equipment involves interactions between the source and the load, and utilises small-signal electronic control circuits as well as power semiconductor devices. Therefore, power electronics draws as well as depends upon all other areas of electrical engineering.

Power electronics constitute a vast, complex and interdisciplinary subject that has gone through rapid technological evolution during the last four decades. As the technology is advancing and apparatus cost is decreasing along with the improvement of reliability, their applications are expanding in industrial, commercial, residential, military, aerospace and utility environments. Many innovations in power semiconductor devices, converter topologies, analytical and simulation techniques, electrical machine drives, and control and estimation techniques are contributing to this advancement. The frontier of the technology has been further advanced by the artificial intelligence (AI) techniques, such as fuzzy logic and artificial neural networks, thus bringing more challenge to power electronic engineers.

In the global industrial automation, energy generation, conservation of the 21st century, the widespread impact of power electronics is inevitable. In this chapter, we will overview the power devices, converters and applications of power electronics.

1.2 HISTORY OF POWER ELECTRONICS DEVELOPMENT

Until 1956, the application of semiconductors was confined to low power circuits and electronic engineering was also called as light current engineering. In September 1956, four engineers of the Bell Telephone Laboratory, USA, published a paper entitled "*PNPN* transistor switches" in the proceedings of the Institute of Radio Engineers. This paper triggered intensive research on *PNPN* devices. In 1957, Gordon Hall of General Electric Company, USA, developed the three terminal *PNPN* silicon based semiconductor device called as *silicon controlled rectifier* (SCR). Continuous modifications and improvement in its design as well as fabrication techniques have made it more and more economical and suitable for various control purposes. Later on, many other power devices having characteristics similar to that of an SCR were developed. Actually, the origin of power electronics can be traced back to the time when mercury arc devices were employed for the rectification of a.c. to d.c. or the inversion of d.c. to a.c. However, the rapidly increasing usage of power electronics nowadays has resulted from the development of solid state power devices.

1.3 POWER ELECTRONIC SYSTEMS

Block diagram of the generalised power electronics system is shown in Fig. 1.1 Power source may be an ac supply system or a dc supply system. In India, 1-phase and 3-phase 50 Hz ac supplies are readily available in most locations. Very low power drives (systems employed for motion control are called drives) are generally fed from 1-phase source. Rest of the drives are powered from 3-phase source. Low and medium power motors (tens of kilowatts) are generally fed from 400 V supply; for high ratings, motors may be rated at 3.3 kV, 6.6 kV, 11 kV and higher. In case of aircraft and space applications, 400 Hz ac supply is generally used to achieve high power to weight ratio for motors. In main line traction, a high voltage supply is preferred because of economy. In India, 25 kV, 50 Hz supply is employed.

Some loads are powered from a battery, e.g. fork lift trucks and milk vans. Depending on size, battery voltage may have typical values of 6 V, 12 V, 24 V, 48 V and 110 V dc. Solar powered drives which are used in space and water pumping applications are fed from a low voltage dc supply. Presently, though these drives are very expensive but have a great future for rural water pumping and low power transport applications.

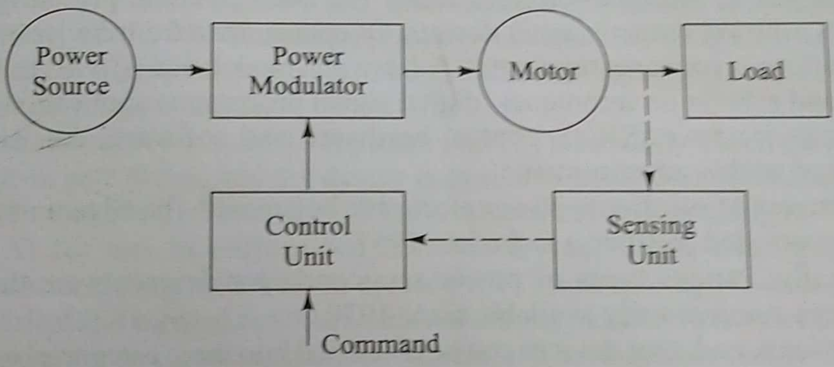


Fig. 1.1 *Block diagram of power electronic system*

Power modulator performs one or more of the following four functions:

(i) Converts electrical energy of the source as per the requirement of the load. For example, if the load is a dc motor, the modulator output must be adjustable direct voltage. In case the load is a 3-phase induction motor, the modulator may have adjustable voltage and frequency at its output terminals. When power modulator performs this function, it is known as converter.

(ii) Selects the mode of operation of the motor, i.e. motoring or braking.

(iii) Modulates flow of power from the source to the motor in such a manner that motor is imparted speed-torque characteristics required by the load.

(iv) During transient operations, such as starting, braking and speed reversal, it restricts source and motor currents within permissible values; excessive current drawn from source may overload it or may cause a voltage dip.

Motors commonly used in power electronic systems are:

(i) DC motors (shunt, series, compound and permanent magnet)

(ii) Induction motors (squirrel-cage, wound rotor and linear)

(iii) Synchronous motors (wound field and permanent magnet)

(iv) Brushless dc motors

(v) Stepper motors and

(vi) Switched reluctance motors

Power modulators are controlled by a control unit. Nature of the control unit for a particular system depends on the power modulator that is used. Control unit operates at much lower voltage and power levels. Sensing unit measures the load parameters, say speed in case of a rotating machine and compares it with the command. The difference of the two parameters processed by the control unit components now controls the turn-on of power semiconductor devices which are used in power modulators. As desired, the behaviour of the load circuit can be controlled over a wide range with the adjustment of the command.

1.5 POWER ELECTRONIC CONVERTERS

The great strides taken in the industrial applications of power electronics during recent years have demonstrated that this versatile tool can be of great importance in increasing production, efficiency and control. Power Electronic Circuits are also called as power converters. A converter uses a matrix of power semiconductor switches to convert electrical power at high efficiency. The converter system is comprised of switches, reactive components L, C, and transformers. Switches include two terminal devices such as diodes and three terminal devices such as transistors or thyristors. These converters/controllers are generally classified into the following five broad categories:

1. Phase Controlled Rectifiers (AC to DC Converters) These controllers convert fixed ac voltage to a variable dc output voltage. These converters takes power from one or more ac voltage/current sources of single or multiple phases and delivers to a load. The output variable is a low-ripple dc voltage or dc current. These controller circuits use line voltage for their commutation. Hence they are also called as line commutated or naturally commutated ac to dc converters. These circuits include diode rectifiers and single/three phase controlled circuits. These controllers are discussed in detail in Chapter 6.

Applications: High voltage dc transmission systems
DC motor drives
Regulated dc power supplies
Static VAR compensator
Wind generator converters
Battery charger circuits

2. Choppers (DC to DC Converter) A chopper converts fixed dc input voltage to a variable dc output voltage. The dc output voltage may be different in amplitude than the input source voltage. Choppers are designed using semiconductor devices such as power transistors, IGBTs, GTOs, Power MOSFETs and thyristors. Output voltage can be varied steplessly by controlling the duty ratio of the device by low power signals from a control unit. Chopper has either a battery, a solar powered dc voltage source or a line frequency (50–60 Hz) derived dc voltage source. Choppers are discussed in Chapter 8.

Applications: DC drives
Subway cars
Battery driven vehicles
Electric traction
Switch mode power supplies

3. Inverters (DC to AC Converter) An inverter converts a fixed dc voltage to an ac voltage of variable frequency and of fixed or variable magnitude. A practical inverter has either a battery, a solar powered dc voltage source or a line frequency (50 Hz) derived dc voltage source (often unregulated). Inverters are widely used from very low-power portable electronic systems such as the flashlight discharge system in a photography camera to very high power industrial systems.

Inverters are designed using semiconductor devices such as power transistors, MOSFETs, IGBTs, GTOs and thyristors. Chapter 9 deals with the study of inverters in detail.

Applications: Uninterruptible power supply (UPS)
Aircraft and space power supplies
Induction and synchronous motor drives
High voltage dc transmission system
Induction heating supplies

4. Cycloconverters (AC to AC Converters) These circuits convert input power at one frequency to output power at a different frequency through one stage conversion. These are designed using thyristors and are controlled by triggering signals derived from a control unit.

The output frequency is lower than the source frequency. Output frequency in cycloconverter is a simple fraction such as $\frac{1}{3}, \frac{1}{5}$ and so on of the source frequency. These are mainly used for slow speed, very high power industrial drives. Cycloconverters are discussed in Chapter 10.

Applications: AC drives like rotary kilns multi-MW ac motor drives.

5. AC Voltage Controllers (AC Regulators) These converters convert fixed ac voltage directly to a variable ac voltage at the same frequency using line commutation. These converters employs a thyristorised voltage controller. Stepless control of the output voltage can be obtained by controlling firing angle of converter thyristors by low power signals from a control unit. This type of converters are briefly discussed in Chapter 11.

Applications: Lighting control
Speed control of large fans and pumps
Electronic tap changers

1.6 POWER ELECTRONIC APPLICATIONS

The importance of power electronics in industrial automation, energy systems, energy generation and conservation, and indirectly for environmental pollution control is tremendous. As the technology is maturing and cost is decreasing, power electronics is expanding in applications, such as switch mode power supplies (SMPS), UPS systems, electrochemical processes, heating and lighting, static VAR compensation, active filtering, high voltage dc system, photo-voltaic system, and variable frequency motor drives. The motor drives possibly constitute the most fascinating and complex applications of power electronics where the applications include computer peripherals, servos and robotics, pumps and fans, paper and textile mills, rolling mills, wind generation system, variable speed heat pump and air-conditioning, transportation system, ship propulsion etc.

The importance of power electronics is being increasingly visible now-a-days in the energy saving of electrical apparatus by more efficient use of electricity. The energy consumption in the world is increasing by leaps and bounds to improve the human living standard, particularly in industrialized countries. The major amount

of this energy comes by burning
which create global warming effe
efficiency improvement of electri
not only reduce electricity cons
generation indirectly helps redu
been estimated that roughly 1
extensive application of power
power-electronics. However, t

Table 1.2 Applications of p

Sectors
1. Home Appliances
2. Games and entertainment
3. Commercial
4. Aerospace
5. Automotive
6. Industrial
7. Medical
8. Security systems
9. Telecommunic
10. Transportatic
11. Utility systems

of this energy comes by burning fossil fuels, such as coal, natural gas and oil which create global warming effect besides urban pollution problem. The energy efficiency improvement of electrical apparatus with the help of power electronics not only reduce electricity consumption but the corresponding reduced power generation indirectly helps reduction of environmental pollution problem. It has been estimated that roughly 15% of electricity consumption can be saved by extensive application of power electronics. Table 1.2 list various applications of power-electronics. However, this list is not exhaustive.

Table 1.2 Applications of power-electronics in various sectors

Sectors	*Applications*
1. Home Appliances	Refrigerators, sewing machines, photography, airconditioning, food warming trays, washing machines, lighting, dryers, vaccum cleaners, electric blankets, grinders and mixers, cooking appliances
2. Games and entertainment	Games and toys, televisions, movie projectors
3. Commercial	Advertising, battery chargers, blenders, computers, electric fans, electronic ballasts, hand power tools, photocopiers, vending machines, light dimmers
4. Aerospace	Aircraft power systems, space vehicle power systems, satellite power systems
5. Automotive	Alarms and security systems, electric vehicles, audio and Rf amplifiers, regulators
6. Industrial	Blowers, boilers, chemical processing equipment, contactor and circuit breakers, conveyors, cranes and hoists, dryers, electric furnaces and ovens, electric, vehicles, electromagnets, electronic ignitions, elevators, flashers, gas-turbine starters, generator exciters, induction heating, linear induction motion control, machine tools, mining power equipments, motor drives and starters, nuclear reactor control, oil-well drilling equipment, paper mill machinery, power-supplies, printing press machinery, pumps and compressors, servo systems, steel mill instrumentation, temperature controls ultrasonic generators, uninterruptible power supplies (UPS), welding equipment
7. Medical	Fitness machines, laser power supplies, medical instrumentation
8. Security systems	Alarms and security systems, radar/sonar
9. Telecommunications	Uninterruptible power supplies (UPS), solar power supplies, VLF transmitters, wireless communication power supplies
10. Transportation	Magnetic levitation, trains and locomotives, motor drives, trolly buses, subways
11. Utility systems	VAR compensators, power factor correction, static circuit breakers, supplementary energy systems (solar, wind)

voltage gradient across its depletion layer. This phenomenon is known as the *Avalanche breakdown*. Since the other junctions, J_1 and J_3 are already forward biased, there will be a free carrier movement across all the three junctions resulting in a large amount of current flowing through the device from anode to cathode. Due to the flow of this forward current, the device starts conducting and it is then said to be in the *conducting state* or on state.

2.3 STATIC ANODE–CATHODE CHARACTERISTICS OF SCR

An elementary circuit diagram for obtaining static V–I characteristics of a thyristor is shown in Fig. 2.2. Here, the anode and cathode are connected to the main source through a load. The gate and cathode are fed from another source E_g.

The static V–I characteristic of an SCR is shown in Fig. 2.3. Here, V_a is the anode–cathode voltage and I_a is the anode current. The thyristor V-I characteristics is divided into three regions of operation. These three regions of operation are described below.

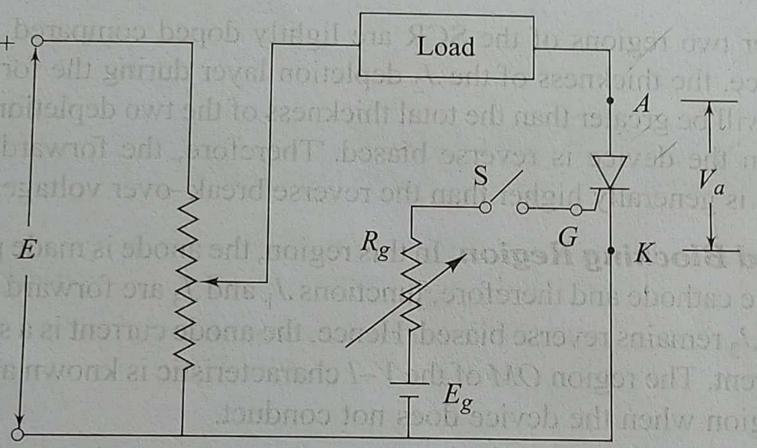
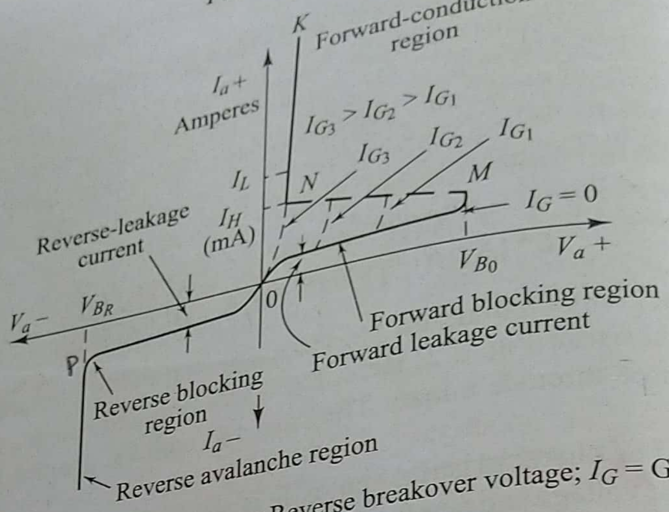


Fig. 2.2 *Elementary circuit*

1. Reverse Blocking Region When the cathode is made positive with respect to anode with the switch *s* open (Fig. 2.2), the thyristor becomes reverse biased. In Fig. 2.3, *OP* is the reverse blocking region. In this region, the thyristor exhibits a blocking characteristic similar to that of a diode. In this reverse biased condition, the outer junction J_1 and J_3 are reverse biased and the middle junction J_2 is forward biased. Therefore, only a small leakage current (in mA) flows. If the reverse voltage is increased, then at a critical breakdown level called reverse breakdown voltage V_{BR}, an avalanche will occur at J_1 and J_3 increasing the current sharply. If this current is not limited to a safe value, power dissipation will increase to a dangerous level that may destroy the device. Region *PQ* is the reverse-avalanche region. If the reverse voltage applied across the device is below this critical value, the device will behave as a high-impedance device (i.e., essentially open) in the reverse direction.



V_{Bo} = Forward breakover voltage; V_{BR} = Reverse breakover voltage; I_G = Gate current; I_L = Latching current; and I_H = Holding current

Fig. 2.3 *V-I characteristics*

The inner two regions of the SCR are lightly doped compared to the outer layers. Hence, the thickness of the J_2 depletion layer during the forward biased conditions will be greater than the total thickness of the two depletion layers at J_1 and J_3 when the device is reverse biased. Therefore, the forward break–over voltage V_{BO} is generally higher than the reverse break–over voltage V_{BR}.

2. Forward Blocking Region In this region, the anode is made positive with respect to the cathode and therefore, junctions J_1 and J_3 are forward biased while the junction J_2 remains reverse biased. Hence, the anode current is a small forward leakage current. The region *OM* of the *V–I* characteristic is known as the forward blocking region when the device does not conduct.

3. Forward Conduction Region When the anode to cathode forward voltage is increased with the gate circuit kept open, avalanche breakdown occurs at the junction J_2 at a critical forward break-over voltage (V_{BO}), and the SCR switches into a low impedance condition (high conduction mode). In Fig. 2.3, the forward breakover voltage is corresponding to the point *M*, when the device latches on to the conducting state. The region *MN* of the characteristic shows that as soon as the device latches on to its *ON* state, the voltage across the device drops from say, several hundred Volts to 1–2 Volt, depending on the rating of the SCR, and suddenly a very large amount of current starts flowing through the device. The part *NK* of the characteristic is called as the forward conduction state. In this high conduction mode, the anode current is determined essentially by the external load impedance. Therefore when the thyristor conducts forward current, it can be regarded as a closed switch.

When a gate-signal is applied, the thyristor turns-on before V_{BO} is reached. The forward voltage at which the device switches to ON state depends upon the magnitude of gate current; higher the gate current, lower is the forward breakover voltage. Figure 2.3 shows that for gate current $I_G = 0$, the forward breakover voltage is V_{BO}. For I_{G1}, the forward breakover voltage is less than V_{BO} and for $I_{G2} > I_{G1}$, it is still further reduced. In practice, the magnitude of gate-current is more than the minimum gate current required to turn-on the SCR. The typical gate current magnitudes are of the order of 20 to 200 mA.

Once the SCR is conducting a forward current that is greater than the minimum value, called the *latching current*, the gate signal is no longer required to maintain the device in its ON state. Removal of the gate current does not affect the conduction of the anode current. The SCR will return to its original forward blocking state if the anode current falls below a low level, called the *holding current* (I_h). For most industrial applications, this holding current (typically 10 mA) can be regarded as being essentially zero. Note that latching current is associated with turn-on process and holding current with turn-off process. The holding current is usually lower than, but very close to the latching current. Hence, from the above discussion it becomes clear that the more convenient, reliable and efficient method of turning on the device employs the gate drive.

is used, which provides for double-sided air or water cooling.

2.6 GATE CHARACTERISTICS OF SCR

In a thyristor, the gate is connected to the cathode through a *PN* junction and resembles a diode. Therefore, the *V-I* characteristic of a gate is similar to a diode but varies considerably in units. The circuit which supplies firing signals to the gate must be designed:

(1) to accommodate these variations,
(2) not to exceed the maximum voltage, and power capabilities of the gate,
(3) to prevent triggering from false signals or noise, and
(4) to assure desired triggering.

The design specification pertaining to gate characteristics are usually provided by the manufacturers. Figure 2.7 shows the gate characteristics of a typical SCR. Here, positive gate to cathode voltage V_g and positive gate to cathode current I_g represent d.c. values.

Applying gate drive increases the minority carrier density in the inner *P* layer and thereby facilitate the reverse breakdown of the junction J_2. There are maximum and minimum limits for gate voltage and gate current to prevent the permanent destruction of junction J_3 and to provide reliable triggering. Similarly, there is also a limit on the maximum instantaneous gate power dissipation ($P_{gmax} = V_g I_g$). The permissible maximum value of P_{gmax} depends on the type of gate drive. The gate signal can be d.c. or a.c. or a sequence of high frequency pulses. With pulse firing, a larger amount of instantaneous gate power-dissipation can be tolerated if the average-value of P_g is within the permissible limits. Hence, the gate can be driven harder (greater V_g and I_g) when pulse firing is used. This provides for reliable and faster turn-on of the device.

All possible safe operating points for the gate are bounded by the low and high current limits for the *V-I* characteristics, maximum gate voltage, and the hyperbola representing maximum gate power. Within these boundaries there are three regions of importance.

(1) The first region *OA* lies near the origin (shown hatched) and is defined by the maximum gate voltage that will not trigger any device. This value is obtained at the maximum rated junction temperature (usually 125°C). The gate must be operated in this region whenever forward bias is applied across the thyristor and triggering is not necessary. In other words, this region sets a limit on the maximum false signals that can be tolerated in the gate-firing circuit.

(2) The second region is further defined by the minimum value of gate-voltage and current required to trigger all devices at the minimum rated junction temperature. This region contains the actual minimum firing points of all devices. In a sense, it is a forbidden region for the firing circuit because a signal in this region may not always fire all devices or never fire any at all. In Fig. 2.7, *OL* and *OV* are the minimum gate-voltage and gate current limits respectively.

(3) The third region is the largest and shows the limits on the gate-signal for reliable firing. Ordinarily, a signal in the lower left part of this region is adequate for firing. For applications, where fast turn-on is required, a 'hard' firing signal in the upper right part of the region may be needed.

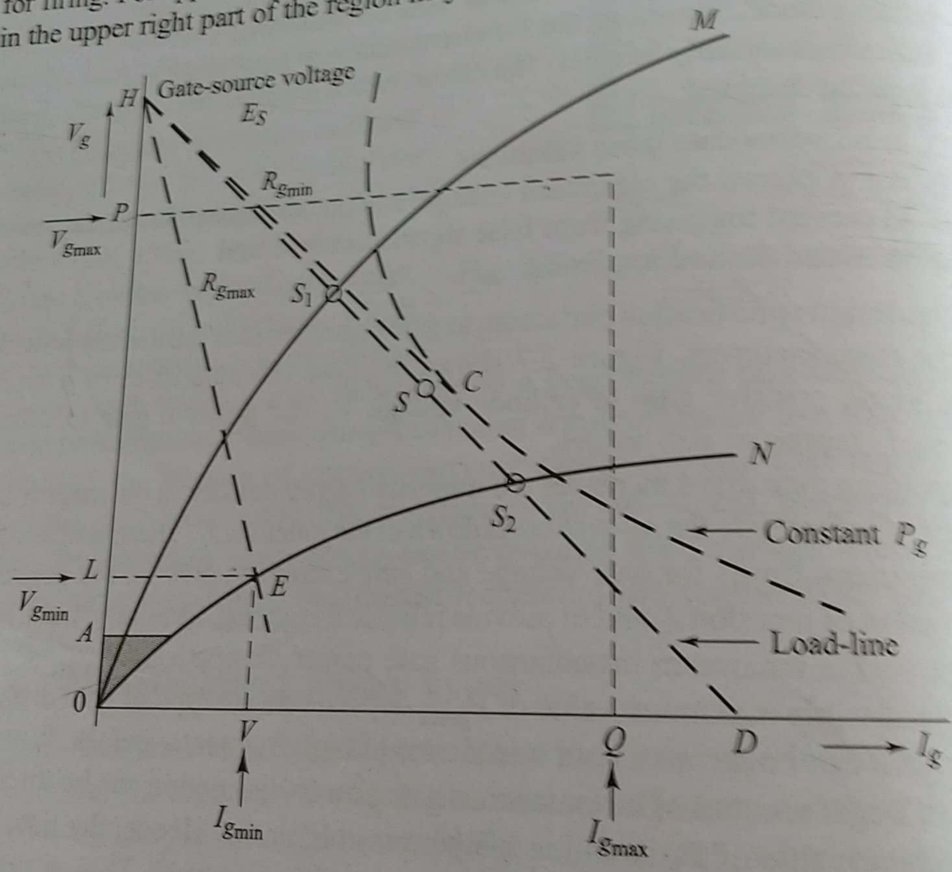


Fig. 2.7 *Gate characteristics*

In Fig. 2.7 curves *ON* and *OM* corresponds to the possible spread of the characteristic for SCRs of the same rating. For best results, the operating point *S*, which may change from S_1 to S_2, must be as close as possible to the permissible

P_g curve and must be contained within the maximum and minimum limits of gate voltage and gate current. This provides the necessary hard drive for the device. For selecting the operating point, usually a load line of the gate source voltage $E_s = OH$ is drawn as HD. The gradient of the load line HD (= OH/OD) will give the required gate source resistance R_g. The maximum value of this series resistance is given by the line HE, where E is the point of intersection of lines indicating the minimum gate voltage and gate current. The minimum value of gate source series resistance is obtained by drawing a line HC tangential to P_g curve.

A thyristor may be considered to be a charged controlled device. Thus, higher the magnitude of gate current pulse, lesser is the time needed to inject the required charge for turning on the thyristor. Therefore the SCR turn-on time can be reduced by using gate current of higher magnitude. It should be ensured that pulse width is sufficient to allow the anode current to exceed the latching current. In practice, the gate pulse width is usually taken as equal to or greater than SCR turn-on time, t_{on}. If T is the pulse width as shown in Fig. 2.8, then

$$T \geq t_{on}$$

With pulse firing, if the frequency of firing f is known, the peak instantaneous gate power dissipation P_{gmax} can be obtained as

$$P_{gmax} = V_g I_g = \frac{P_{g_{av}}}{fT} \tag{2.10}$$

where $f = \dfrac{1}{T_1}$ = frequency of firing or pulse repetition rate in Hz

and T = pulse width in second

A duty cycle is defined as the ratio of pulse-on period to the periodic time of pulse. In the Fig. 2.8 pulse-on period is T and the periodic time is T_1. Therefore, duty-cycle is given by

$$\delta = \frac{T}{T_1} = fT \tag{2.11}$$

From Eq. (2.10)

$$\frac{P_{g_{av}}}{\delta} \leq P_{gmax} \tag{2.12}$$

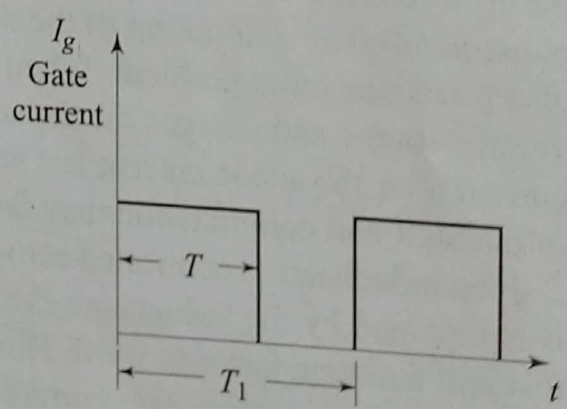


Fig. 2.8 *Pulse gating*

2.6.1 Gate Circuit Parameters

The gate cathode circuit with different circuit parameters ... series resistance ...

Power dissipation,

$$P_g = V_g I_g$$

$$= 8 \times 0.5 = 4 \text{ W}$$

Now, $P_{g\max} = \dfrac{P_{gav}}{f \cdot T_{on}}$ $\quad \therefore \quad 4 = \dfrac{0.3 \times 10^6}{f \times 4}$

$f = 18.75$ kHz $\quad \therefore \quad F \approx 19$ kHz.

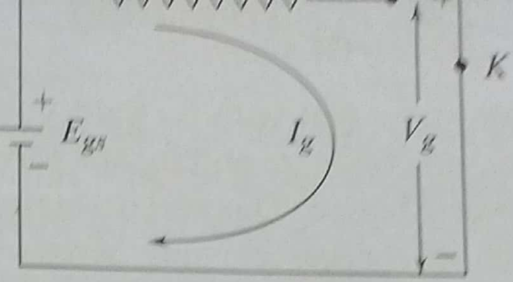


Fig. E2.8

2.7 TURN-ON METHODS OF A THYRISTOR

A thyristor can be switched from a nonconducting state to a conducting state in several ways described as follows.

2.7.1 Forward Voltage Triggering

When anode-to-cathode forward voltage is increased with gate circuit open, the reverse biased junction J_2 will have an avalanche breakdown at a voltage called forward breakover voltage V_{BO}. At this voltage, a thyristor changes from OFF state (high voltage with low leakage current) to ON-state characterised by a low voltage across it with large forward current. The forward voltage-drop across the SCR during the ON state is of the order of 1 to 1.5 V and increases slightly with load current.

2.7.2 Thermal Triggering (Temperature Triggering)

Like any other semiconductor, the width of the depletion layer of a thyristor decreases on increasing the junction temperature. Thus, in a thyristor when the voltage applied between the anode and cathode is very near to its breakdown voltage, the device can be triggered by increasing its junction temperature. By increasing the temperature to a certain value (within the specified-limits), a situation comes when the reverse biased junction collapses making the device conduct. This method of triggering the device by heating is known as the thermal triggering process.

2.7.3 Radiation Triggering (Light Triggering)

In this method, as the name suggests, the energy is imparted by radiation. Thyristor is bombarded by energy particles such as neutrons or photons. With the help of this external energy, electron-hole pairs are generated in the device, thus increasing the number of charge carriers. This leads to instantaneous flow of current within the device and the triggering of the device. For radiation triggering to occur, the device must have high value of rate of change of voltage (dv/dt). Light activated silicon controlled rectifier (LASCR) and light activated silicon controlled switch (LASCS) are the examples of this type of triggering.

2.7.4 $\frac{dv}{dt}$ Triggering

We know that with forward voltage across the anode and cathode of a device, the junctions J_1 and J_3 are forward biased, whereas the junction J_2 becomes reverse biased. This reverse biased junction J_2 has the characteristics of a capacitor due to charges existing across the junction. If a forward voltage is suddenly applied, a charging current will flow tending to turn the device ON. If the voltage impressed across the device is denoted by V, the charge by Q and the capacitance by C_j, then

$$i_c = \frac{dQ}{dt} = \frac{d}{dt}(C_j V) = C_j \frac{dV}{dt} + V\frac{dc_j}{dt} \qquad (2.13)$$

The rate of change of junction capacitance may be negligible as the junction capacitance is almost constant. The contribution to charging current by the later term is negligible. Hence, Eq. (2.13) reduces to

$$i_c = C_j \frac{dV}{dt} \qquad (2.14)$$

Therefore, if the rate of change of voltage across the device is large, the device may turn-on even though the voltage appearing across the device is small.

2.7.5 Gate Triggering

This is the most commonly used method for triggering SCRs. In laboratories, almost all the SCR devices are triggered by this process. By applying a positive signal at the gate terminal of the device, it can be triggered much before the specified breakover voltage. The conduction period of the SCR can be controlled by varying the gate signal within the specified values of the maximum and minimum gate currents.

For gate triggering, a signal is applied between the gate and the cathode of the device. Three types of signals can be used for this purpose. They are either d.c. signals, pulse signals or ac signals.

1. D.C. Gate Triggering In this type of triggering, a d.c. voltage of proper magnitude and polarity is applied between the gate and the cathode of the device in such a way that the gate becomes positive with respect to the cathode. When the applied voltage is sufficient to produce the required gate current, the device starts conducting.

One drawback of this sch
d.c. and there is no isolation
Another disadvantage of
applied, at the gate causing

2. A.C. Gate Triggerir
signal in all application of
scheme provides the prope
The firing angle control [
by changing the phase an
However, the gate dri
turned ON, and a revers
during the negative half
transformer is required

3. Pulse Gate Trig
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of this method is that
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2.8

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2.10. As shown,
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(i) ***Delay tim***
current reaches
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It can also be
0.9 V_a, where V
The gate cu
cathode surfac
decreases rapi
anode current
is the highest.

One drawback of this scheme is that both the power and control circuits are d.c. and there is no isolation between the two.

Another disadvantage of this process is that a continuous d.c. signal has to be applied, at the gate causing more gate power loss.

2. A.C. Gate Triggering a.c. source is most commonly used for the gate signal in all application of thyristor control adopted for a.c. applications. This scheme provides the proper isolation between the power and the control circuits. The firing angle control [discussed in Chapter 6] is obtained very conveniently by changing the phase angle of the control signal.

However, the gate drive is maintained for one half cycle after the device is turned ON, and a reverse voltage is applied between the gate and the cathode during the negative half cycle. The drawback of this scheme is that a separate transformer is required to step down the a.c. supply, which adds to the cost.

3. Pulse Gate Triggering This is the most popular method for triggering the device. In this method, the gate drive consists of a single pulse appearing periodically or a sequence of high frequency pulses. This is known as carrier frequency gating. A pulse transformer is used for isolation. The main advantage of this method is that there is no need of applying continuous signals and hence, the gate losses are very much reduced. Electrical isolation is also provided between the main device supply and its gating signals.

2.8 DYNAMIC TURN-ON SWITCHING CHARACTERISTICS

The static characteristics gives no indication as to the speed at which the SCR is capable of being switched from the forward blocking voltage to the conducting state and vice-versa. However, the transition from one state to the other does not take place instantaneously, it takes a finite period of time. This is illustrated in Fig. 2.10. As shown, the total turn-on time t_{on} of the SCR is subdivided into three distinct periods, called the *delay time*, *rise time* and *spread time*. These time periods are defined in terms of the waveforms of the anode voltage and current

... and the gate signal waveshapes.

The width of the firing pulse should, therefore, be more than 10 μs, preferably in the range of 20 to 100 μs. The amplitude of the gate-pulse should be 3 to 5 times the minimum gate current required to trigger the SCR.

From Fig. 2.10, it is noted that during rise-time, the SCR carries a large forward current and supports an appreciable forward voltage. This may result in high-instantaneous power dissipation creating local internal hot-spots which could destroy the device. It is, therefore, necessary to limit the rate of rise of current. Normally, a small inductor, called *di*/*dt* inductor is inserted in the anode circuit to limit the *di*/*dt* of the anode current.

The shadow area under the power-curve in Fig. 2.10 represents the switching loss of the device. This loss may be significant in high-frequency applications.

2.9 TURN-OFF MECHANISM (TURN-OFF CHARACTERISTIC)

Once the SCR starts conducting an appreciable forward current, the gate has no control on it 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. Process of turn-off is also called as commutation. Various methods used for turning off thyristors will be discussed in Section 2.10. However, if a forward voltage is applied immediately after reducing the anode current to zero, it will not block the forward voltage and will start conducting again, although it is not triggered by a gate pulse. It is, therefore, necessary to keep the device reverse biased for a finite period before a forward anode voltage can be reapplied.

The turn-off time of the thyristor is defined as the minimum time interval between the instant at which the anode current becomes zero, and the instant at which the device is capable of blocking the forward voltage. The turn-off time is illustrated by the waveforms shown in Fig. 2.11. The total turn-off time t_{off} is divided into two time intervals the reverse, recovery time t_{rr} and the gate recovery time t_{gr}.

At the instant t_1, the anode forward current becomes zero. During the reverse recovery time, t_1 to t_3, the anode current flows in the reverse direction. At the instant t_2, a reverse anode voltage is developed and the reverse recovery current continues to decrease. At t_3, junction J_1 and J_3 are able to block a reverse voltage. However, the thyristor is not yet able to block a forward voltage because carriers, called *trapped charges*, are still present at the junction J_2. During the interval t_3 to t_4, these carriers recombine. At t_4, the recombination is complete and therefore, a forward voltage can be reapplied at this instant. The SCR turn-off time is the interval between t_4 and t_1. In an SCR, this time varies in the range 10 to 100 μs. Thus, the total turn-off time (t_q) required for the device is the sum of the duration for which the reverse recovery current flows after the application of reverse voltage, and the time required for the recombination of all excess carriers in the

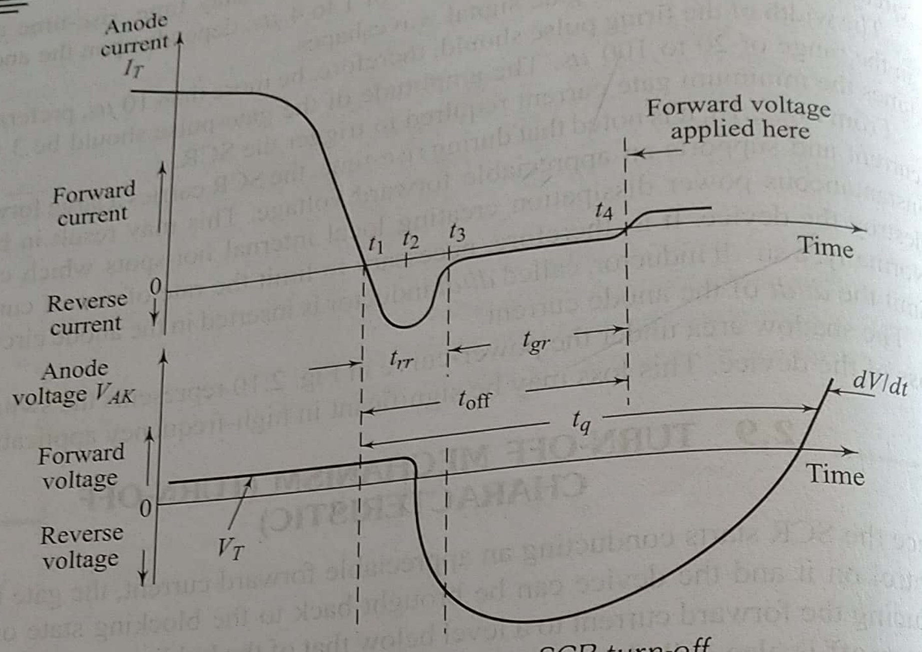


Fig. 2.11 *Waveforms during SCR turn-off*

inner two layers of the device. This may be noted that in case of highly inductive load circuit, the current cannot change abruptly at t_1. Also, the fast change in current at t_2 may give rise to high voltage surges in the inductance, which will then appears across the terminals of the thyristor.

In practical applications, the turn-off time required to the SCR by the circuit, called the circuit turn-off time t_q, must be greater than the device turn-off time t_{off} by a suitably safe margin, otherwise the device will turn-on at an undesired instant a process known as *commutation failure*. Thyristor having large turn-off time (50–100 μs) are called as slow switching or phase control type thyristors (or converter grade thyristors), and those having low turn-off time (10–50 μs) are called fast switching or inverter type thyristors. In high frequency applications, the required circuit turn-off time consumes an appreciable portion of the total cycle time and therefore, inverter grade thyristors must be used.

2.10 TURN-OFF METHODS

The term *commutation* basically means the transfer of current from one path to another. In thyristor circuits, this term is used to describe process of transferring current from one thyristor to another. As explained earlier, it is not possible for a thyristor to turn itself OFF; the circuit in which it is connected must reduce the thyristor current to zero to enable it to turn-off. 'Commutation' is the term to describe the methods of achieving this.

Commutation is one of the fundamental principles the use of thyristors for control purposes. A thyristor can only operate in two modes: it is either in the OFF state, i.e., open circuit, or in the ON state, i.e., short circuit. By itself it cannot control the level of current or voltage in a circuit. Control can only be achieved by variation in the time thyristors when switched ON and OFF, and commutation is central to this switching process. All thyristor circuits, therefore, involve the cyclic or sequential switching of thyristors. The two methods by which a thyristor can be commutated are as follows.

2.10.1 Natural Commutation

The simplest and most widely used method of commutation makes use of the alternating, reversing nature of a.c. voltages to effect the current transfer. We know that in a.c. circuits, the current always passes through zero every half cycle. As the current passes through natural zero, a reverse voltage will simultaneously appear across the device. This immediately turns-off the device. This process is called as *natural commutation* since no external circuit is required for this purpose. This method may use a.c. mains supply voltages or the a.c. voltages generated by local rotating machines or resonant circuits. The line commutated converters and inverters comes under this category.

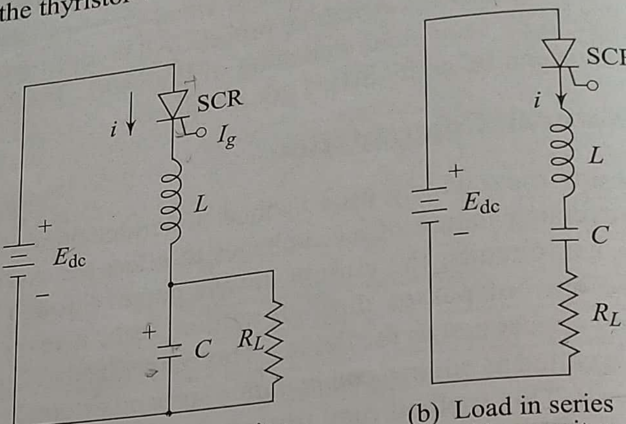
2.10.2 Forced Commutation

Once thyristors are operating in the ON state, carrying forward current, they can only be turned OFF by reducing the current flowing through them to zero for sufficient time to allow the removal of charged carriers. In case of d.c. circuits, for switching off the thyristors, the forward current should be forced to be zero by means of some external circuits. The process is called *forced commutation* and the external circuits required for it are known as commutation circuits. The components (inductance and capacitance) which constitute the commutating circuits are called as commutating components. A reverse voltage is developed across the device by means of a commutating circuit that immediately brings the forward current in the device to zero, thus turning off the device. Producing reliable commutation is a difficult problem to be tackled while designing chopper and inverter circuits. The most important stage in the designing process is choosing a forced turn-off method and deciding its components.

The classification of the methods of forced commutation is based on the arrangement of the commutating components and the manner in which zero current is obtained in the SCR. There are six basic methods of commutation by which thyristors may be turned OFF.

1. Class A–self Commutation by Resonating the Load This is also known as resonant commutation. This type of commutation circuit using *L-C* components-in-series-with the load are shown in Fig. 2.12. In Fig. 2.12(a), load R_L is in parallel with the capacitor and in Fig. 2.12(b) load R_L is in series with the

L–C circuit. In this process of commutation, the forward current passing through the device is reduced to less than the level of holding current of the device. Hence, this method is also known as the current commutation method. The waveforms of the thyristor voltage, current and capacitor voltages are shown in Fig. 2.13.



(a) Load in parallel with capacitor

(b) Load in series with capacitor

Fig. 2.12 *Class A commutation circuit*

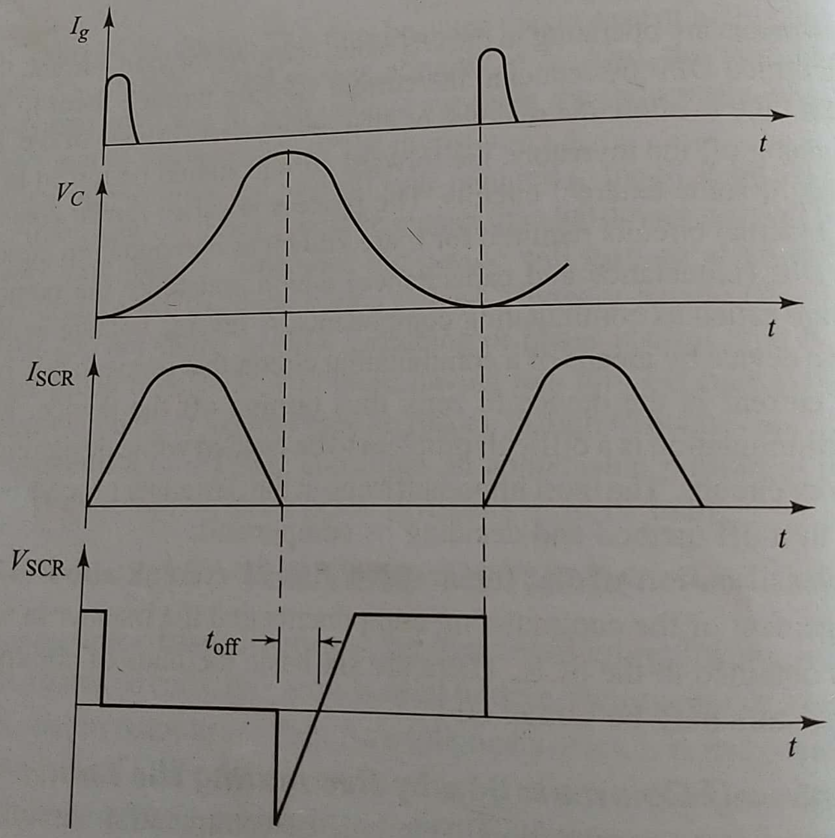
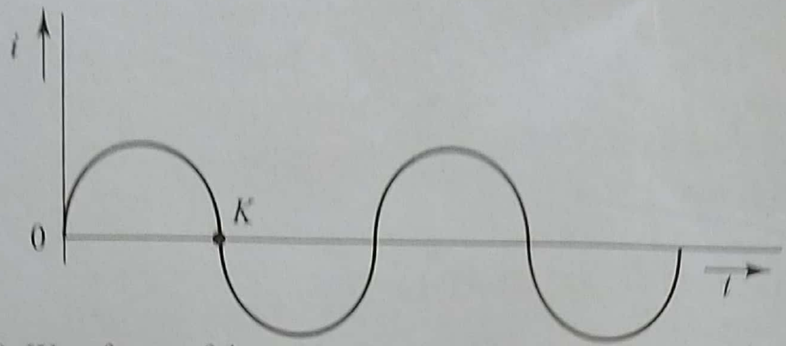
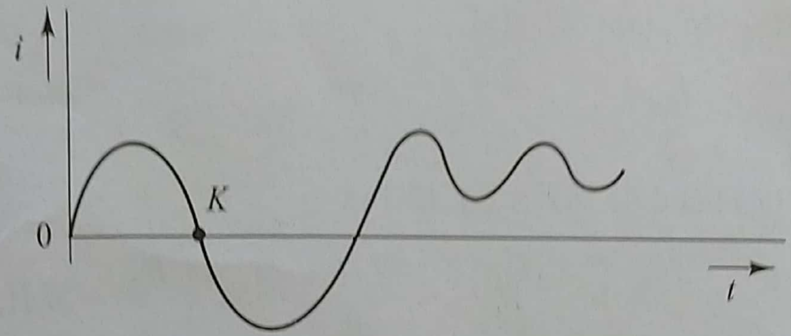


Fig. 2.13 *Voltages and currents in Class A (load is parallel with capacitor)*



(a) Waveforms of the current produced in Fig. 2.12(b) (series capacitor)



(b) Waveforms of the current produced in Fig. 2.12(a) (parallel capacitor)

Fig. 2.14

The load resistance R_L and the commutating components are so selected that their combination forms an underdamped resonant circuit. When such a circuit is excited by a d.c. source, a current of the nature shown in Fig. 2.14 will be obtained across the device. This current, as evident from its shape, has zero value at the point K where the device is automatically turned OFF. Beyond point K, the current is reversed in nature which assures definite commutation of the device. The thyristor when ON carries only the charging current of capacitor C which will soon decay to a valueless than the holding current of the device, when capacitor C is charged up to the supply voltage E_{dc}. This simultaneously switches off the thyristor. The time for switching off the device is determined by the resonant frequency which in turn depends on the values of the commutating components L and C, and the total load resistance.

This type of commutation circuits are most suitable for high frequency operation, i.e., above 1000 Hz, because of the need for an L–C resonant circuit which carries the full load current. This commutation circuit is used in series inverter.

Design Considerations

(a) Load in parallel with capacitor C Let us consider the resonant circuit of Fig. 2.12 (a). Let E_{dc} be the applied d.c. voltage, V be the load voltage, and i be the load current.

The circuit equation is

2. Class B—Self Commutation by an LC Circuit In this method, the LC resonating circuit is across the SCR and not in series with the load. The commutating circuit is shown in Fig. 2.15 and the associated waveforms are shown in Fig. 2.16.

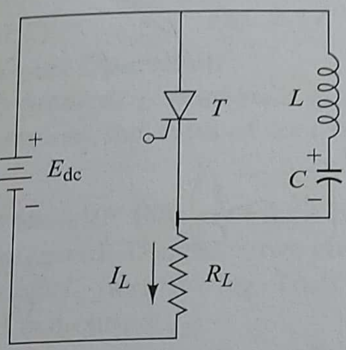


Fig. 2.15 *Class B commutation circuit*

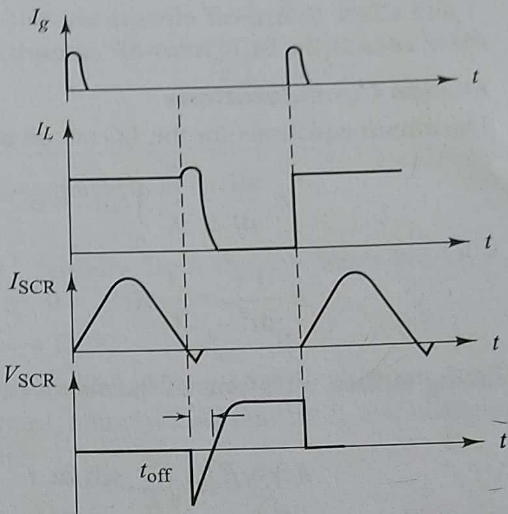


Fig. 2.16 *Associated waveforms*

Initially, as soon as the supply voltage E_{dc} is applied, the capacitor C starts getting charged with its upper plate positive and the lower plate negative, and it charges up to the voltage E_{dc}.

When thyristor T is triggered, the circuit current flows in two directions: (1) The load current I_L flows through the path $E_{dc\,+} - T - RL - E_{dc\,-}$, and (2) Commutating current I_c.

The moment thyristor T is turned ON, capacitor C starts discharging through the path $C_+ - L - T - C_-$. When the capacitor C becomes completely discharged, it starts getting charged with reverse polarity. Due to the reverse voltage, a commutating current I_C starts flowing which opposes the load current I_L. When the commutating current I_C is greater than the load current I_L, thyristor T becomes turned OFF. When the thyristor T is turned OFF, capacitor C again starts getting charged to its original polarity through L and the load. Thus, when it is fully charged, the thyristor will be ON again.

Hence, from the above discussion it becomes clear that the thyristor after getting ON for sometime automatically gets OFF and after remaining in OFF state for sometime, it again gets turned ON. This process of switching ON and OFF is a continuous process. The desired frequency of ON and OFF states can be obtained by designing the commutating components as per the requirement. The main application of this process is in d.c. chopper circuits, where the thyristor is required to be in conduction state for a specified duration and then to remain in the OFF state also for a specified duration. Morgan chopper circuit using a saturable reactor in place of the ordinary inductor L is a modified arrangement for this process. The circuit has the advantage of longer oscillation period and therefore of more assurance of commutation. In this Class B commutation method, the commutating component does not carry the load current. Both Class A and Class B turn-off circuits are self-commutating types, that is in both of these circuits the SCR turns-off automatically after it has been turned on.

Design Considerations

The circuit equations for the LC circuit are:

$$L\frac{di}{dt} + \frac{1}{C}\int i\,dt = 0 \tag{2.36}$$

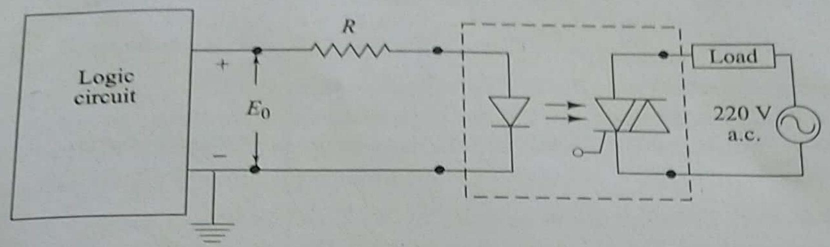


Fig. 3.8 *Electrical isolation provided by an optoisolator*

3.5 GATE TRIGGER CIRCUITS

The general block diagram of a gate-trigger circuit, for example, in a single phase converter is shown in Fig. 3.9. The thyristors are at line potential and the trigger circuit must be referenced within respect to a logic ground associated with

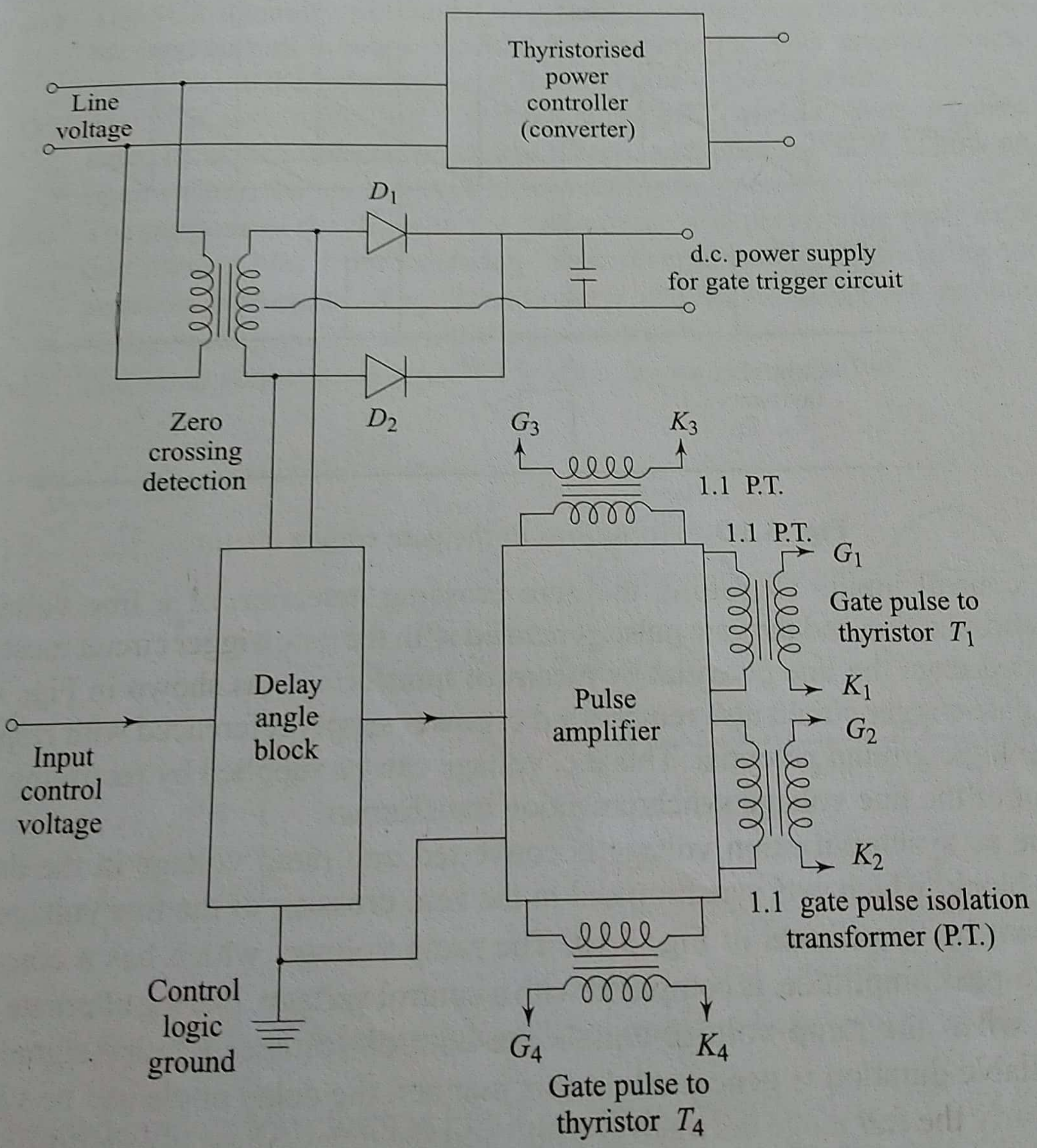


Fig. 3.9 *General block diagram of a thyristor gate trigger circuit*

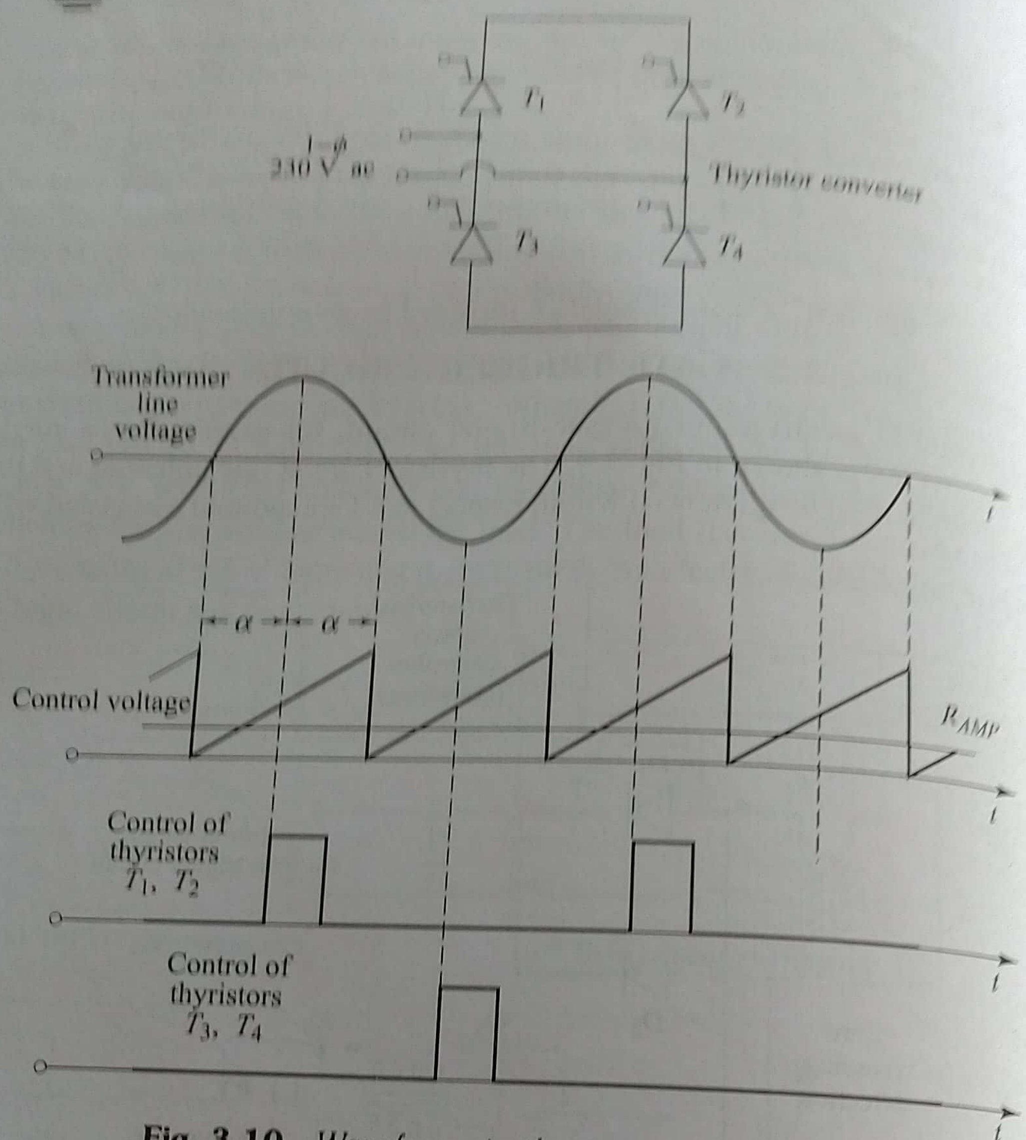


Fig. 3.10 *Waveforms in the gate trigger circuit*

the control input. Therefore, the zero crossing detection of a line voltage synchronization and the gate pulse generated with the gate trigger circuit must be isolated from the line potential by means of transformers as shown in Fig. 3.9. The gate-trigger circuit also requires a d.c. power supply referenced with respect to the logic ground potential. This d.c. voltage can be supplied by rectifying the output of the line voltage synchronization transformer.

The ac synchronization voltage is converted into ramp voltage in the delay angle block, which gets synchronised to the zero crossing of the line voltage, as is shown by waveforms in Fig. 3.10. The ramp voltage, which has a constant peak-to-peak amplitude, is compared with a control voltage. During alternate half cycles when the ramp voltage equals the control voltage, a pulse signals of controllable duration is generated. In this manner, the delay angle can be varied over nearly the full range between 0–180° and the delay-angle is proportional to the control voltage.

3.5.1 Resistance Firing Circuit

The circuit in Fig. 3.11 shows a simple method for varying the trigger angle and therefore, the power in the load. Instead of using a gate pulse to trigger the SCR, the gate current is supplied by an a.c. source of voltage e_s through R_{min}, R_v, and the series diode D. The circuit operates as follows:

(i) As e_s goes positive, the SCR becomes forward-biased from anode to cathode; however, it will not conduct ($e_L = 0$) until its gate current exceeds $I_{g(min)}$.

(ii) The positive e_s also forward biases the diode and the SCRs gate–cathode junction; this causes flow of a gate current i_g.

(iii) The gate current will increase as e_s increases towards its peak value. When i_g reaches a value equal to $I_{g(min)}$, the SCR turns "on" and e_L will approximately equal e_s (refer to point P on the waveform in Fig. 3.11).

(iv) The SCR remains "on" and $e_L \approx e_s$ until e_s decreases to the point where the load current is below the SCR holding-current. This usually occurs very close to the point until $e_s = 0$ and begins to go negative.

(v) The SCR now turns "off" and remains "off" while e_s goes negative since its anode–cathode is reverse biased, and since the SCR is now an open switch, the load voltage is zero during this period.

(vi) The purpose of the diode in the gate-circuit is to prevent the gate–cathode reverse bias from exceeding peak reverse gate voltage during the negative half-cycle of e_s. The diode is chosen to have peak reverse-voltage rating greater than the input voltage E_{max}.

(vii) The same sequence is repeated when e_s again goes positive.

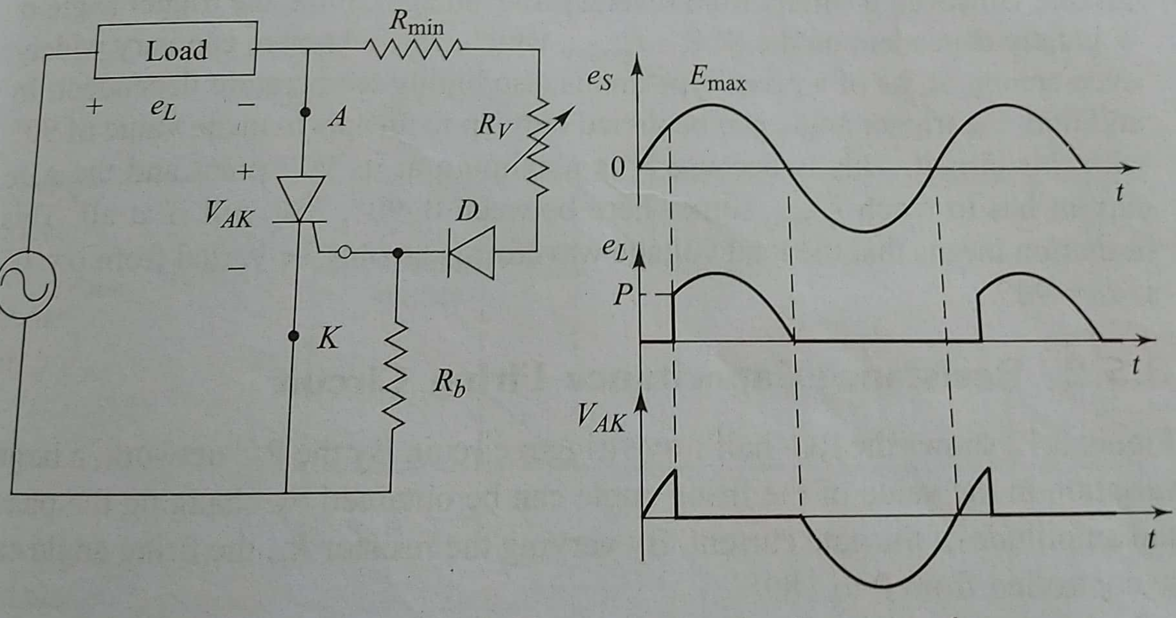


Fig. 3.11 *R-firing circuit and associated voltage waveforms*

The load-voltage waveform in Fig. 3.11 can be controlled by varying R_v which varies the resistance in the gate circuit. If R_v is increased, the gate current will reach its trigger value $I_{g(min)}$ at a greater value of e_s making the SCR to trigger at a latter point in the e_s positive half-cycle. Thus, the trigger angle α will increase. The opposite will occur if R_v is decreased. Of course, if R_v is made large enough the SCR gate current will never reach $I_{g(min)}$ and the SCR will remain "off". The minimum trigger angle is obtained with R_v equal to zero.

As shown in Fig. 3.11, the limiting resistor $R_{(min)}$ is placed between anode and gate so that the peak gate current of the thyristor I_{gm} is not exceeded. In the worst case, that is when the supply voltage has reached its peak, E_{max},

$$R_{min} \geq \frac{E_{max}}{I_{gm}} \tag{3.1}$$

The stabilising resistor R_b should have such a value that the maximum voltage drop across it does not exceed maximum possible gate voltage $V_{g(max)}$. From the voltage distribution,

$$R_b \leq \frac{(R_v + R_{min})\cdot V_{g(max)}}{(E_{max} - V_{g(max)})} \tag{3.2}$$

The thyristor will trigger when the instantaneous anode voltage, e_s , is

$$e_s = I_{g(min)}\,(R_v + R_{min}) + V_d + V_{g(min)} \tag{3.3}$$

where $I_{g(min)}$ = minimum gate current to trigger the thyristor

V_d = voltage drop across the diode

$V_{g(min)}$ = gate-voltage to trigger, corresponding to $I_{g(min)}$.

The resistance trigger shown in Fig. 3.11 is the simplest and most economical circuit. However, it suffers from several disadvantages. First, the trigger angle α is greatly dependent on the SCR's $I_{g(min)}$, which, as we known, can vary widely even among SCRs of a given type and is also highly temperature dependent. In addition, the trigger angle can be varied only up to an approximate value of 90° with this circuit. This is because e_s is maximum at its 90° point and the gate current has to reach $I_{g(min)}$ somewhere between 0–90°, if it will if at all. This limitation means that the load voltage waveform can only be varied from $\alpha = 0°$ to $\alpha = 90°$.

3.5.2 Resistance-Capacitance Firing Circuit

Figure 3.12 shows the RC–half wave trigger circuit. By the RC network, a larger variation in the value of the firing angle can be obtained by changing the phase and amplitude of the gate current. By varying the resistor R_v, the firing angle can be controlled from 0 to 180°.

In the negative half-cycle, capacitor C charges through diode D_2 with lower plate positive to the peak supply voltage E_{max}. This capacitor voltage remains constant at $-E_{max}$ until supply voltage attains zero value. Now, as the SCR anode voltage passes through zero and becomes positive, capacitor C begins to charge through R_v from the initial voltage $-E_{max}$. When the capacitor charges to positive voltage equal to gate trigger voltage V_{gt} $(= V_{g(min)} + V_{D1})$, SCR is triggered and after this, the capacitor holds to a small positive voltage, as shown in Fig. 3.12. During negative half-cycle, the diode D_1 prevents the breakdown of the gate to cathode junction. In the range of power-frequencies, the RC for zero output voltage is given by

$$R_v C \geq \frac{1.3T}{2} = \frac{4}{w} \tag{3.4}$$

where $T = 1/f$ = period of ac line frequency in seconds.

As discussed above, the thyristor will turn ON when the capacitor voltage e_c equals $(V_{g(min)} + V_{D1})$, provided the gate current $I_{g(min)}$ is available. Therefore, the maximum value of R_v is given by

$$e_s \geq I_{g(min)} R_v + e_c$$

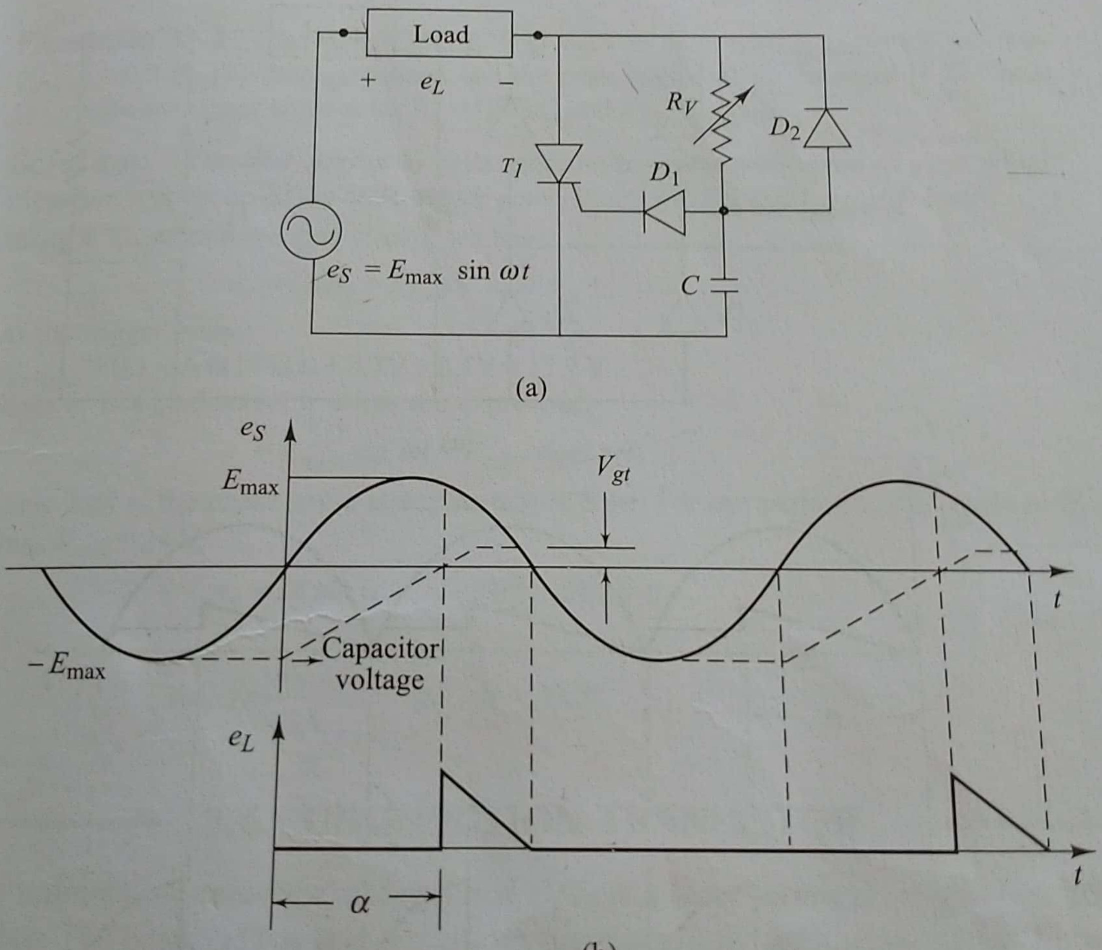


Fig. 3.12 ***(a)*** *RC firing circuit,* ***(b)*** *voltage-waveform*

$$\sin \alpha = \frac{12.2}{24} \qquad \therefore \quad \alpha = 30.6^\circ.$$

3.6 UNIJUNCTION TRANSISTOR

The unijunction transistor, abbreviated UJT, is a three-terminal, single-junction device. The basic UJT and its variations are essentially latching switches whose operation is similar to the four-layer diode, the most significant difference being that the UJT's switching voltage can be easily varied by the circuit designer. Like the four-layer diode, the UJT is always operated as a switch and finds most frequent applications in oscillators, timing circuits and SCR/TRIAC trigger circuits.

3.6.1 Basic Operation

A typical UJT structure, pictured in Fig. 3.14, consists of a lightly doped, *N*-type silicon bar provided with ohmic contacts at each end. The two end connections are called base-1, designated B_1 and base-2, B_2. A small, heavily doped *P*-region is alloyed into one side of the bar closer to B_2. This *P*-region is the UJT emitter, *E*, and forms a *P-N* junction with the bar.

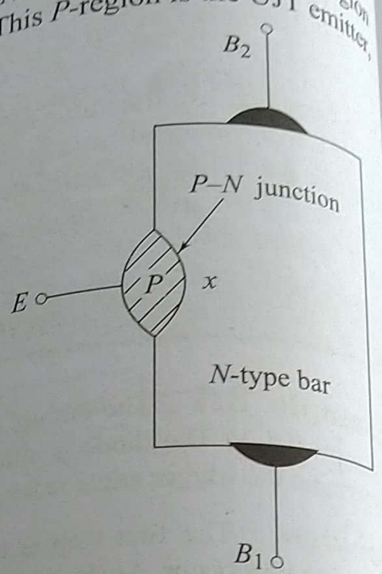


Fig. 3.14 *Basic UJT structure*

An interbase resistance, R_{BB}, exists between B_1 and B_2. It is typically between 4 kΩ and 10 kΩ, and can easily be measured with an ohmeter with the emitter open. R_{BB} is essentially the resistance of the *N*-type bar. This interbase resistance can be broken up into two resistances, the resistance from B_1 to emitter, called R_{B_1}, and resistance from B_2 to emitter, called R_{B_2}. Since the emitter is closer to B_2, the value of R_{B_1} is greater than R_{B_2} (typically 4.2 kΩ versus 2.8 kΩ).

The operation of the UJT can better be explained with the aid of an equivalent circuit. The UJT's circuit symbol and its equivalent circuit are shown in Fig. 3.15. The diode represents the *P-N* junction between the emitter and the base-bar (point *x*). The arrow through R_{B_1} indicates that it is variable since during normal operation it may typically range from 4 kΩ down to 10 Ω.

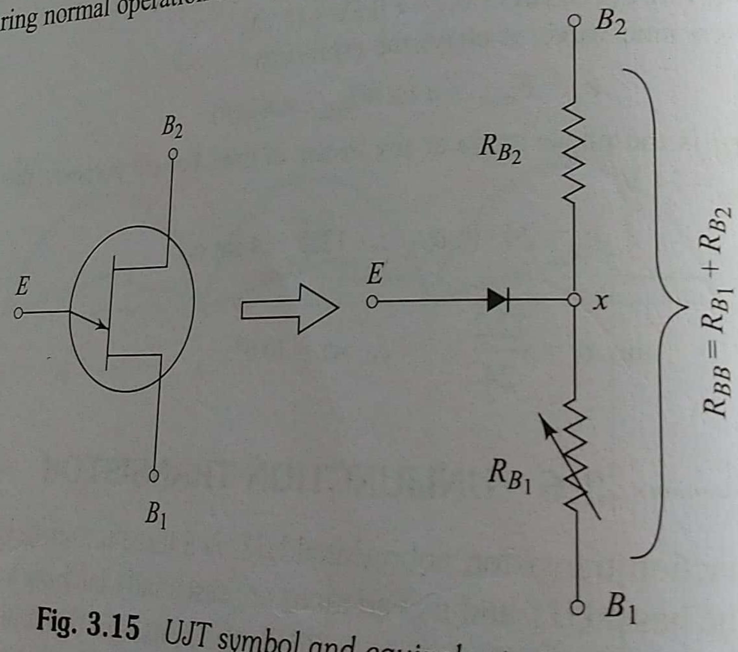


Fig. 3.15 *UJT symbol and equivalent circuit*

The essence of UJT operation can be stated as follows:

(a) When the emitter diode is reverse biased, only a very small emitter current flows. Under this condition, R_{B_1} is at its normal high-value (typically 4 kΩ). This is the UJT's "off" state.

(b) When the emitter diode becomes forward biased, R_{B1} drops to a very low value (reason to be explained later) so that the total resistance between E and B_1 becomes very low, allowing emitter current to flow readily. This is the "on" state.

Circuit-operation The UJT is normally operated with both B_2 and E biased positive relative to B_1 as shown in Fig. 3.16. B_1 is always the UJT reference terminal and all voltages are measured relative to B_1. The V_{BB} source is generally fixed and provides a constant voltage from B_2 to B_1. The V_{EE} source is generally a variable voltage and is considered the input to the circuit. Very often, V_{EE} is not a source but a voltage across a capacitor.

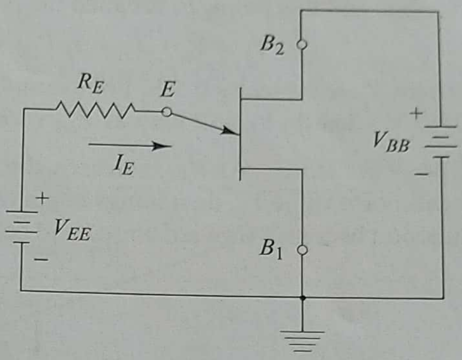


Fig. 3.16 *Normal UJT biasing*

We will analyze the UJT circuit operation with the aid of the UJT equivalent circuit, shown inside the dotted lines in Fig. 3.17(a). We will also utilize the UJT emitter-base-1 V_E-I_E curve shown in Fig. 3.17(b). The curve represents the variation of emitter current I_E, with emitter-base-1 voltage, V_E, at a constant B_2-B_1 voltage. The important points on the curve are labelled, and typical values are given in parentheses.

The "Off" state If we neglect the diode for a moment, we can see in Fig. 3.17(a) that R_{B_1} and R_{B_2} form a voltage divider that produces a voltage V_x from point x relative to ground.

$$\because \quad V_x = \frac{R_{B_1}}{R_{B_1} + R_{B_2}} \times V_{BB} = \underbrace{\frac{R_{B_1}}{R_{BB}}}_{\eta} \times V_{BB}$$

or simply,

$$V_x = \eta\, V_{BB} \tag{3.9}$$

where η (the greek letter "eta") is the internal UJT voltage divider ratio $\dfrac{R_{B_1}}{R_{BB}}$ and is called the *intrinsic stand off ratio*.

Values of η typically range from 0.5 to 0.8 but are relatively constant for a given UJT.

The voltage at point x is the voltage on the N-side of the P-N junction. The V_{EE} source is applied to the emitter which is the P-side. Thus, the emitter diode will be reverse-biased as long as V_{EE} is less than V_x. This is the "off" state and is shown on the V_E–I_E curve as being a very low current region. In the "off" state, then, we can say that the UJT has a very high resistance between E and B_1, and I_E is usually a negligible reverse leakage current. With no I_E, the drop across R_E is zero and the emitter voltage, V_E, equals the source-voltage.

The UJT "off" state, as shown on the V_E–I_E curve, actually extends to the point where the emitter voltage exceeds V_x by the diode threshold voltage, V_D, which is needed to produce forward current through the diode. The emitter voltage and this point, P, is called the *peak-point voltage*, V_p, and is given by

$$V_p = V_x + V_D = \eta V_{BB} + V_D \tag{3.10}$$

where V_D is typically 0.5 V. For example, if $\eta = 0.65$ and $V_{BB} = 20$V, then V_p = 13.5 V. Clearly, V_p will vary as V_{BB} varies.

The "On" state As V_{EE} increases, the UJT stays "off" until V_E approaches the peak-point value V_p, then things begin to happen. As V_E approaches V_P, the P-N junction becomes forward biased and begins to conduct in the opposite direction.

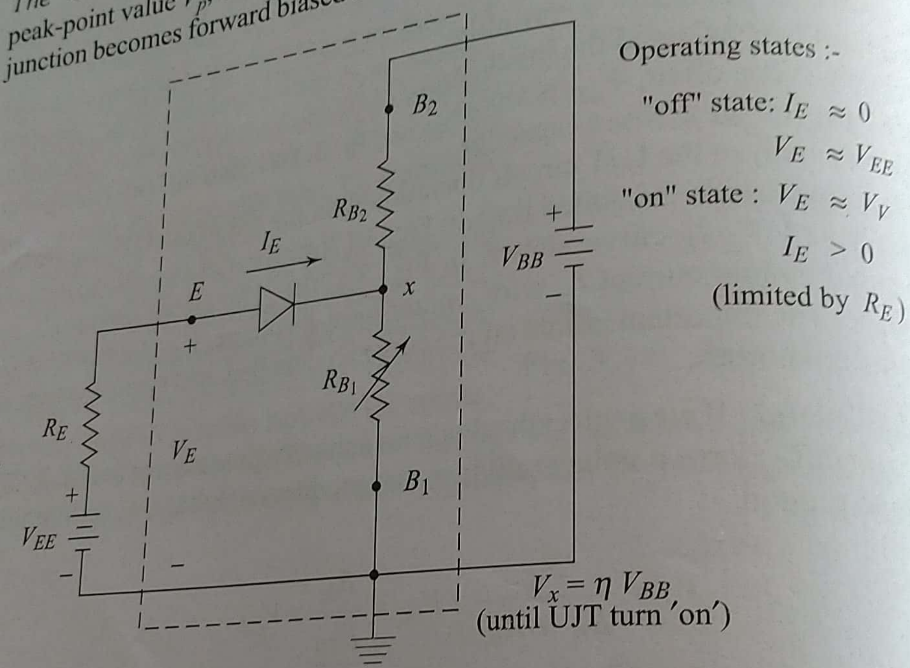


Fig. 3.17 ***(a)*** *Equivalent-circuit for UJT analysis*

Note on the V_E–I_E curve that I_E becomes positive near the peak point P. When V_E exactly equals V_p, the emitter current equals I_P, the *peak-point current*. At this point, holes from the heavily doped emitter are injected into the N-type bar, specially into the B_1 region. The bar, which is lightly doped, offers very little chance for these holes to recombine. As such, the lower half of the bar becomes replete with additional current carriers (holes) and its resistance R_{B_1} is drastically reduced. The decrease in R_{B_1} causes V_x to drop. This drop in turn causes the diode to become more forward biased, and I_E increases even further. The larger I_E injects more holes into B_1, further reducing R_{B_1}, and so on. When this *regenerative* or *snowballing* process ends, R_{B_1} has dropped to a very small value (2–25 Ω) and I_E can become very large, limited mainly by external resistance R_E.

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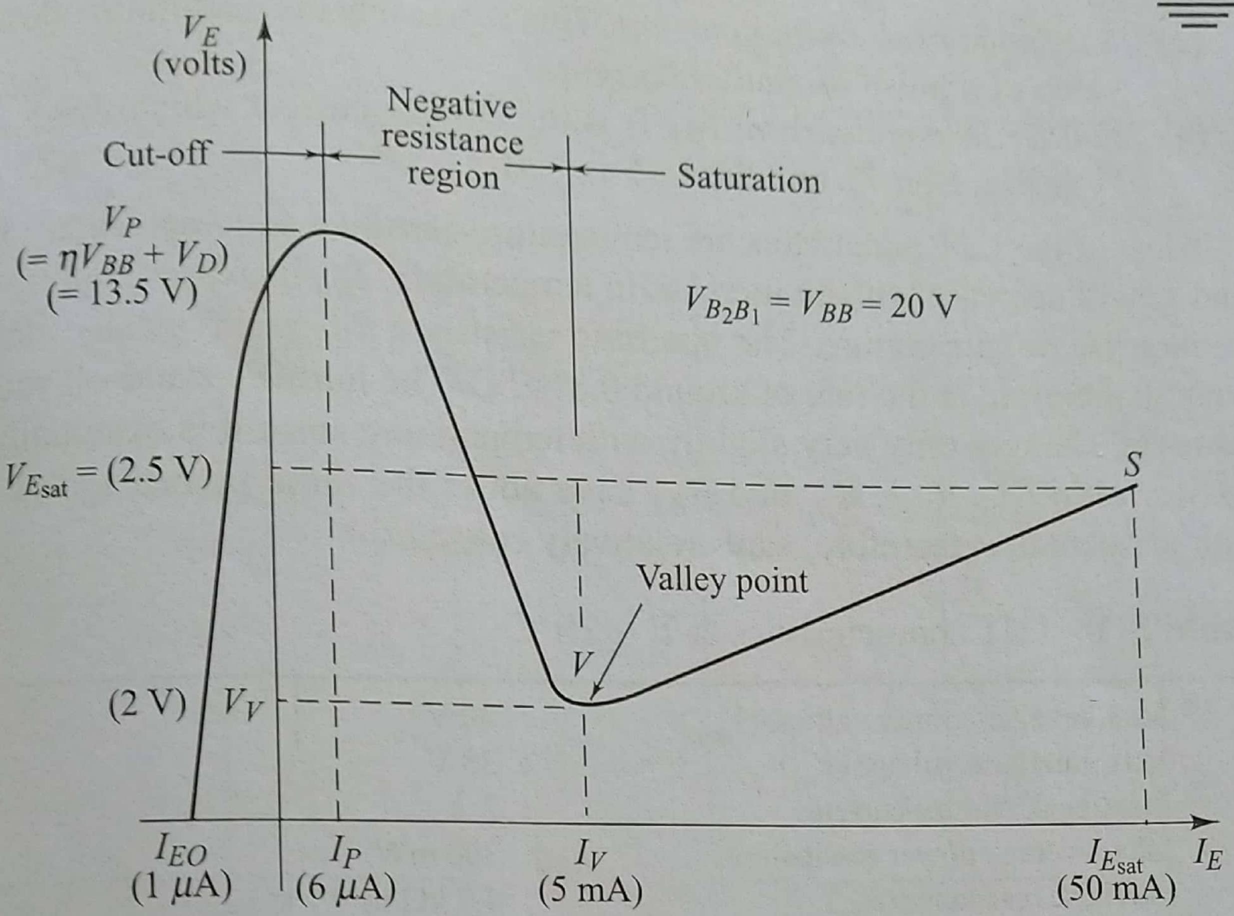

Fig. 3.17 ***(b)*** *Typical UJT V-I characteristic curve*

The UJT operation has switched to the low-voltage, high-current region of its $V_E - I_E$ curve. The slope of this "on" region is very steep, indicating a low resistance. In this region, the emitter voltage V_E, will be relatively small, typically 2 V, and remains fairly constant as I_E is increased up to its maximum rated value, $I_{E(\text{sat})}$. Thus, once the UJT is "on," increasing V_{EE} will serve to increase I_E while V_E remains around 2V.

Turning "Off" the UJT Once it is "on," the UJT's emitter current depends mainly on V_{EE} and R_E. As V_{EE} decreases, I_E will decrease along the "on" portion of the $V_E - I_E$ curve. When I_E decreases to point V, the valley point, the emitter current is equal to I_V, the *valley current*, which is essentially the holding current needed to keep the UJT "on". When I_E is decreased below I_V, the UJT turns "off" and its operation rapidly switches back to the "off" region of its $V_E - I_E$ curve, where $I_E \approx 0$ and $V_E - V_{EE}$. The valley current is the counterpart of the holding current in *PNPN* devices, and generally ranges between 1 and 10 mA.

3.6.2 UJT Parameters and Ratings

A set of parameter and ratings for a typical UJT (2N2646) are listed in Table 3.1. ... explanation are:

Max. average p...	
Interbase resistance (R_{BB})	
Intrinsic stand-off ratio (η)	
Emitter-leakage current (I_{E0})	: 12 μA (max) at $V_{B_2E} = 30$ V
Valley current (I_v)	: 4 mA (min) at $V_{BB} = 20$ V
Valley voltage (V_v)	: 2 V (typical) at $V_{BB} = 20$ V
Peak-point current (I_p)	: 5 μA (max) at $V_{BB} = 25$ V

3.6.3 UJT Relaxation Oscillator

The UJT is often used as a trigger device for SCRs and TRIACs. Other applications include nonsinusoidal oscillators, sawtooth generators, phase-control, and timing circuits.

The most common UJT circuit in use today is the relaxation oscillator shown in Fig. 3.18. Also, this type of circuit is basic to other timing and trigger circuits. The operation is as follows:

Let us consider the situation in which the capacitor is at zero volts ($V_c = 0$) and the switch is suddenly closed at $t = 0$ applying E_{dc} to the circuit. Since $V_E = V_c = 0$, the UJT emitter diode is reverse-biased and the UJT is "off." The amount of reverse bias is V_x volts which can be obtained using the voltage divider rule:

$$V_x = \frac{(R_1 + R_{B_1})E_{dc}}{R_1 + R_{B1} + R_2 + R_{B2}} \tag{3.11}$$

In many cases, R_1 and R_2 are much smaller than R_{B_1} and R_{B_2}, and V_x becomes approximately equal to ηE_{dc} (Eq. 3.9).

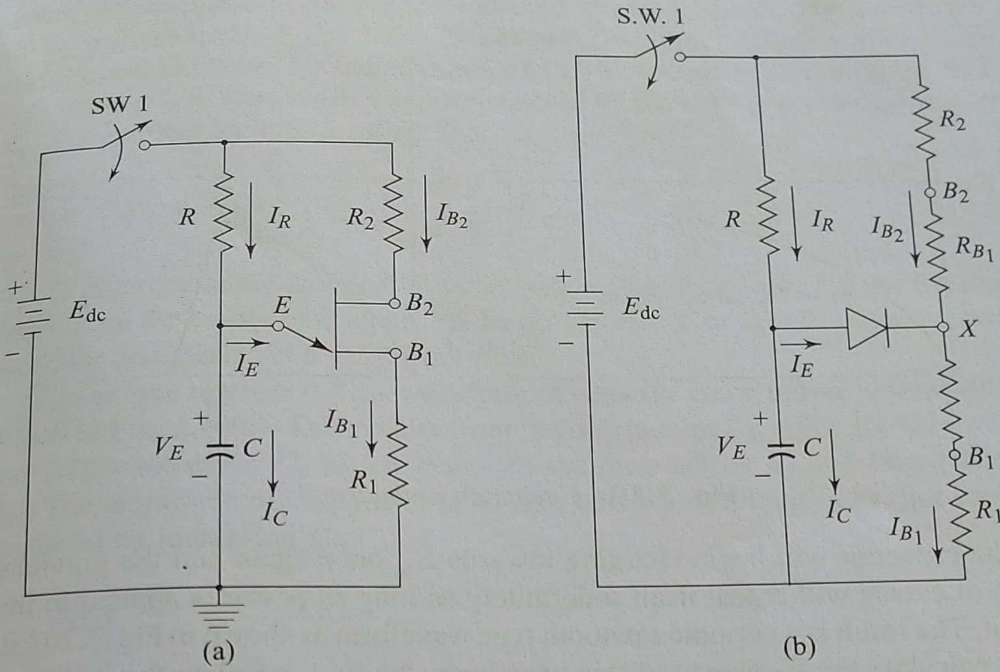

Fig. 3.18 ***(a)*** *UJT basic-relation oscillator* ***(b)*** *its equivalent circuit*

In this condition the only emitter current flowing will be small-reverse-leakage, I_{E0}. Also, R_{B_1} will be at its "off" value (typically 4 kΩ). Thus we can consider the emitter to be open ($I_E \approx 0$) and the capacitor will begin to charge toward the input voltage E_{dc} through resistor R. The capacitor voltage increases with a time constant of RC as illustrated in Fig. 3.19 (a). It will continue to increase until the voltage at the emitter reaches the peak-point value, V_{p_1} given by Eq. (3.10). At this time, the emitter diode becomes forward biased and the UJT turns 'on' with R_{B_1} dropping to a very low value (typically 10 ohms). Since the diode is now forward-biased, the capacitor will discharge through the low-resistance path containing the diode, R_{B_1} and R_1.

The capacitor discharge time constant is normally very short compared to its charging time constant (see Fig. 3.19(b)). An analytical expression for the discharge time constant is difficult to obtain since R_{B1} will continually change as the current I_E decreases. The discharging capacitor provides the emitter current needed to keep the UJT "on"; it will remain "on" until I_E drops below the valley current I_V, at which time the UJT will turn "off." This occurs at time T_2 when the capacitor voltage has dropped to the valley voltage V_v (typically 2–3 volts). At this time, R_{B1} returns to its "off" value, the diode is again reverse-biased and $I_E \approx 0$.

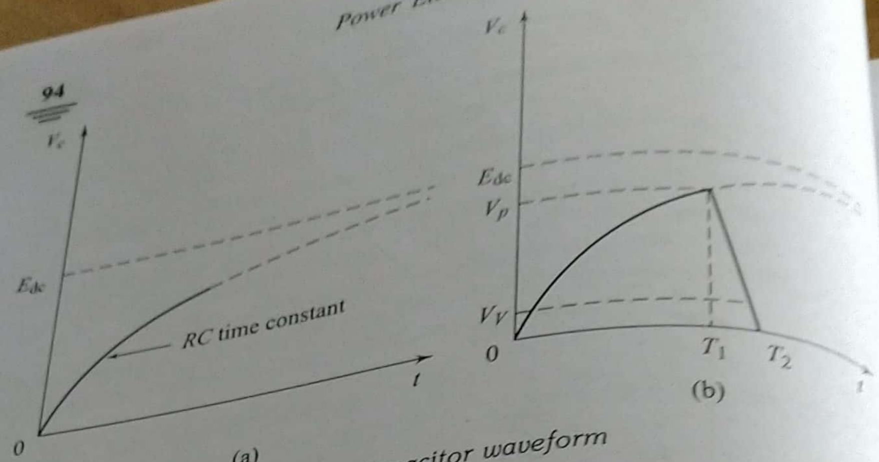

Fig. 3.19 *Capacitor waveform*

The capacitor will begin charging towards E_{dc} once again and the previous chain of events will repeat itself indefinitely as long as power is applied to the circuit. The result is a periodic sawtooth type waveform as shown in Fig. 3.20(a).

To calculate the frequency of this waveform, we first calculate the period of one cycle. The length of one period, T_1, is essentially the time it takes for the capacitor to charge to V_p since the discharge time T_2 is usually relatively short. Thus $T \approx T_1$ and is given by

$$T = R.C.\log_e\left(\frac{E_{dc}}{E_{dc} - V_p}\right) \tag{3.12}$$

In most cases, $V_p \approx \eta\, E_{dc} + V_0$ and the period can be written as

$$T \approx R.C.\log_e\left[\frac{E_{dc}}{E_{dc}(1-\eta) - V_D}\right] \tag{3.13}$$

The small diode drop V_D can often be ignored if $E_{dc} > 10$ V, resulting in the more approximate expression,

$$T \approx R.C.\log_e\left[\frac{1}{1-\eta}\right] \tag{3.14}$$

Examination of Eq. 3.14 brings out an important point, namely that T is relatively independent of supply voltage E_{dc}. This characteristic is important when designing a stable oscillator circuit. The oscillator frequency is given by $1/T$ and can be obtained by using either of the three previous equations for T.

Pulse outputs The UJT relaxation oscillator circuit can also supply pulse waveforms. If the output is taken from B_1, the result is a train of pulses occurring during the discharge of the capacitor through the UJT emitter. The waveforms of V_{B_1} is illustrated in Fig. 3.20(b). The amplitude of the B_1 pulses is always less than V_P, but is greater for larger values of C. The voltage at B_1 during the UJT "off" time will be very small and is determined by the voltage divider formed by R_1, R_{BB} and R_2 [see Fig. 3.18(b)] That is,

$$V_{B1}\ (\text{off}) = \left(\frac{R_1}{R_1 + R_{BB} + R_2}\right)E_{dc} \tag{3.15}$$

The rise time of the pulses at B_1 is very short (less than 1 μs), but the fall time depends on the values of C and R_1. A larger value of C or R_1 will cause a slower capacitor discharge and a longer fall-time.

If the output is taken at B_2, a waveform of negative going pulses is obtained as shown in Fig. 3.20(c). This results from the decrease in R_{B_1} when the UJT turns "on". This increases I_{B_2} which increases the drop across R_2 and thus reduces V_{B_2}. The amplitude of this pulses is usually about a couple of volts, but can be increased by increasing R_2.

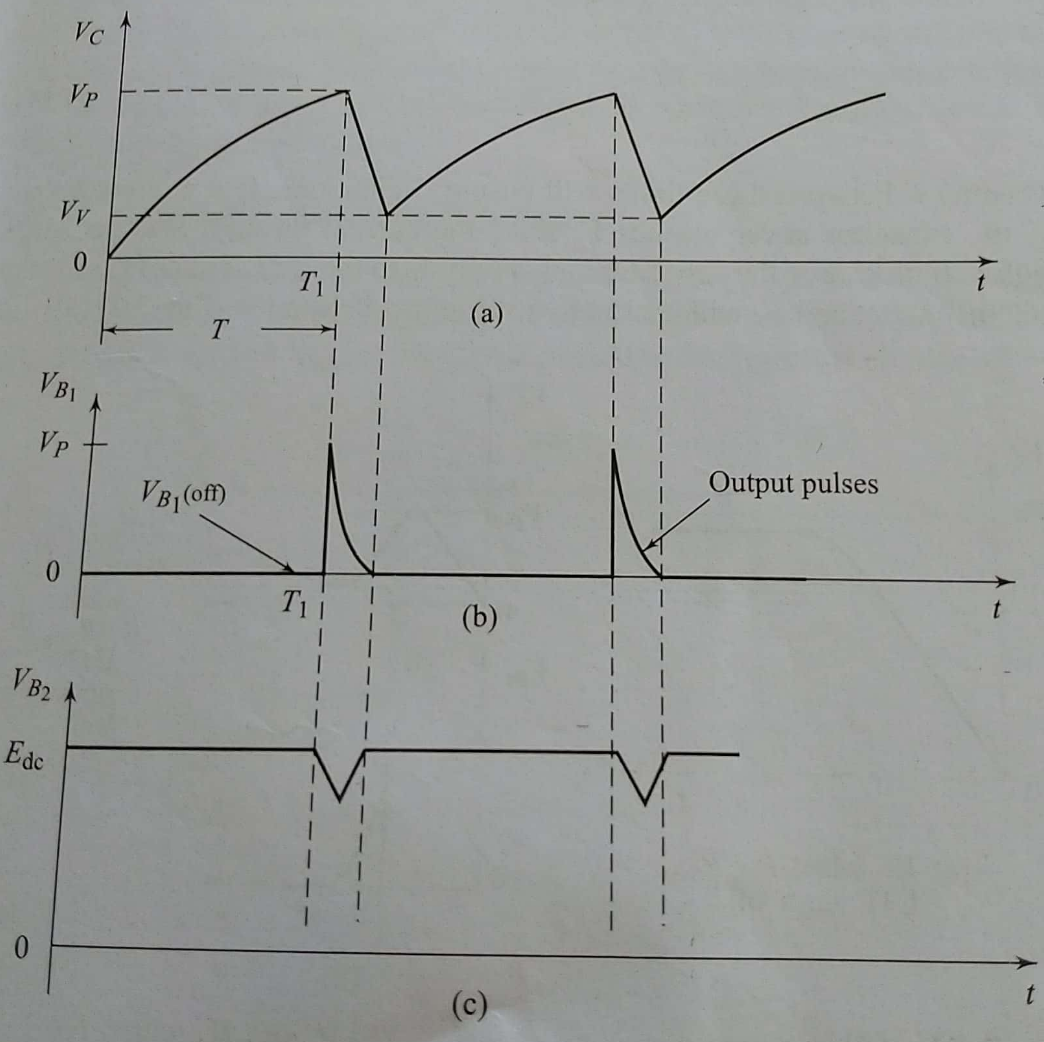

Fig. 3.20 *Waveform for UJT relation — oscillator*

The pulses at B_1 are usually the ones of most interest, they are of relatively high amplitude and are not affected by loading since they appear across a low-valued resistor R_1. These positive pulses are often used to trigger SCRs or other gated *PNPN* devices. The amplitude of these pulses is to some degree dependent on the value of C. For values of C of 1 μF or greater, the amplitude of the pulses is approximately equal to V_P (less than 2–3 V VJT drop). As C becomes smaller, the B_1 pulse decrease in amplitude. The reason for this is that the smaller value for C discharges a significant amount during the time that the UJT is making its transition from the "off" to "on" state. Thus, when the UJT finally reaches the "on" state, C has lost some of its voltage (V_p) and less voltage can appear across R_1 as the capacitor continues its discharge.

Varying the frequency The frequency of oscillations is normally controlled by varying the charging time constant RC. There are, however, limits on R. These limits are:

$$R_{\min} = \frac{E_{dc} - V_v}{I_v} \tag{3.16}$$

$$R_{\max} = \frac{E_{dc} - V_p}{I_p} \tag{3.17}$$

Keeping R between these limits will ensure oscillations. If R is greater than $R_{\max}$, the capacitor never reaches V_p since the current through R is not large enough to both charge the capacitor and supply I_P to the UJT. The UJT will stay in the "off" state, and V_c will charge to a value just below V_P. (Fig. 3.21(a))

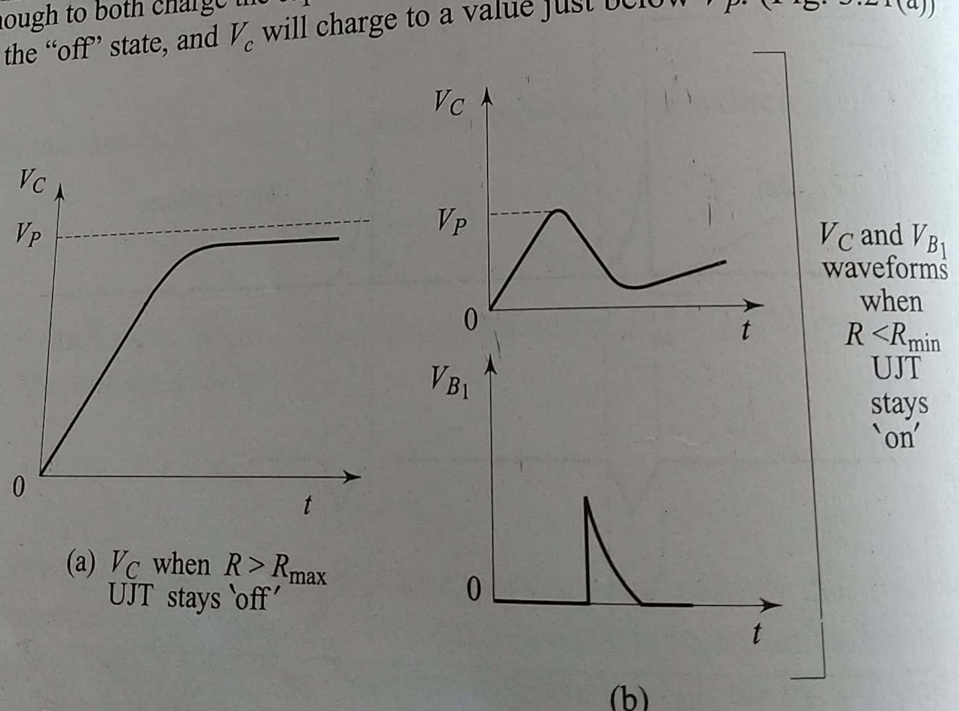

Fig. 3.21 ***(a)*** *V_c waveform when $R > R_{max}$;* ***(b)*** *V_c and R_{B_1} when $R<R_{min}$*

If R is smaller than R_{min}, the capacitor will reach V_P and discharge through the UJT, but the UJT will not turn "off" since the current through R is greater than the I_V needed to hold the UJT "on". The capacitor and V_{B_1} waveforms will consists of a single (Fig. 3.21(b)) representing one charge and discharge interval. This single pulse operation is sometimes used in time delay applications. The time delay is given by Eq. 3.12.

Examination of Eq. 3.16 indicates that to obtain a greater upper limit on frequency (a lower R_{min}) the value of I_V should be made larger. Similarly, to obtain a smaller lower limit on frequency (a higher R_{max}) the value of I_p should be made smaller. UJTs with I_v as high as 20 mA and I_p as low as 1 μA are presently available, resulting in a possible frequency range of 4000 : 1.

The frequency may also be varied by varying C. The lower limit on C is normally around 0.001 μF, while the upper limit depends on the size of R_1 (which limits on discharge current). In most applications of this circuit, the value of C is kept fixed and a variable resistor is used for R.

The temperature stability of the UJT relaxation oscillator frequency is normally very good. This is because η varies only slightly with temperature and the only variation in V_p is due to the small decrease in VD (2 mV/°C) with temperature. Its stability of frequency with variations in temperature and supply voltage coupled with its simplicity and low cost make the UJT oscillator a popular circuit for timing and pulsing applications.

3.6.4 The UJT as an SCR Trigger

The circuit examined in the previous section is often used as the gate trigger source in SCR applications. The basic circuit is shown in Fig. 3.22 when the B_1 pulse output is used to trigger the SCR a predetermined interval of time after the

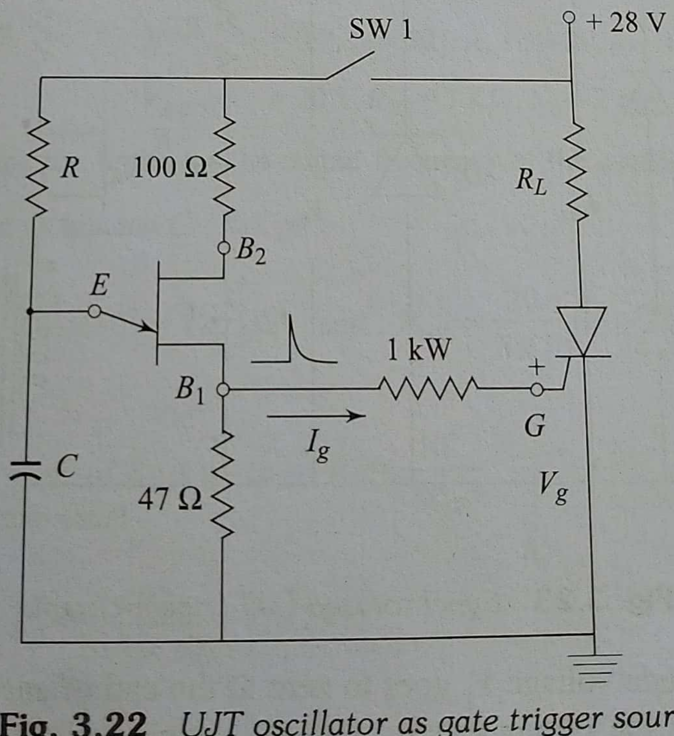

Fig. 3.22 *UJT oscillator as gate trigger source*

switch is closed. That is, the first B_1 pulse occurs T seconds after the 28 V is supplied to the UJT circuit. After the SCR has been triggered "on," subsequent pulses at its gate have no effect.

An important design consideration in this type of circuit concerns premature triggering of the SCR. The voltage at B_1 when the UJT is "off" (Eq. 3.15) must be smaller than the voltage needed to trigger the SCR, otherwise the SCR will be triggered immediately upon switch closure. Thus, we have the requirement

$$V_{B_1}(\text{off}) < (I_g \times 1\ \text{k}\Omega + V_g) \tag{3.18}$$

3.6.5 Synchronized UJT-Triggering (Ramp Triggering)

...ized UJT triggering circuit is shown in Fig. 3.23. The diode bridge ... Resistor R_s lowers E_{dc} to a suitable value for the ... the rectified-voltage to a ...