av)} - E_a}{R_a}$$

$$= \frac{2V_m}{\pi R_a}(cos\alpha) - \frac{E_a}{R_a} \qquad (13.24)$$

(D) Power and power factor

The power taken by the motor can be calculated as

$$P_{in} = P_a = I_a^2 R_a + E_a I_a \qquad (13.25)$$

$E_a I_a$ represents the output power plus the motor friction and windage losses. If the mechanical losses in the motor and electrical losses in the rectifier switches are neglected, the output power and the operating efficiency are

$$P_{out} = E_a I_a = T\omega_m \qquad (13.26)$$

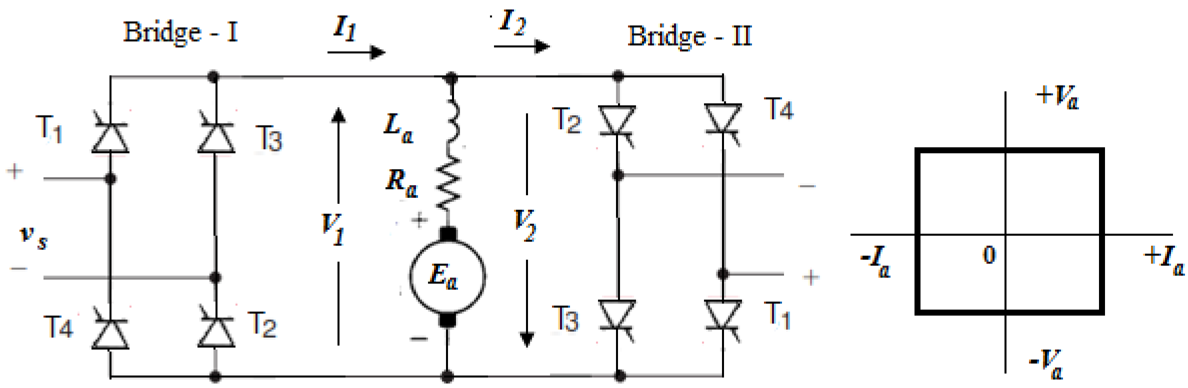
$$\eta = \frac{P_{out}}{P_{in}} = \frac{E_a I_a}{P_a} \qquad (13.27)$$

Since the input current has *rms* value equal to that of the motor current, thus the operating power factor is

$$\text{Power factor} = \frac{P_a}{\frac{E_m}{\sqrt{2}} . I_L} \qquad (13.28)$$

Single-Phase Dual Converter Drives

In some industrial applications, d.c. motor may require to be operated in four quadrants without a switching changeover. In this case, duplication of power electronics converters is used. Fig.13.11 shows a simple dual converter drive circuit diagram which consists of two single-phase full bridge converters connected in inverse-parallel supplying a d.c motor. One bridge for one direction of motor current and the other bridge for the opposite direction of current. The controls are interlock to prevent their simultaneous operation to avoid short circuits on one another. Bridge-I provides operation in the first and fourth quadrants while bridge-II provides operation in second and third quadrants. Therefore, the dual converter is a four quadrant drive which allows four quadrant of machine operation without a switching changeover.



(a) (b)

Fig.13.11 Dual converter drive: (a) Circuit diagram, and (b) Quadrants of operation.

To illustrate how a speed reversal takes place, bridge-I has its firing signals removed; i_1 falls to zero and after few milliseconds delay, bridge- II is fired. This drive is employed for motors of rating up to 15 kW. On the circuit of Fig.13.13, positive voltages are shown by the arrowheads, though in the equations, these voltages may have negative values. These equations are:

Bridge – I operating:

$$V_1 = V_{a(av)\,1} = \frac{2V_m}{\pi}\cos\alpha_1 = V_{do}\ \cos\alpha_1 = E_a + I_1\ R_a \qquad (13.29)$$

Bridge – II operating:

$$V_2 = V_{a(av)\,2} = -\left(\frac{2V_m}{\pi}\cos\alpha_2\right) = V_{do}\ \cos\alpha_2 = E_a - I_1\ R_a \qquad (13.30)$$

PE where

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

Which is the output voltage of the converter when $\alpha = 0^\circ$.

Equations (13.29) and (13. 30) are shown as straight lines on Fig.13.12, the intersection of the machine and bridge characteristics giving the operating points.

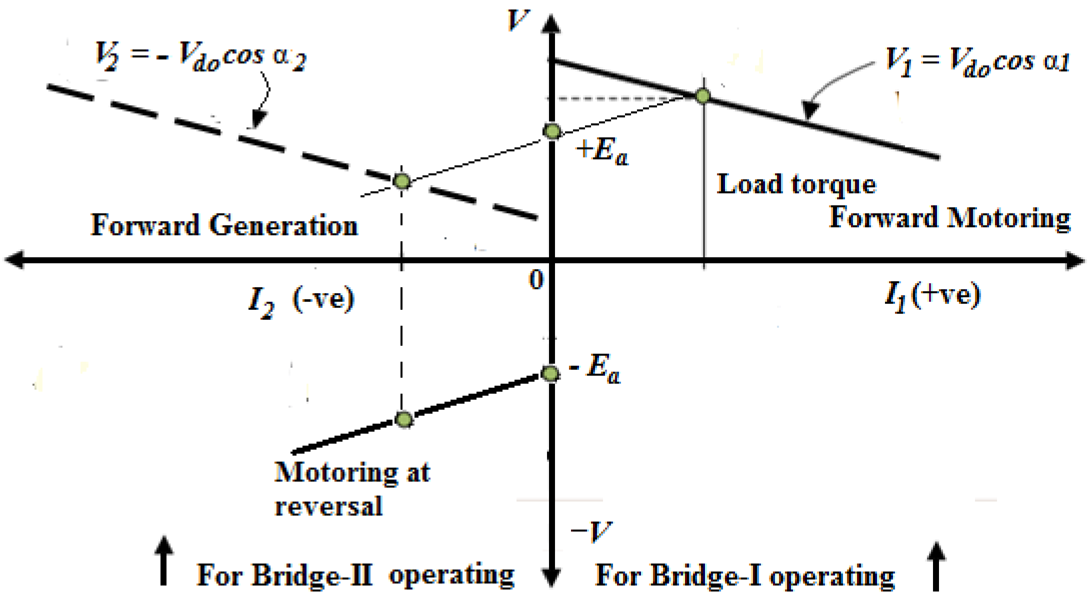


Fig.13.12 Dual converter.

THREE-PHASE DC DRIVES

Three-phase converters are commonly used in adjustable speed drives from about 15 kW up to several thousand kilowatts ratings. The output voltage of a three-phase converter has less ripple contents than the single- phase converter, and therefore, the armature current will be smoother and mostly continuous.

The theory and operation of all types of the three-phase converters was fully discussed in chapter Three-Part I, for the case of passive impedance load. However, three-phase converters could be half-wave, full-wave fully-controlled, full-wave half-controlled (semiconverter) and dual converter when reversible armature current is needed. The three-phase half-wave circuit is only of theoretical importance and is generally not used in industrial applications because of the d.c. components inherent in its line currents. For medium size motors, in the range 15 – 120 kW, either the full-converters or semi converter are used.

Three-Phase Half-Wave (or p = *3*) Converter

In the three-phase half-wave converter, the motor load is connected between the converter positive terminal (cathodes of all thyristors) and the supply neutral as shown in Fig.13.13 .The firing angle α is also defined to be zero from the zero crossings of the input voltages. This converter is used for motor ratings from 10 to 50 hp and it is rarely used in practice because of the d.c. component in the line current. Waveforms of the armature (load) voltage and current are shown in Fig.13.14.

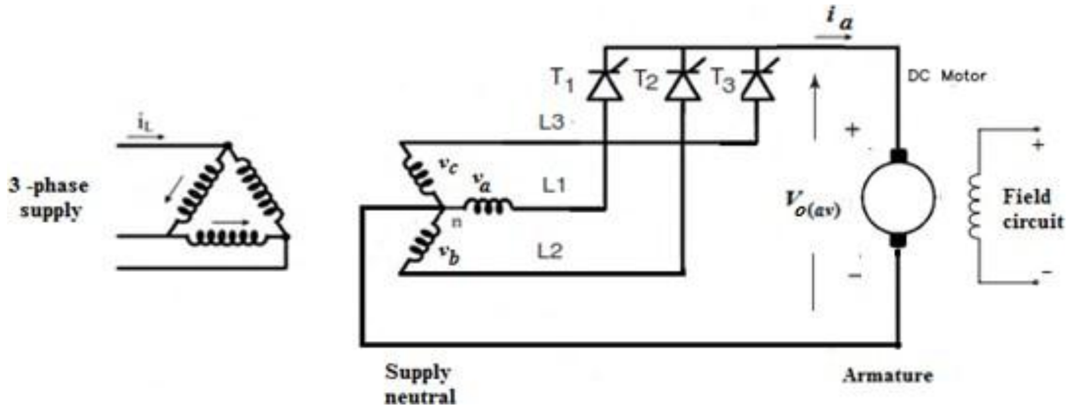


Fig.13.13 Three-phase, half-wave controlled converter with motor load.

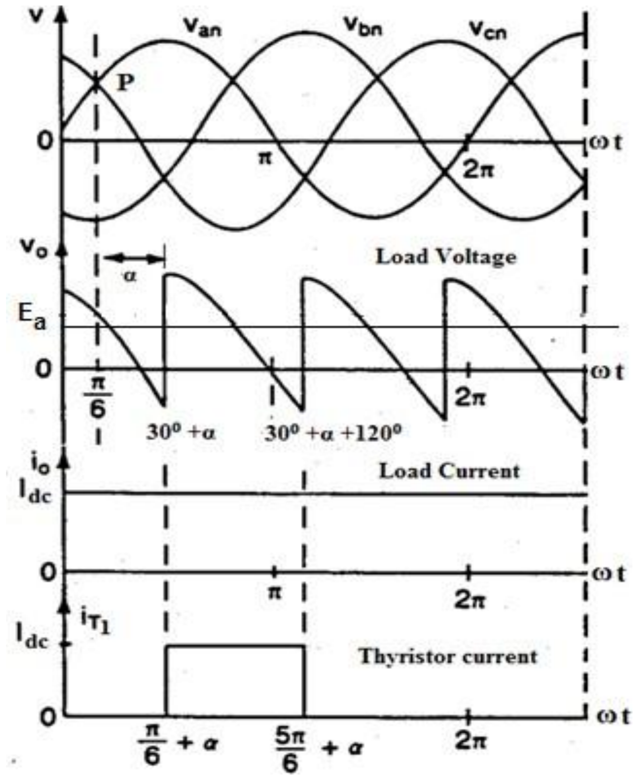


Fig.13.14.Wave forms for the three-phase half-wave converter d.c. drive.

To find the average value of the output d.c. voltage of the converter, let the transformer secondary phase to neutral voltages be,

$$v_{an} = V_m\, sin\omega t$$
$$v_{ab} = V_m \sin(\omega t - 120°)$$
$$v_{ab} = V_m \sin(\omega t - 240°)$$

Assuming continuous current conduction, the average output voltage is,

$$V_{o(av)} = \frac{1}{\frac{2\pi}{3}} \int_{30°+\alpha}^{30°+\alpha+120°} V_m\ sin\omega t\ d\omega t = \frac{3V_m}{2\pi} [-cos\omega t] \begin{matrix} 150° + \alpha \\ 30° + \alpha \end{matrix}$$

$$V_{o(av)} = \frac{3V_m}{2\pi} [-(\cos(150° + \alpha) - \cos(30° + \alpha))]$$

$$V_{o(av)} = \frac{3\sqrt{3}V_m}{2\pi} cos\alpha \qquad (13.31)$$

where V_m is the peak of the supply line-neutral voltage.

A firing angle of zero degree produces the maximum output d.c. voltage for all ac-to-dc converter circuits. For continuous current conduction, each thyristor carries current for 120°, followed by 240° of non-conduction. The firing angle α can be varied in the range of ±180°. For α > 90°, the output d.c. voltage becomes negative, whilst the motor current is positive and continuous. This implies operation of the converter in the fourth quadrant of the *V-I* plane as shown in Fig.13.15 where the converter operates in the inversion mode. In this mode of operation the motor supplies power to the a.c. source through the converter steadily. This mode of operation is called regenerative conversion. For example, an overhauling motor can supply its energy to the a.c. mains in this way. However, in the case of the overhauling motor, controlled braking is thus possible.

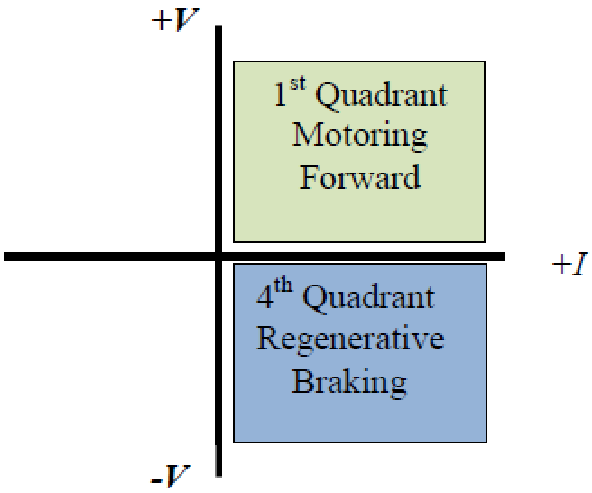


Fig.13.15 Two-quadrant operation of three-phase half-wave converter drive.

Three-Phase Semiconverter Drive

The three-phase semiconverter is a one-quadrant drive. Its circuit includes a freewheeling diode D_{FW} to maintaining continuous load current. It uses three thyristors and three diodes; hence a cost advantage is obtained compared with the full-converter. The circuit diagram for a separately-excited d.c. motor supplied from a three-phase a.c. supply through a three-phase semiconverter is shown in Fig.13.16. This converter

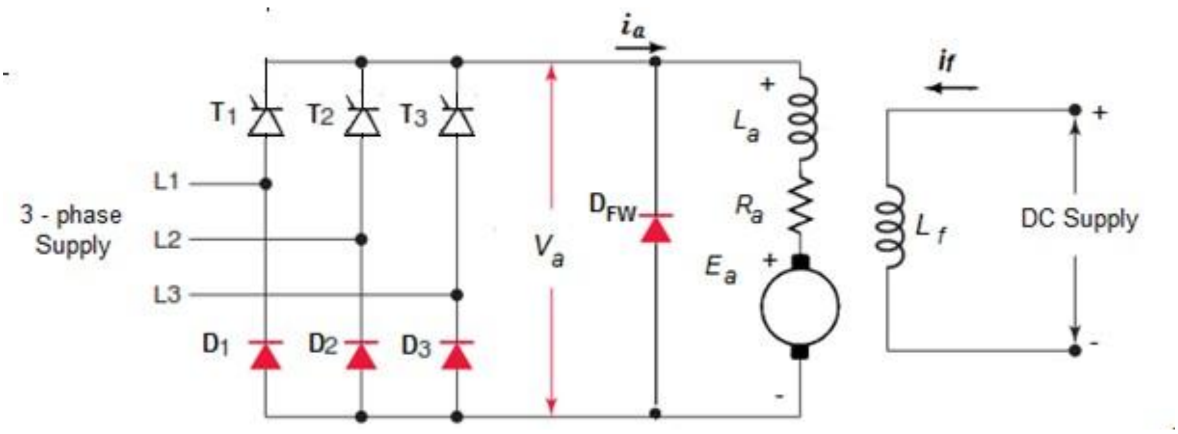


Fig.13.16 Three-phase semiconverter drive.

is used for motor ratings from 15 to 150 hp. The field converter may be single-phase or three-phase semiconverter with firing angle of α_f .

Assuming continuous current operation, the average value of the armature voltage at the motor terminals is a contribution from the upper half-bridge plus a contribution from the uncontrolled lower bridge. Hence for all firing angles we can write:

$$V_{o(av)} = \frac{3\sqrt{3}}{2\pi} V_m\, cos\alpha + \frac{3\sqrt{3}}{2\pi} V_m$$

$$= \frac{3\sqrt{3}}{2\pi} V_m(1 + \cos\alpha) \qquad (13.32)$$

The average armature current is:

$$I_{a(va)} = \frac{V_{a(av)} - E_a}{R_a} \qquad (13.33)$$

$$= \frac{3\sqrt{3}V_m}{2\pi R_a}(1 + cos\alpha) - \frac{E_a}{R_a} \qquad (13.34)$$

For discontinuous current operation, the above equations are not valid.

Features:

- Since only three thyristors are used, the circuit is not expensive and a simple control circuitry is required.
- Dynamic braking can be performed by switching armature connection to an external resistance.
- Operation is in the first quadrant only (Fig.13.17). However, bi-directional rotation can be obtained by reversing field current or armature terminals when the motor has been stopped.

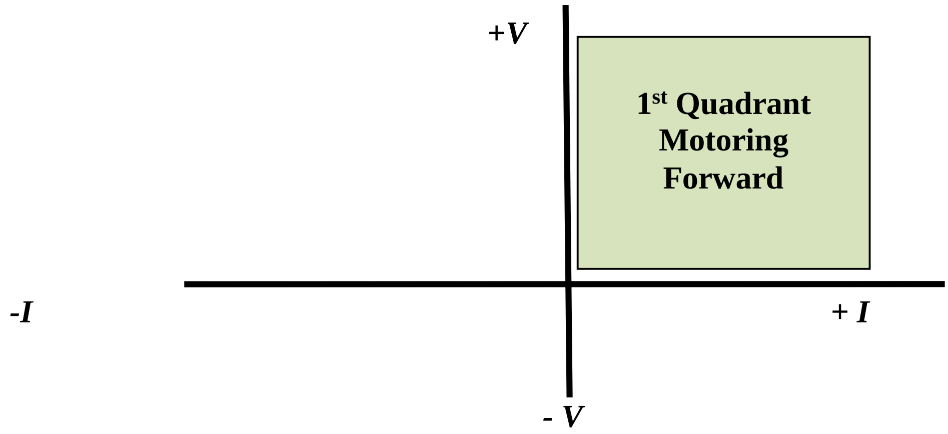


Fig.13.17 One-quadrant operation of the three-phase semiconverter drive.

Three-Phase Full-Converter Drive

The circuit diagram for a separately-excited d.c. motor supplied from a three-phase a.c. supply through a three-phase full-converter is shown in Fig.13.18.

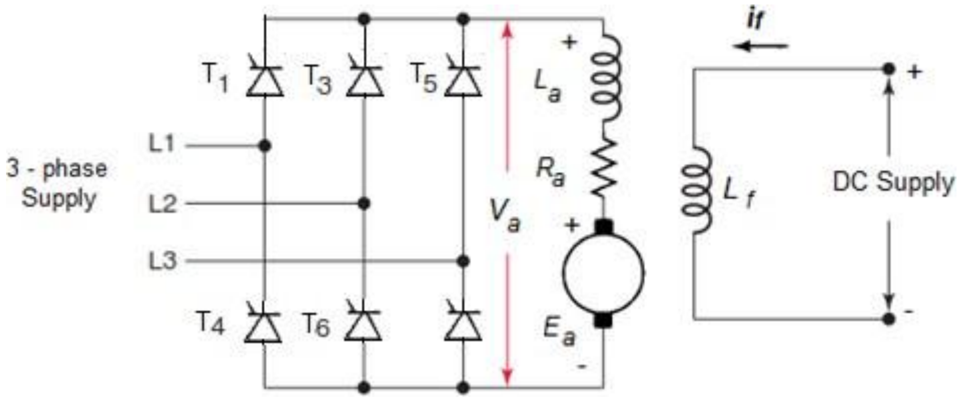


Fig.13.18 Three-phase full-converter with a separately-excited d.c. motor load.

If the motor armature inductance is large and the firing angle is small then the armature current is likely to be continuous. However, with small armature inductance and large firing angles the armature current may become discontinuous particularly when the back *emf* is relatively high.

- For continuous current operation, the armature voltage has an average value

$$V_{a(av)} = \frac{3\sqrt{3}}{\pi} V_m \cos\alpha \qquad (13.35)$$

Thus, the average armature current is:

$$I_{a(va)} = \frac{V_{a(av)} - E_a}{R_a} \qquad (13.36)$$

$$= \frac{3\sqrt{3}V_m}{\pi R_a}\cos\alpha - \frac{E_a}{R_a} \qquad (13.37)$$

- For discontinuous current operation of the full-converter, the waveforms of the voltage and currents are shown in Fig.13.19.

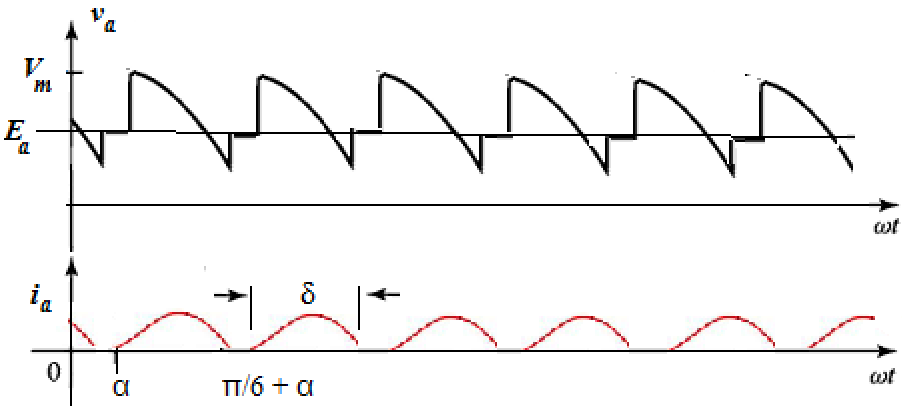


Fig.13.19 Discontinuous current operation waveforms of the full-converter.

The differential equations describing the motor system, during the period the thyristors conduct, are

$$v_a = i_a R_a + L_a \frac{di_a}{dt} + e_a \qquad (13.38)$$

$$\sqrt{3}V_m \sin(\omega t + 30°) = i_a R_a + L_a \frac{di_a}{dt} + K\emptyset\omega_m$$

$$T_m = K\emptyset\, I_a = J\,\frac{d\omega_m}{dt} + B.\omega_m + T_L \qquad (13.39)$$

If it is assumed that the inertia of the rotating system is large then speed fluctuations will be negligible. If each term of is integrated from α to (α + π/6) and then divided by π/3, the instantaneous voltage, current and speed will be converted to their respective average values,

$$V_{a(av)} = \frac{3}{\pi}\int_{\alpha}^{\alpha+\pi/6} \sqrt{3}V_m\,(\sin\omega t + 30°)\ d\omega t$$

PF

$$= \frac{1}{\pi}\int_{\alpha}^{\alpha+\pi/6} L_a \frac{di_a}{dt}\, d\omega t + \frac{1}{\pi}\int_{\alpha}^{\alpha+\pi/6} R_a\, i_a\ d\omega t + \frac{1}{\pi}\int_{\alpha}^{\alpha+\pi/6} K\emptyset\, \omega_m\, d\omega t$$

Thus

$$V_{a(av)} = I_{a(av)} R_a + K\emptyset\, \omega_{m(av)} \qquad (13.40)$$

where

$$V_{a(av)} = \frac{3\sqrt{3}V_m}{\pi}[\cos(\alpha + 30°) - \cos(\alpha + 90°)] \qquad (13.41)$$

and the average voltage across L_a is zero.

Similarly,

$$T_{m(av)} = K\emptyset\, I_{a(av)} = B \cdot \omega_{m(av)} + T_L \qquad (13.42)$$

Three-Phase Dual Converter Drive

Four-quadrant operation of a medium and large size d.c. motor drive (200-2000 hp) can be obtained by the three-phase dual converter shown in Fig.13.20. The average motor voltage is required to be equal for both converters, which required that the firing angles of the two sets of the thyristors should sum to 180˚.

The armature voltage supplied by converter-1 (for continuous current operation) is

Bridge – I operating:

$$V_1 = V_{a(av)1} = \frac{3\sqrt{3}V_m}{\pi}\cos\alpha_1$$
$$= V_{do}\cos\alpha_1 = E_a + I_1 R_a \qquad (13.43)$$

Bridge – II operating:

$$V_2 = V_{a(av)2} = -\left(\frac{3\sqrt{3}V_m}{\pi}\cos\alpha_2\right)$$
$$= V_{do}\cos\alpha_2 = E_a - I_1 R_a \qquad (13.44)$$

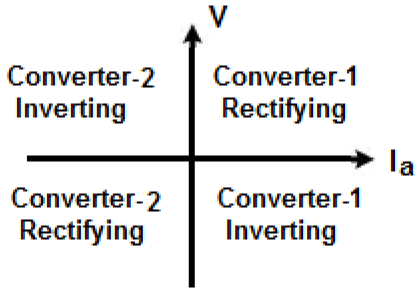


Fig.13.20 Four-quadrant three-phase d.c. drive.

where $$V_{do} = \frac{3\sqrt{3}V_m}{\pi}$$

and $$\alpha_2 = \pi - \alpha_1 \; .$$

Two modes of operation can be achieved with this circuit: Two modes of operation can be achieved with this circuit:

(a) Circulating current operating mode:

Here, instantaneous values of circulating current are limited by use of reactors and mean level is controlled by current loop. Circulating current may be constant giving linear characteristic or it may be reduced to zero giving higher gain portion of overall characteristic.

Advantage: Continuous bridge current maintain armature current at all times, no discontinuity occurs.

Disadvantage: Presence of circulating current reduces efficiency.

(b) Circulating current-free operation mode:

In this mode only one converter operates at a time. Logic used to prevent the two bridges being turn on at the same time. Reactors or inductors used to maintain continuous current down to acceptable low levels. Discontinuity occurs at zero and also a time delay (*ms*) introduced at the zero current level.

Advantage: Higher efficiency than circulating current schemes, hence used more widely.

Disadvantage: Dead time, discontinuity in zero current regions.

Control of AC drives

This chapter is to describe the control of induction motor. There are a number of control methods. We will concentrate the variable voltage control, constant V/F control, and vector control. There are too many references in these areas and we have listed a few of them at the last section.

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1. Equivalent circuit

The induction motor is a complex electromagnetic structure. At one level it consists of three stator phase windings and three effective rotor windings. Six voltage equations can therefore be written down, with self and mutual inductances between all windings. The mutual inductances between stator and rotor windings vary with position. By inserting three phase symmetry, and sinusoidal space and time quantities into these equations, the six equations can be reduced a single phase equivalent circuit as shown in Fig 1. Most calculations relating to induction motor performance can be undertaken using this equivalent circuit.

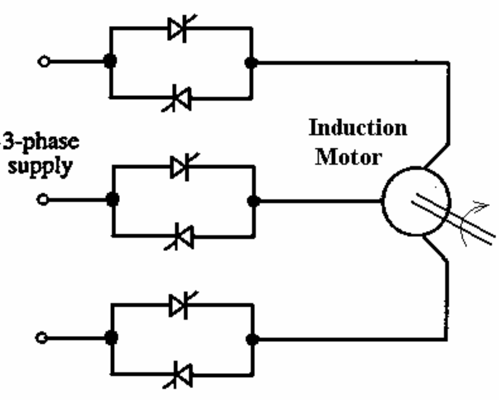

Fig 2.Induction motor with a voltage controller

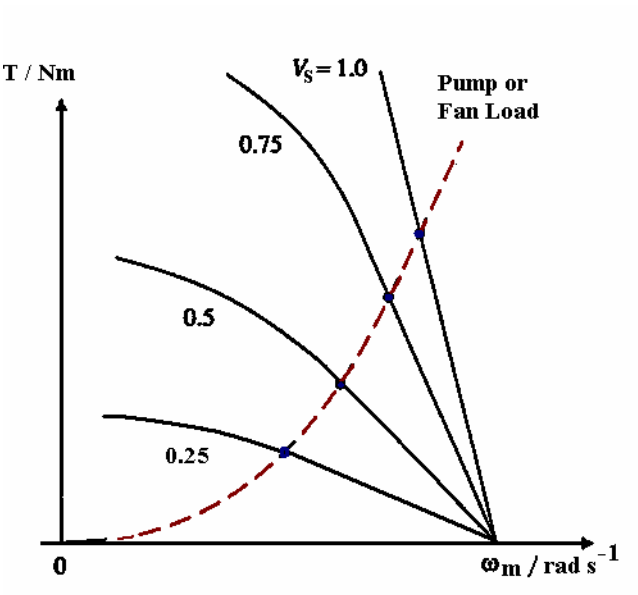

Fig 3. Control of fan speed by stator voltage

A typical fan load speed-torque relation is:

$$T = k\omega_r^2$$

Fig 3 shows a torque-speed characteristic of the variable speed controlled drives, The fan load is also shown. The intersections of this curve with the motor curve gives possible values of steady state speed. The rotor current and rotor losses in a motor driving a fan load is not necessary be maximum at the full-voltage. It can be shown that the condition for maximum rotor current is:

$$s = \frac{1}{3}$$

Hence, the motor should be appropriately derated in order to avoid overheat of the rotor winding.

2. *Constant flux control*

Why constant flux

The characteristics of an induction motor with fixed frequency excitation (Fig 4) have indicated that it only operates efficiently under low slip conditions. Therefore, for inverter fed variable frequency operation the inverter frequency must be kept close to the motor speed and varied to vary the motor speed. This produces a drive which operates efficiently over a wide speed range.

To achieve maximum torque production at all speeds, the motor flux loading should be maintained at its design level. The ideal criterion for variable frequency operation is to maintain constant flux.

$$E_1 = 4.44\, f_1 N \hat{\phi}_m$$

Therefore E_1/f_1 should be kept constant at all speed if the motor flux loading is maximised.

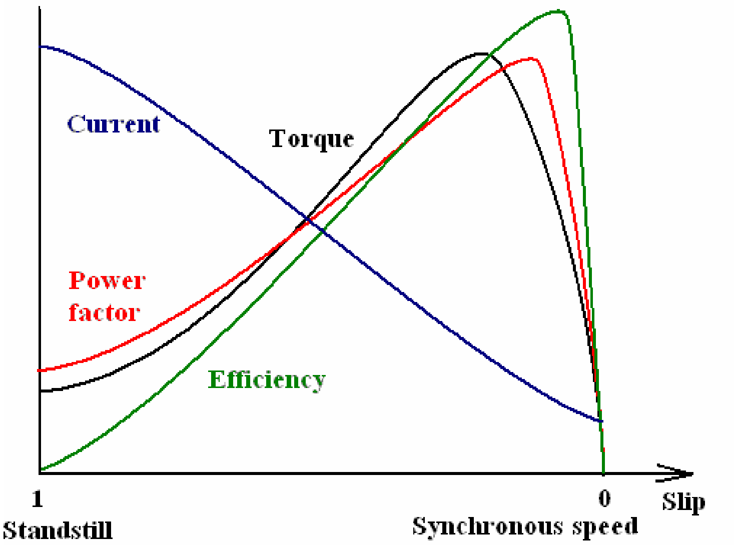

Fig 4 Classical characteristics of fixed frequency induction motor

V/F control

What is V/F control

The V/F control is a popular way to implement the constant flux operation of induction motor. Fig 6 shows a simplified method to implement the V/F control. ω_1* denotes the command speed of the motor. The phase voltage V_{ph} is calculated by the function generator V/F block. The disadvantages of this simple control system are:

1) This control is to use the rotor speed to approximate the supply frequency and hence the speed control is poor.
2) There is no feedback of rotor speed. The speed regulation is poor.
3) Although there is V/F control, but the flux cannot be ensured constant. This is especially important during the low speed operation. The magnetising current becomes significant and the voltage drop on the

stator winding becomes large comparing to the voltage across the magnetising inductance.

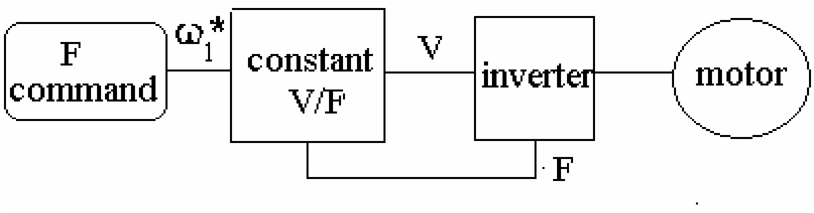

Fig 6.An open-loop V/F control of induction motor

Boost voltage

A boost voltage is therefore needed in order to compensate for the voltage drop in the stator winding during the low frequency operation. A typical V/F characteristic is shown in fig 7. The boost voltage V_b is depended on the induction motor and varies between 2% and 20%.

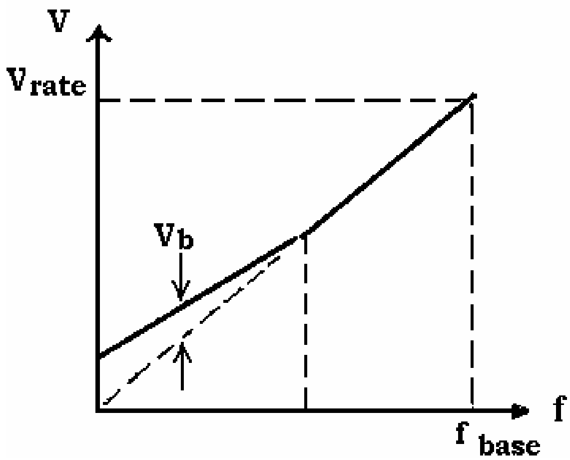

Fig 7. V/F characteristics with boost voltage

Closed loop V/F control

Closed V/F control can be implemented by measuring the speed of the motor. The speed can be measured by a tacho-generator or incremental encoder. (Control diagrams: review in the lecture)

3. *Field weakening and region of operation (Fig 8)*

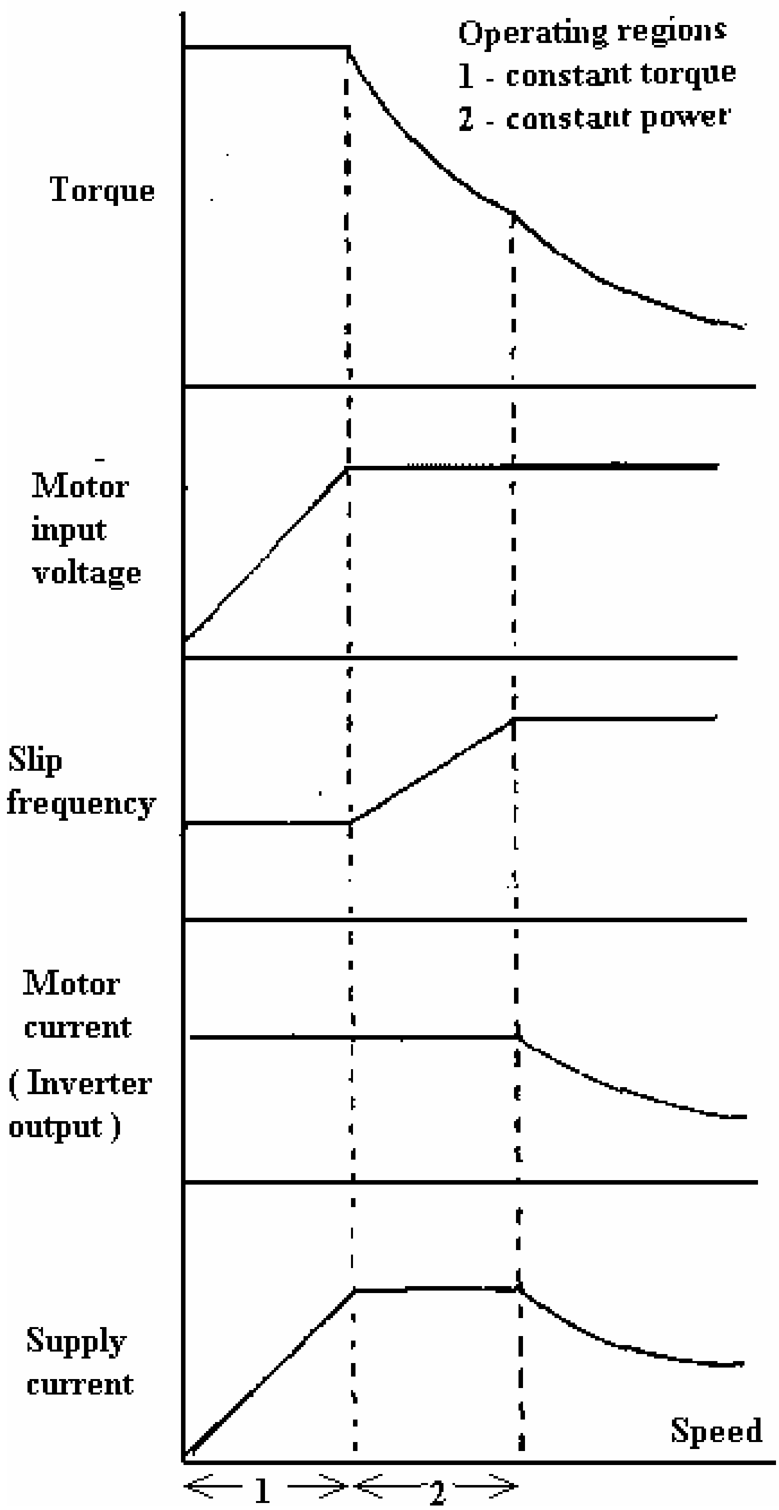

Fig 8. Idealised motor characteristics

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Region of operation

Region 1. Constant torque region:
This is the constant volts/frequency ratio region, apart from the voltage boost needed at low speed to overcome stator voltage drop. Ideally, this results in constant flux in the air-gap; hence with constant rotor current and slip frequency, constant torque is achieved.

Region 2. Constant power region
Once the motor speed is increased beyond its base speed, there is no further voltage increase available because it will exceed the rated voltage. Further increase of the motor speed is achieved by weakening the field.

The function generation of the V/F control diagram can now be modified such that the voltage is hold constant at the rated voltage beyond base speed. Further increase in speed will cause the motor to operate in field weakening region.

Plugging, Motoring and Regenerative braking

Motoring - The source frequency of the motor can be adjusted by changing the frequency demand of the inverter (the PWM pulse width equation). It should be noted that using PWM inverter, the frequency can be positive or negative in order to control the motor in forward or reverse rotation.

Plugging - The rotor rotates in the opposite direction to that of the air gap flux, ie S >

1. This condition could arise if the speed command is reverse when the rotor is still moving or because of an overhauling type of load which drives the rotor in the opposite direction. The electrical power supplied by the source and mechanical power generated by the load and inertia are dissipated in the motor circuit resistance. This method is not suitable for braking but quite suitable for reversing the motor.

Regenerative braking - When the motor runs at a speed greater than the synchronous, the relative speed between the rotating field and the rotor is negative ($S < 0$). The stator current will reverse the direction and the power flow becomes negative. The motor works as a generator and the DC link current will reverse. Fig 9 shows the speed- torque of an induction motor during motoring, plugging and regenerating.

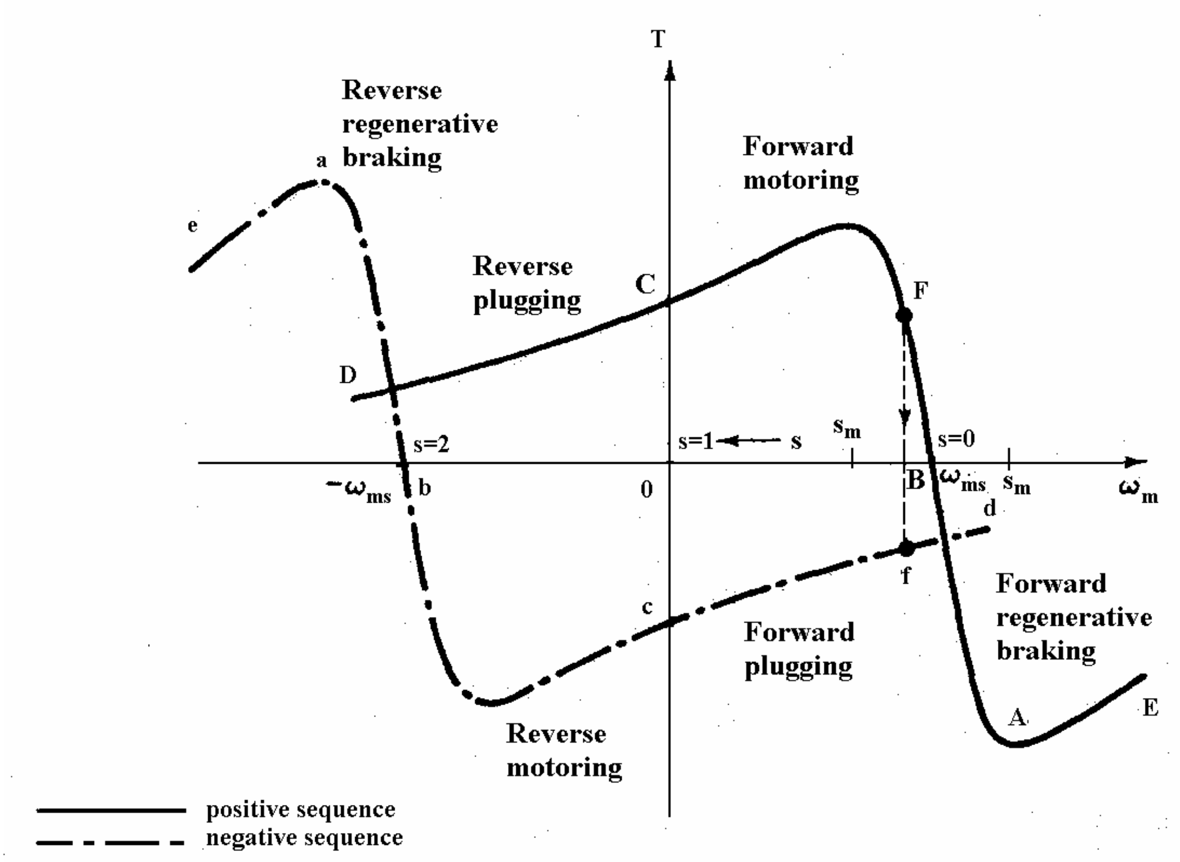

Fig 9 Speed-torque curves of an induction motor in forward and reverse direction

5. Wound -rotor induction motor drives

The wound motor

A wound-rotor induction motor has a three phase winding in the slots on its rotor. These windings are connected to the external circuits by three slip rings. A fraction, S, of the power is converted to the rotor circuit. In a squirrel-cage motor, this is dissipated as winding loss. For the wound-rotor motor, this power can be connected to the external circuit and regulate the torque and speed of the motor. Two methods can be used:

1) rotor resistance control
2) rotor-power recovery

Rotor resistance control

Fig 10 shows a rotor resistance control of a wound-rotor induction motor. The rotor circuit resistance can be varied by means of a chopper. Alternatively, the rotor resistance control circuit can be replaced by a 6-pulse controlled rectifier. The power consumed by the load can be regulated by the firing angle. This type of control method is obviously not very efficient because of extra power loss.

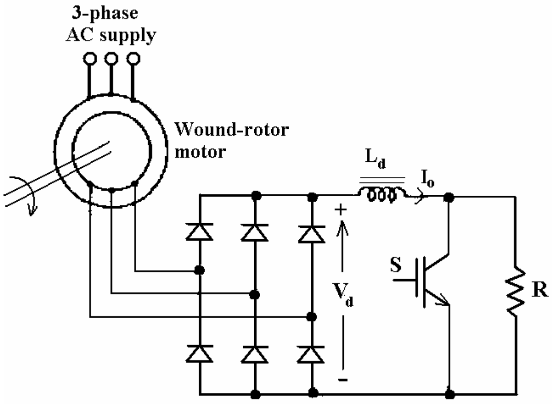

Fig 10. Rotor resistance control of wound-rotor induction motor

Rotor power recovery

Fig 11 shows a typical rotor-power recovery system, and is known as Scherbius drive. The power brought out from the rotor through the bridge diode is inverted to ac and is returned to the source. The advantages of this drive are: high efficiency, simple drive and very stable in operation. The disadvantages are: Wound-rotor motor is needed, overall power factor is low under rated operating conditions, limited speed range and static converter is needed.

The performance of the drive can be approximated as follow. Assume the voltage drops on the stator and rotor resistive and leakage impedance are negligible.

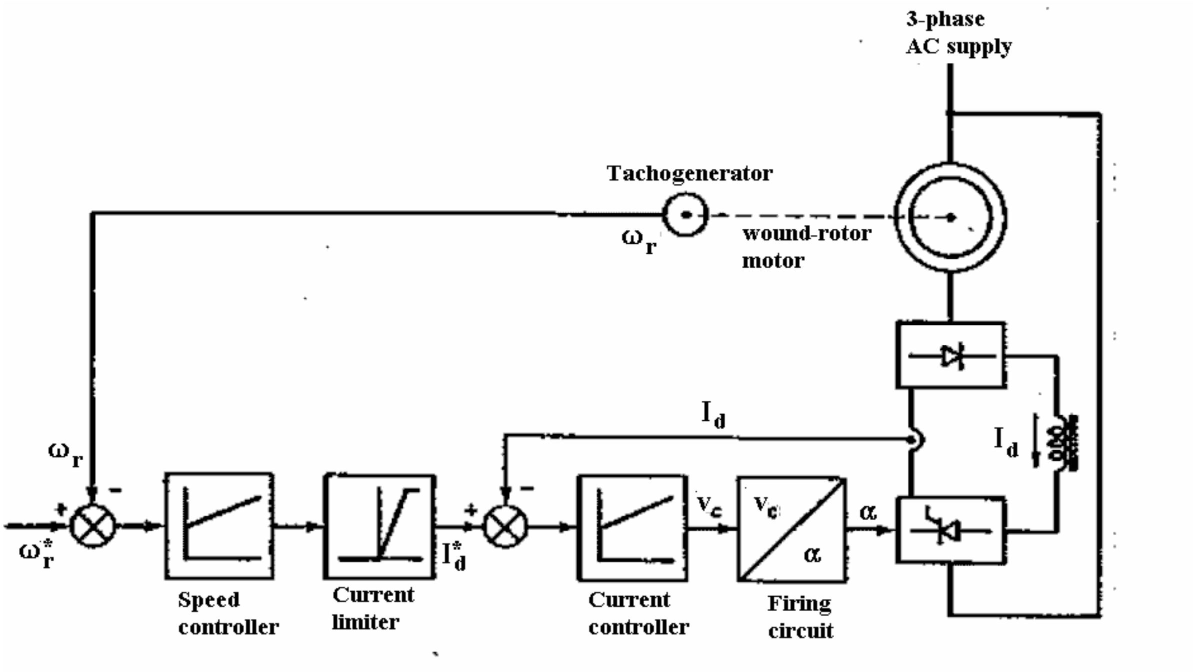

Fig 12. Closed-loop speed control of a Scherbius drive

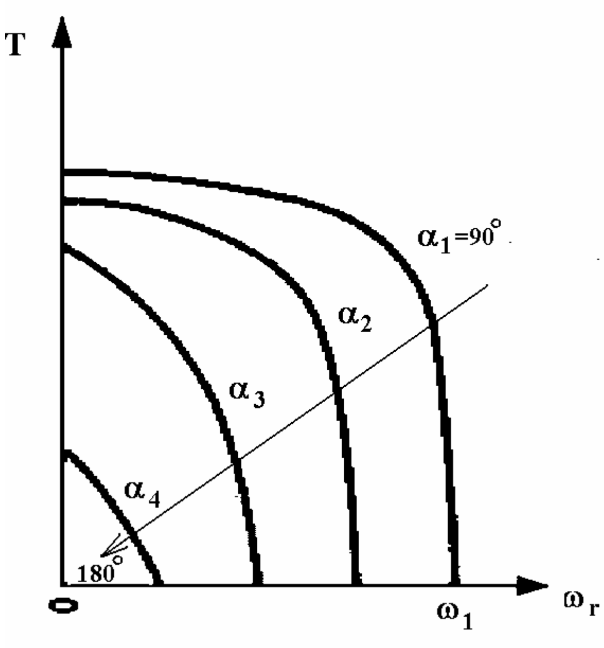

Fig 13. Nature of speed-torque curves of the Scherbius drive

6.Closed loop control of AC machines

Introduction

In general, there are many methods for formulating an induction motor drive system. A typical industrial drive can be identified as a speed control system. There are many applications in which the torque, current and position is controlled. In multiple loop control system, the inner loops are designed to be progressively faster. The primary objective of a drive system is to design transient response and steady state error of the control parameters within the desired specifications.

Summary of control methods

Stator voltage control with current limit

The block diagram of a stator voltage speed control system using thyristor ac control method is shown in Fig 14. The speed command is compared with the measured speed from the speed sensor. The speed regulator can be a PI controller or a lag-lead type $(1 + T_1S) / (1 + T_2S)$. An inner current limit control loop may be added if start up current transient is higher than the circuit limit but scarifying the acceleration time.

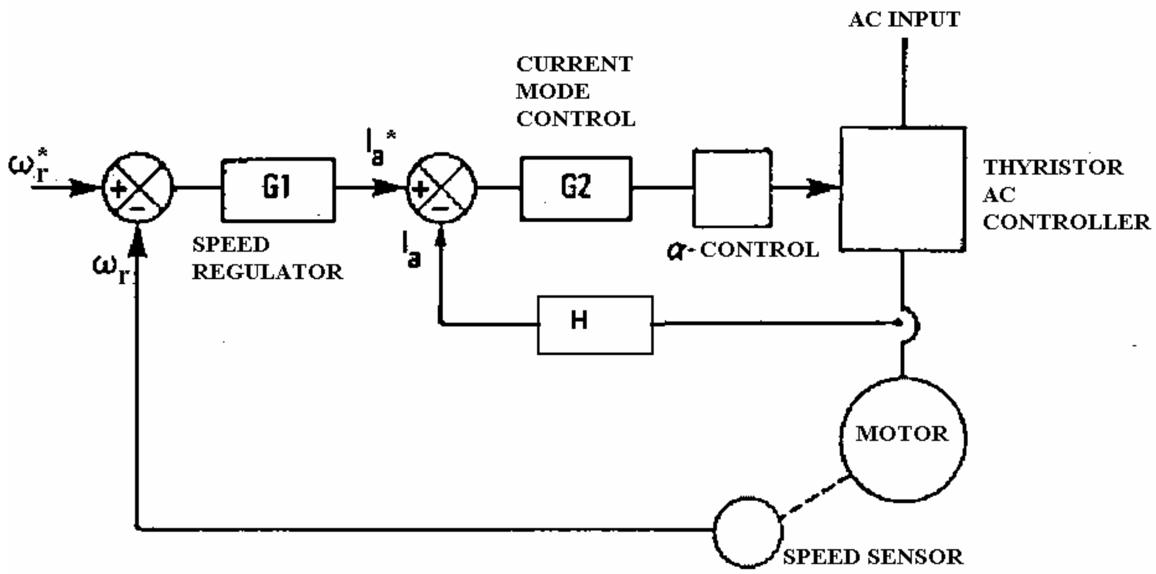

Fig 14. Stator voltage speed control with current limit

Constant flux control

A simple form of speed control of a voltage source inverter is the constant V/F control and has shown in section 3.3 and 3.4. An improved control system can be designed where the flux and torque are close loop controlled as shown in Fig. 15. The torque error generates the slip command which is added with the speed to generate the frequency command. The air- gap flux ψ_g is maintained constant in order to obtain steady-state efficiency optimization.

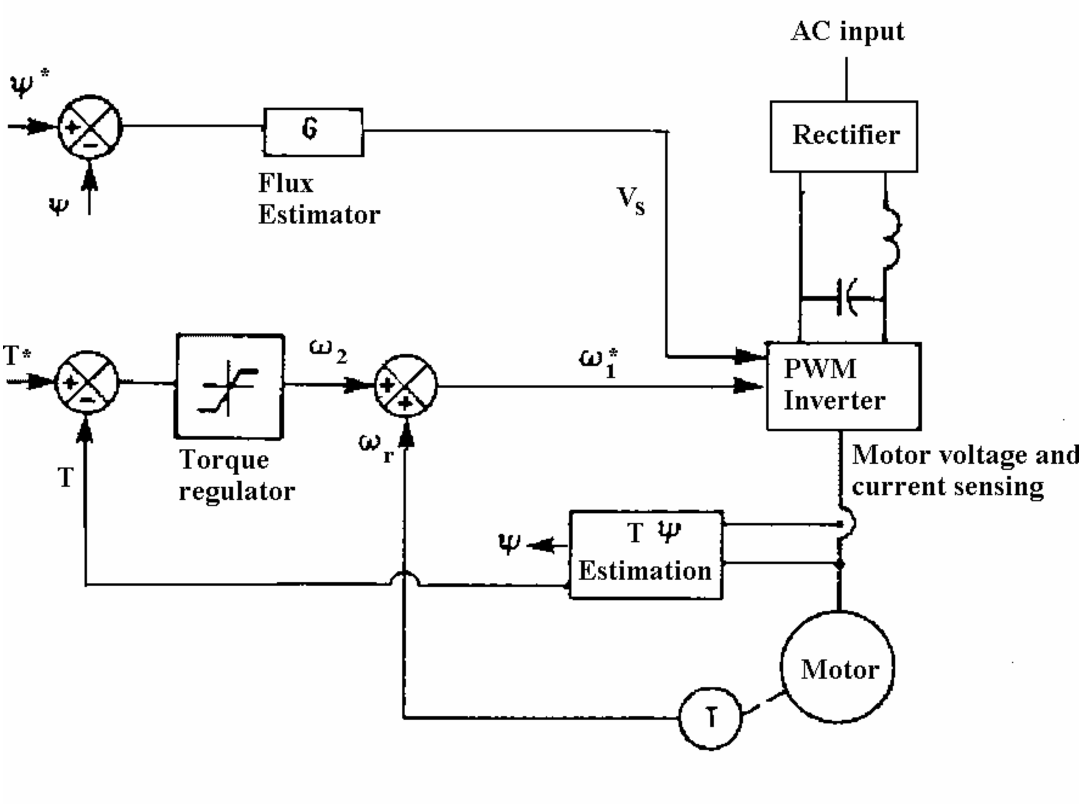

Fig. 15 Independent torque and air-gap flux control.

Current source inverter

The inverter mentioned in this chapter is the voltage source inverter. The input parameters to the inverter are the modulation index, M, which control the output voltage and the output frequency. The input source is a voltage and something is refereed as DC link voltage.

Another type of the inverters is called current source inverter. The input source is current. The input parameters are the stator current command. The machine is controlled continuously in current mode control.

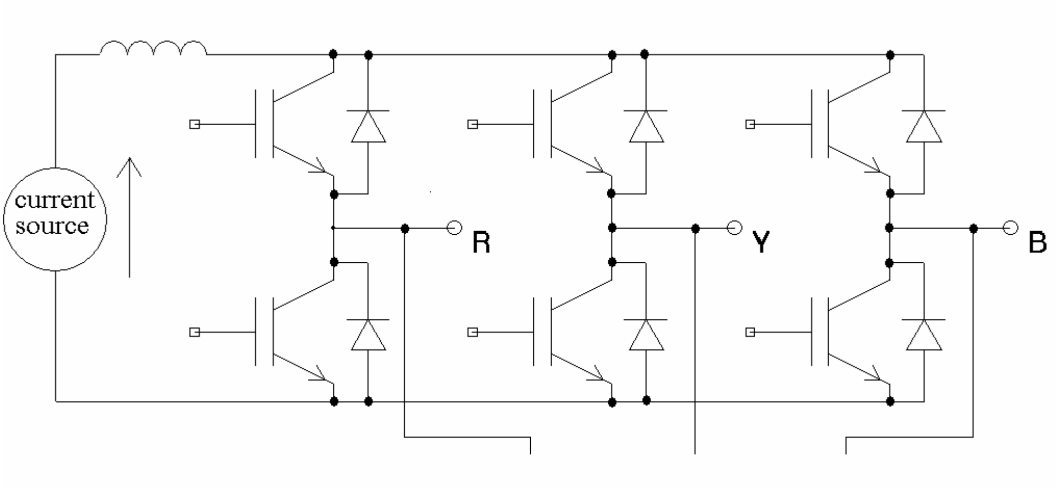

Fig 16: A typical diagram of current source inverter

Vector Control

Vector control was introduced about more than 10 years ago. The control method is based on the fact that both torque and airgap flux are controlled independently, so that it looks like the way to control a DC machine. The control variables which are originally in stationary reference frame, can be transformed to rotating so that the parameters appears as DC quantities. These DC quantities can be regulated easily.

The Vector Control can also be called as Field oriented control because the control parameters are transformed to a stator, rotor or air-gap flux reference frame. Fig 17 shows a block diagram of field oriented control diagram.

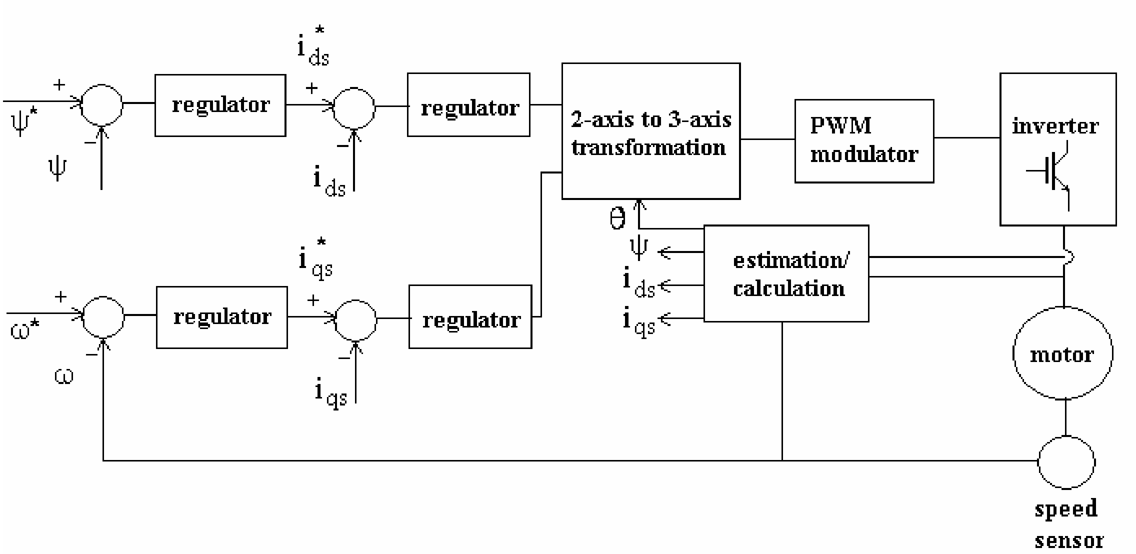

Fig. 17. A block diagram of field oriented control diagram

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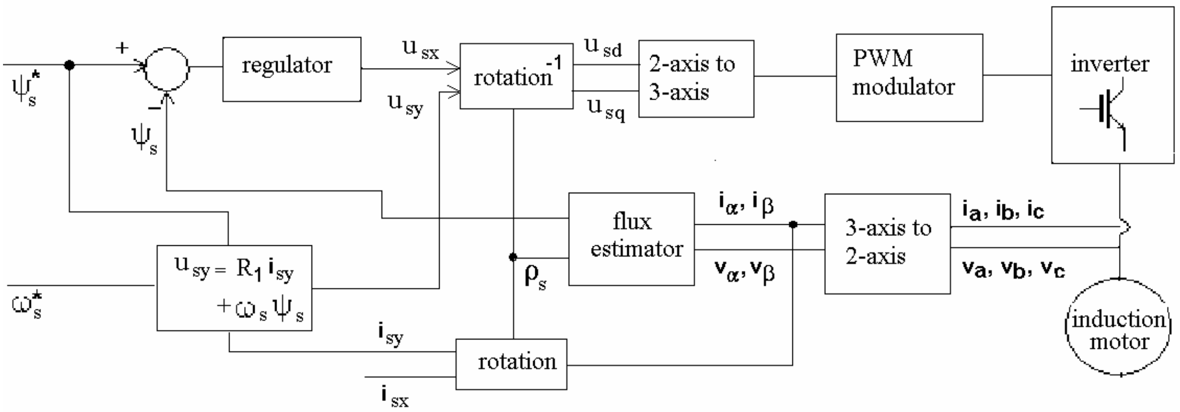


Fig 18 Simplified stator flux oriented control of induction motor

2 phase to 3 phase transformation

The 2 phase (d, q) quantities can be resolved into 3 phase (a,b,c) quantities by park transformation:

The control diagram

It can be see than the flux and torque are controlled independent as shown. The flux and torque component can be represented by x and y components. In a more detailed study, the x and y component are not purely decoupled. Some decoupling calculation is needed. In practice, the rotation from u_{sx} and u_{sy} back to u_{sd} and u_{sq} may not be necessary because the amplitude by can simply be obtained by :

$$u_s = \sqrt{{u_{sx}}^2 + {u_{sy}}^2}$$

and the frequency is simply derived from ρ_s.

Advantage and disadvantage of vector control

This method can controlled an AC machine by independently regulated the torque and flux components hence the dynamic response is very good. The application includes the servo drives and any industrial drives.

However, this method requires the measurement of the flux quantities and hence extra sensor is needed. If the measurement of flux is not possible due to the constraint of the application and environment, indirect method can be used. The flux can be determined by some forms of estimators which calculate the flux from the current, voltage and speed measurement.

In general, the vector control is parameter sensitive hence recently many researches have been concentrated on the motor parameter estimation and also the speed estimation.

7. Comparison of AC drives

	Stator voltage control	Square-wave inverter	PWM inverter
Speed range	Limited low speed	medium to high speed (10:1) ratio	Very wide speed range, up to very low speed
Harmonics	Rich harmonics in line and machine. Problem in motor heating	relative lower in low frequency harmonics	Very low in harmonics especially the switching frequency is high
Torque pulsation	High pulsating torque frequency	Problem at low speed	Minimal
Power factor	Poor in line power factor because of the phase control	low line power factor because of phase control	High power factor
Efficiency	Poor	Medium but improves with better transistors	Good and improves with better transistors
Cost	Low cost because simple in control circuit	Moderate ly but improvin g	Moderate ly because of the control circuits but improvin g
Application	Low to medium power blower and pump drives. Low power appliance	General purpose drives. Low to medium power	General purpose drive, medium to high power application, servo drives

General comments	Application is limited	Usually for open loop control. May have stability problem at low speed.	Fast transient response. popular drive

Most of us take the mains ac supply for granted and use it almost casually without giving the slightest thought to its inherent shortcomings and the danger posed to sophisticated and sensitive electronic instruments/equipments. For ordinary household appliances such as incandencent lamps, tubes, fans, TV and fridge, the mains ac supply does not make much difference, but when used for computers, medical equipments and telecommunication systems, a clean, stable interruption free power supply is of the utmost importance. Of the myriad of devices, processes and systems which rely on ac power, computers are probably the most sensitive to power disturbances and failures. Interruptions in power supply may cause the contents of a memory to be lost or corrupted, the entire system to malfunction or fail, or even variety of components failures to occur, all of which not only result in inconvenience but also loss of money.
As more and more PCs, word processors and data terminals find their way into small business, UPS systems that meet the power requirements and price range needs of even the small business organizations and offices are being manufactured.

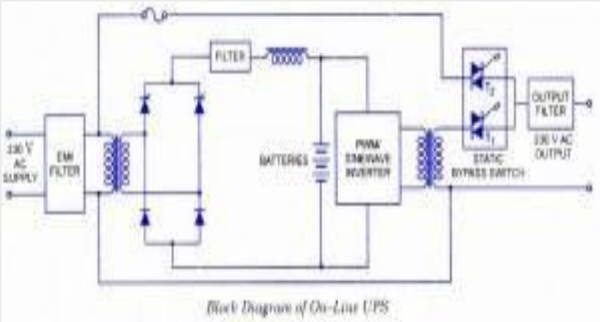
Uninterruptible Power Supply Systems.

There are three distinct types of uninterrupted power supplies, namely, (£) on-line UPS (ii) off-line UPS, and (Hi) electronic generators. In the on-line UPS, whether the mains power is on or off, the battery operated inverter is on all the time and supplies the ac output voltage. When the mains power supply goes off, the UPS will be on only until the battery gets discharged. When the main power resumes, the battery will get charged again. In off-line UPS and electronic generators, ther inverter is off when the mains power is present and the output voltage derived directly from the mains is the same as the mains supply voltage. The inverter turns on only when the mains supply goes off.
The block diagrams of on-line UPS, off-line UPS and electronic generators are given in figs The ever increasing importance of computers in industry and commerce will increase the need for quality, high stability and interruption free power supplies.

A clean ac power source is the fundamental to the operation of most sensitive electronic equipment, and many new and sophisticated circuits are designed to overcome the effects of disturbances normally found in the mains ac supply.
In order to protect a sensitive system from power losses and blackouts, an alternative power source is required that can switch into operation immediately when disruption occurs. An interruptible power supply (UPS) is just such an alternative source. A UPS generally consists of a rectifier, battery charger, a battery bank and inverter circuit which converts the commercial ac input into dc suitable for input to the battery bank and the inverter. The rectifier should have its input protected and should be capable of supplying power to the inverter when the commercial supply is either slightly below the normal voltage or slightly above.

Online UPS:

In case of On-line UPS, the battery operated inverter works continuously whether the mains supply is present or not. Triac T_1 is on for all the times while Triac T_2 has been provided to bypass the UPS inverter, only when a fault

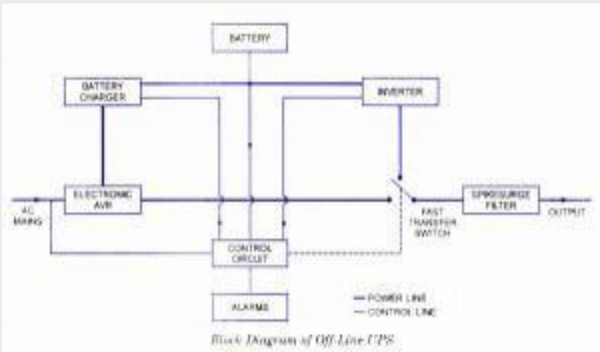


Online UPS Block Diagram

develops in the UPS inverter. When the mains supply fails, the UPS supplies power only until the batteries get discharged. However, once the mains power resumes, the batteries will get charged again. The switching times of these supplies is considered to be zero. Usually sealed maintenance free batteries are used and the running time of the inverter is low (approximately 10 to 30 minutes).

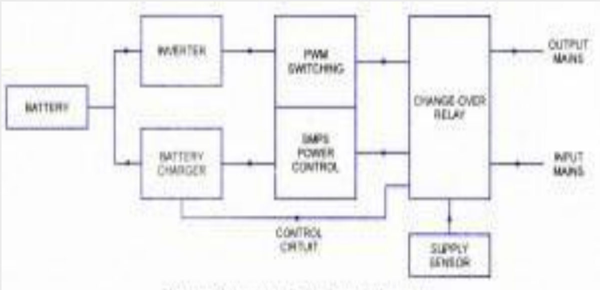
Off Line UPS:

In the case of Off-Line UPS, the inverter is off when the mains power is on and the output voltage is derived directly from the mains. The inverter turns on only when the mains supply fails. Its switching time is less than 5 ms. These UPS are generally used with PCs or computers or other appliances where a small duration (5 ms or less) interruption in power supply can be tolerated. Usually, sealed batteries or lead-acid batteries are used. The running time of these supplies is also low (about 10 to 30 minutes).



Offline UPS Block Diagram

Electronic Generators:



Block Diagram of an Electronic Generator

Electronic Generator

An electronic generator is the same as the off-line UPS system except for one difference that switching time from the mains supply to battery driven inverter supply will not be small (over 10 ms) for the electronic generator. Also, the electronic generators will run for longer time (1 to 4 hours) than off-line UPS systems because, usually large size lead-acid batteries are used with/electronic generators. These are meant for household applications to run fans, coolers, fridge, lights, TV and VCR.

The demand is the highest for the electronic generators meant for house hold applications, followed by the off-line UPS, and then the on-line UPS systems. The off-line or online UPS systems are mainly used in places where PCs or computers are used. The demand for on- line UPS systems is less than for off-line UPS systems because the price of the on-line UPS systems is higher.

Switched Mode Power Supply (SMPS)

Like a linear power supply, the switched mode power supply too converts the available unregulated ac or dc input voltage to a regulated dc output voltage. However in case of SMPS with input supply drawn from the ac mains, the input voltage is first rectified and filtered using a capacitor at the rectifier output. The unregulated dc voltage across the capacitor is then fed to a high frequency dc-to-dc converter. Most of the dc-to-dc converters used in SMPS circuits have an intermediate high frequency ac conversion stage to facilitate the use of a high frequency transformer for voltage scaling and isolation. In contrast, in linear power supplies with input voltage drawn from ac mains, the mains voltage is first stepped down (and isolated) to the desired magnitude using a mains frequency transformer, followed by rectification and filtering. The high frequency transformer used in a SMPS circuit is much smaller in size and weight compared to the low frequency transformer of the linear power supply circuit.

The ‘Switched Mode Power Supply’ owes its name to the dc-to-dc switching converter for conversion from unregulated dc input voltage to regulated dc output voltage. The switch employed is turned ‘ON’ and ‘OFF’ (referred as switching) at a high frequency. During ‘ON’ mode the switch is in saturation mode with negligible voltage drop across the collector and emitter terminals of the switch where as in ‘OFF’ mode the switch is in cut-off mode with negligible current through the collector and emitter terminals. On the contrary the voltage- regulating switch, in a linear regulator circuit, always remains in the active region.

Details of some popular SMPS circuits, with provisions for incorporating high frequency transformer for voltage scaling and isolation, have been discussed in next few lessons. In this lesson a simplified schematic switching arrangement is described that omits the transformer action. In fact there are several other switched mode dc-to-dc converter circuits that do not use a high frequency transformer. In such SMPS circuits the unregulated input dc voltage is fed to a high frequency voltage chopping circuit such that when the chopping circuit (often called **dc to dc chopper**) is in ON state, the unregulated voltage is applied to the output circuit that includes the load and some filtering circuit. When the chopper is in OFF state, zero magnitude of voltage is applied to the output side. The ON and OFF durations are suitably controlled such that the average dc voltage applied to the output circuit equals the desired magnitude of output voltage. The ratio of ON time to cycle time (ON + OFF time) is known as duty ratio of the chopper circuit. A high switching frequency (of the order of 100 KHz) and a fast control over the duty ratio results in application of the desired mean voltage along with ripple voltage of a very high frequency to the output side, consisting of a low pass filter circuit followed by the load. The high frequency ripple in voltage is effectively filtered using small values of filter capacitors and inductors.

SMPS versus linear power supply

As discussed above, in a linear regulator circuit the excess voltage from the unregulated dc input supply drops across a series element (and hence there is power loss in proportion to this voltage drop) whereas in switched mode circuit the unregulated portion of the voltage is removed by modulating the switch duty ratio. The switching losses in modern switches (like: MOSFETs) are much less compared to the loss in the linear element.

In most of the switched mode power supplies it is possible to insert a high frequency transformer to isolate the output and to scale the output voltage magnitude. In linear power supply the isolation and voltage-scaling transformer can be put only across the low frequency utility supply. The low frequency transformer is very heavy and bulky in comparison to the high frequency transformer of similar VA rating. Similarly the output voltage filtering circuit, in case of low frequency ripples is much bulkier than if the ripple is of high frequency. The switched mode circuit produces ripple of high frequency that can be filtered easily using smaller volume of filtering elements.

Linear power supply though more bulky and less efficient has some advantages too when compared with the switched mode power supply. Generally the control of the linear power supply circuit is much simpler than that of SMPS circuit. Since there is no high frequency switching, the switching related electro-magnetic interference (EMI) is practically absent in linear power supplies but is of some concern in SMPS circuits. Also, as far as output voltage regulation is concerned the linear power supplies are superior to SMPS. One can more easily meet tighter specifications on output voltage ripples by using linear power supplies.

Hybrid (SMPS followed by linear) power supply

A comparison of linear and switched mode power supplies tells about the advantages and disadvantages of the two. Linear power supply is highly inefficient if it has to work over large variations in input voltage, is more bulky because of the use of low frequency transformer and filter elements (inductors and capacitors). On the other hand linear power supplies give better output voltage regulation. It may sometimes be required to have output voltage regulation similar to the one provided by linear supplies and compactness and better efficiency of a switched mode supply. For this, the linear power supply may be put in tandem with a switched mode supply. Let us consider a case where one needs an isolated and well-regulated 5 volts output while input power is drawn from utility supply that has large voltage fluctuation. In such a situation one may generate an isolated 7.5 volts from an SMPS and follow it by a 5 volts linear power supply set to work with 7.5 volts input. The input to linear power supply must be few volts more than the required output (for proper biasing of the switches) and hence SMPS tries to maintain around 7.5 volts input. It can be seen that the linear power supply now does not have large input voltage variation in spite of large variations in the utility rms voltage. The SMPS portion of the power supply efficiently performs the job of voltage isolation and conversion from widely varying utility voltage to fairly regulated 7.5 volts dc. Under the given condition it may not be difficult to see that the overall efficiency of this hybrid power supply will lie between that of a SMPS and a linear supply. The overall cost may or may not increase even though two supplies in tandem are used. It is to be kept in mind that to achieve the same output voltage specification by an SMPS circuit alone, the control and filtering circuit may become more costly and complex (than the one used in the hybrid power supply unit). Similarly if the linear supply has to be designed for larger fluctuation in input voltage the component ratings, including heat-sink ratings, will be higher and may cost as much as the hybrid unit.

Multiple output SMPS

A single power supply unit may need to output several different voltages. The individual output voltages may have different ratings in terms of output current, voltage regulation and ripple voltages. These outputs may need isolation between them. Generally a common high frequency transformer links the input and output windings and in spite of output voltage feedback all the outputs can not have same regulation because of different loads connected to different outputs and hence different ohmic (resistive) drops in the output windings (loads are generally variable and user dependent). Also the coupling between the different secondary windings and the primary winding may not be same causing different voltage drops across the respective leakage inductances. Barring this mismatch in the voltage drops across the resistances and leakage inductances of the secondary windings their output voltages are in proportional to their turns ratios. The turns ratios are properly chosen to give fairly regulated individual output voltages (even if only one output voltage feedback is used for SMPS switch control). The output that needs to have tighter voltage regulation may be used for output voltage feedback. In case another output needs to have similarly tight regulation then that particular output may be passed through an additional linear regulator circuit as in the case of hybrid power supply circuit discussed in the previous section (Sec. 21.5).

Resonant Mode Power Supplies

Resonant mode power supplies are a variation over SMPS circuits where the switching losses are significantly reduced by adapting zero-voltage or zero-current switching techniques. In non- resonant mode SMPS circuits the switches are subjected to hard switching (during hard- switching, both the voltage and current in the switch are of considerable magnitude resulting in large instantaneous switching power loss). Efficiency of resonant mode power supplies is generally higher than non-resonant mode supplies.

Power supply specifications

Power supplies may have several specifications to be met, including their voltage and current ratings. There may be short time ratings of higher magnitudes of current and continuous ratings of somewhat lower magnitudes. One needs to specify the tolerable limits on the ripple voltages, short-circuit protection level of current (if any) and the nature of output volt-current curve during over-current or short circuit (the output voltage magnitude should reduce or fold back towards zero, gradually, depending on the severity of over-current). The fuse requirement (if any) on the input and the output side may need to be specified. One needs to specify the type of input supply (whether ac or dc) or whether the power supply can work both from ac or dc input voltages. Acceptable range of variation in input voltage magnitude, supply frequency (in case of ac input) are also to be specified. Efficiency, weight and volume are some other important specifications. Some applications require the electro-magnetic compatibility standards to be met. By electromagnetic compatibility it is meant that the level of EMI generation by power supply should be within tolerable limits and at the same time the power supply should have the ability to work satisfactorily in a limited noisy environment. It is quite common to have output voltage isolation and it is specified in terms of isolation breakdown voltage. In case of multiple power supplies it needs to be specified whether all the outputs need to be isolated or not and what should be the acceptable ripple voltage range for each.

In majority of the cases the available source of input power is the alternating type utility voltage of 50 or 60 Hz. The voltage levels commonly used are 115V (common in countries like, USA) and 230 volts (common in India and many of the European countries). Most utility (mains) power supplies are expected to have 10% voltage regulation but for additional precaution the SMPS circuits must work even if input voltages have 20% variation. Now-a-days universal power supplies that work satisfactorily and efficiently both on 115 V and 230 V input are quite popular. These power supplies are very convenient for international travelers who can simply plug-on their equipments, like laptop computer and shaving machine, without having to pay much attention on the exact voltage and frequency levels of the utility supply. In contrast some of the other power supplies have a selector switch and the user is required to adjust the switch position to match the utility voltage. In case user forgets to keep the selector switch at correct position, the equipment attached may get damaged.

Some common types of SMPS circuits

There are several different topologies for the switched mode power supply circuits. Some popular ones are: fly-back, forward, push-pull, half bridge and H-bridge circuits. Some of these configurations will be discussed in the coming lessons. A particular topology may be more suitable than others on the basis of one or more performance criterions like cost,

efficiency, overall weight and size, output power, output regulation, voltage ripple etc. All the topologies listed above are capable of providing isolated voltages by incorporating a high frequency transformer in the circuit. There are many commercially available power supply controller ICs that can readily be used to control the duty ratio of the SMPS switches so that the final output is well regulated. Most of these ICs are capable of driving MOSFET type of switches. They also provide features like under voltage lock-out, output over-current protection etc.