NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronic Engineering

VARIABLE – POWER CIRCUIT

Graduation Project
EE – 400

Student: Sercan Gündeş (990449)

Supervisor: Özgür Cemal Özerdem

Lefkoşa – 2003

ACKNOWLEDGEMENTS

First I want to thank Özgür Cemal Özerdem to be my advisor. Under his guidence, I succesfully study and overcome many difficulties and learn a lot about Power Electronic Circuits, Devices and Applications. In each discussion, he explained my questions patiently, and because of his endless help the progress of project passed quickly. He always helps me a lot either in my study or my life. I asked him many questions in Power Electronics and Applications and he always answered my questions quickly and in detail. And also again special thanks to him for his practical advises practical advices .

Thanks to Faculty of Engineering for having such a good computational and electronical environment which allows us to do our practical works easely, faster and accurately .

I also want to thank to my friends for their helps supports and friendships whose are in Near East University: Hakan, Hamza, Deniz, Cihan. Being with them make my 4 years in Near East University unforgatable and full of fun.

Finally, I want to thank my family, especially my wife Esra my brother Murat and my mother-in-law and father-in-law. Without their endless support and love for me, I would never achieve my current position.My mother and father is Also take my thanks because of their corporal and spiritual helps and supports..

Sercan

ABSTRACT

As the technology for the power semiconductor devices and integrated circuits develops, the potential for the applications of power electronics becomes wider. There are already many power semiconductor devices that are commercially available; however, the development in this direction is continuing.

The design of a Variable-Power Circuit requires a priori information about the characteristics and capabilities of thyristor and unijunction transistor which are very important Power Electronic Devices and we are using them as key devices for Variable-Power Circuit.

For understand the circuit well, I examined the power and output voltage waveforms of the each devices at different conditions very carefully under the guidance of Power Electronics Analysis Principles.

In this circuit we are showing how the thyristor (SCR) can be used under the control of unijunction transistor, as a variable power unit feeding a dc load. By using the unijunction transistor with the adjustable resistor we are determining the dealy and fire time of SCR.

We are trying to obtain corresponding results for power and output voltage waveforms by making appropriate adjustments on the circuit.

Sercan

INTRODUCTION

As the technology for the power semiconductor devices and integrated circuits develops, the potential for the applications of power electronics becomes wider. There are already many power semiconductor devices that are commercially available; however, the development in this direction is continuing.

The Variable-Power Circuit is one of the application of the Power Electronics which is built by thyristor, unijunction transistor and other passive devices (resistor, capacitor etc.) to reinforce our theoretical knowledege with practical experiment.

This Thesis is aimed to provide the enough information about the Power Electronics, Power Electronic Devices and Applications of Power Electronics. Analysis of the Variable-Power Circuit will help us to understand better the application of Power Electronics.

The Thesis Consists of the introduction, three chapters and conclusion.

In chapter 1 the principles and applications of Power Electronics were mentioned to give information about the topic to understand where Power Electronics is used and which type of applications does it have. Also in this chapter the types of static switches and ideal and real switches were mentioned with their characteristics, performances and relationships with power electronic devices.

In chapter 2 Power Electronic Devices were introduced on a large scale. Especially thyristors and unijunction transistor (in power transistors section) where both have quite importance for my practical application. Power Electronic Devices types, ratings, their structures, characteristics, operations, protections and their applications were briefly and clearly explained in this chapter also.

In chapter 3 Variabale-Power Circuit is discussed in details as an application of Power Electronics to explain how it is works and how we can achieve the important aspects with analyzing its output waveforms

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CHAPTER 1

POWER ELECTRONICS

1.1 APLLICATION OF POWER ELECTRONICS

The demand for control of electric power for electric motor drive systems and industrial controls existed for many years, and this led to early development of the Ward-Leonard system to obtain a variable dc voltage for the control of dc motor drives. Power electronics have revolutionized the concept of power control for power conversion and for control of electrical motor drives.

Power electronics combine power, electronics, and control. Control deals with the steady-state and dynamic characteristics of closed-loop systems. Power deals with the static and rotating power equipment for the generation, transmission, and distribution of electric power. Electronics deal with the solid-state devices and circuits for signal processing to meet the desired control objectives. *Power electronics* may be defined as the applications of solid-state electronics for the control and conversion of electric power. The interrelationship of power electronics with power, electronics, and control is shown in Fig. 1.1.

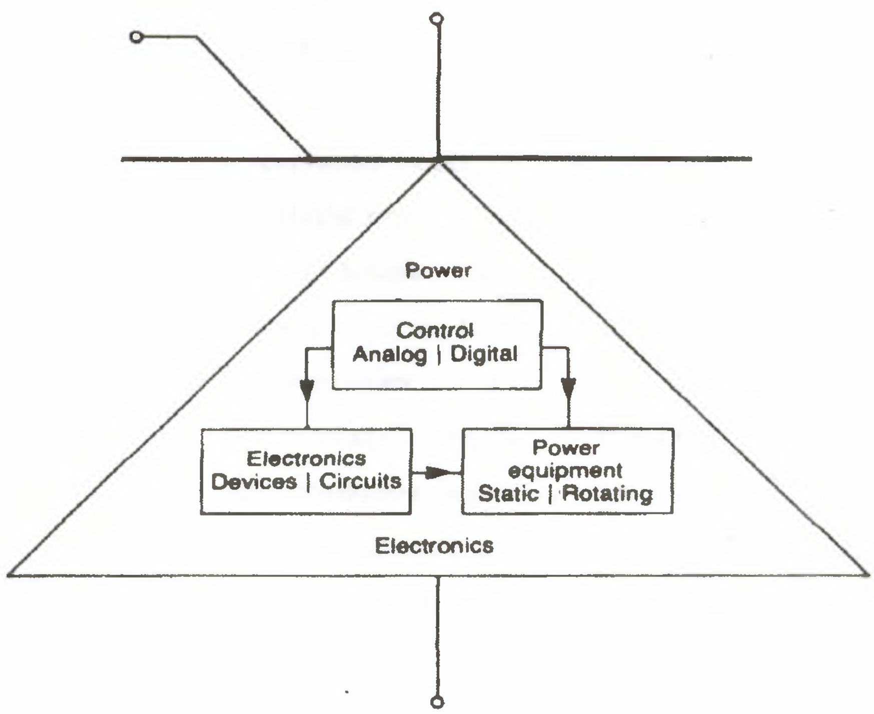


Figure 1.1 Relationship of power electronics to power, electronics and control

Power electronics is based primarily on the switching of the power semiconductor devices. With the development of power semiconductor technology, the power-handling capabilities and the switching speed of the power devices have improved tremendously. The development of microprocessors/microcomputer technology has a great impact on the control and synthesizing the control strategy for the power semiconductor devices. Modern power electronics equipment uses (1) power semiconductors that can be regarded as the muscle, and (2) microelectronics that has the power and intelligence of a brain.

Power electronics have already found an important place in modern technology and are now used in a great variety of high-power products, including heat controls, light controls, motor controls, power supplies, vehicle propulsion systems, and high-voltage direct-current (HVDC) systems. It is difficult to draw the boundaries for the applications of power electronics; especially with the present trends in the development of power devices and microprocessors, the upper limit is undefined. Table 1-1 shows some applications of power electronics.

1-2 HISTORY OF POWER ELECTRONICS

The history of power electronics began with the introduction of the mercury arc rectifier in 1900. Then the metal tank rectifier, grid-controlled vacuum-tube rectifier, ignitron, phanotron, and thyratron were introduced gradually. These devices were applied for power control until the 1950s.

The first electronics revolution began in 1948 with the invention of the silicon transistor at Bell Telephone Laboratories by Bardeen, Brattain, and Schockley. Most of today's advanced electronic technologies are traceable to that invention. Modern microelectronics evolved over the years from silicon semiconductors. The next breakthrough, in 1956, was also from Bell Laboratories: the invention of the *PNPN* triggering transistor, which was defined as a thyristor or silicon-controlled rectifier (SCR).

The second electronics revolution began in 1958 with the development of the commercial thyristor by the General Electric Company. That was the beginning of a new era of power electronics. Since then, many different types of power semiconductor devices and conversion techniques have been introduced. The microelectronics revolution gave us the ability to process a huge amount of information at incredible speed. The power electronics revolution is giving us the ability to shape and control large amounts of power with ever-increasing efficiency. Due to the marriage of power electronics, the muscle, with microelectronics, the brain, many potential applications of power electronics are now emerging, and this trend will continue.

Table 1.1 Some applications of power electronics

Advertising	Air conditioning	Aircraft power supplies
Alarms	Appliances	Audio amplifiers
Battery charger	Blenders	Blowers
Boilers	Burglar alarms	
Cement kiln	Chemical processing	Clothes dryers
Computers	Conveyers	Cranes and hoists
Dimmers	Displays	
Electric blankets	Electric door openers	Electric dryers
Electric fan	Electric vehicles	Electromagnets
Electromechanical electroplating	Electronic ignition	Electrostatic precipitators
Fans	Flashers	Food mixers
Food warmer trays	Forklift trucks	Elevators
Games	Garage door openers	Gas turbine starting
Generator exciters	Grinders	
Hand power tools	Heat controls	High-frequency lighting
High-voltage dc (HVDC)		
Laser power supplies	Latching relays	Light dimmers
Linear induction motor controls	Light flashers	Locomotives
Machine tools	Magnetic recordings	Magnets
Mass transits	Mercury-arc lamp ballasts	Mining
Model trains	Motor controls	Motor drives
Paper mills	Particle accelerators	People movers
Phonographs	Photocopies	Photographic supplies
Power supplies	Printing press	Pumps and compressors
Radar/sonar power supplies	Range surface unit	Refrigerators
Regulators	RF amplifiers	
Security systems	Servo systems	Sewing machines
Solar power supplies	Solid-state contactors	Solid-state relays
Space power supplies	Static circuit breakers	Static relays
Steel mills	Synchronous machine	
Starting	Synthetic fibers	
Television circuits	Temperature controls	Timers
Toys	Traffic signal controls	Trains
Ultrasonic generators	UPS	
Vacuum cleaners	VAR compensation	Vending machines
VLF transmitters	Voltage regulators	
Washing machines	Welding	

Within the next 30 years, power electronics will shape and condition the electricity somewhere in the transmission line between its generation and all its users. The power electronics revolution has gained momentum since the late 1980s and early 1990s.

1-3 POWER SEMICONDUCTOR DEVICES AS STATIC SWICHES

In a static power converter, the power semiconductor devices function as switches, which operate statically, that is, without moving contacts. The time durations, as well as the turn ON and turn OFF operations of these switches, are controlled in such a way that an electrical power source at the input terminals of the converter appears in a different form at its output terminals. In most types of converters, the individual switches in the converter are operated in a particular sequence in one time period, and this sequence is repeated at the switching frequency of the converter. Figure 1.2 shows two simple conversion schemes to illustrate this statement.

Figure 1.2(a) shows a scheme for converting DC to AC. This type of power conversion is called inversion, and the circuit itself is called an inverter. Our inverter power circuit consists of four switches labeled S_1, S_2, S_3 and S_4 connected in the manner shown in Fig. 1.2(a). The input is a DC voltage source of magnitude V (in V) connected to the input terminals of the inverter, which are P (positive) and Q (negative). The timing of the switches is shown in Fig. 1.2(d). For example, from instant $t = 0$ to instant $t = t_1$, switches S_1 and S_4 are kept ON, the other two being kept OFF. Therefore the input DC voltage appears at the output terminals with terminal A positive. During the next interval from $t_1$to t_2, S_1 and S_4 are kept OFF, but S_2 and S_3 are kept ON. Therefore, during this interval, the input DC voltage appears at the output terminals with reversed polarity (A negative). This sequence of switching is repeated, and in this way the input voltage V *of* fixed polarity shown in Fig. 1.2(b) is presented at the output terminals PQ as an AC square wave voltage as shown in Fig. 1.2(c).

Figure 1.2(e) shows how the same circuit configuration of four switches can be used to convert an AC voltage source connected at the terminals AB into a DC voltage at the terminals PQ. This type of power conversion is called rectification, and the circuit itself a rectifier. In our rectifier of Fig. 1.2(e), the switches S_1, S_2, S_3 and S_4 are operated according to the same timing as indicated by Fig. 1.2(d). The input terminals of the rectifier are now A and B, to which the AC square wave voltage such as that shown in Fig. 1.2(c) is connected. The output terminals of our rectifier are P (positive) and Q (negative). The input AC voltage is now presented as a unidirectional voltage at the output terminals. Notice that the directions of current through the switches will be different in Figs 1.2(a) and (e).

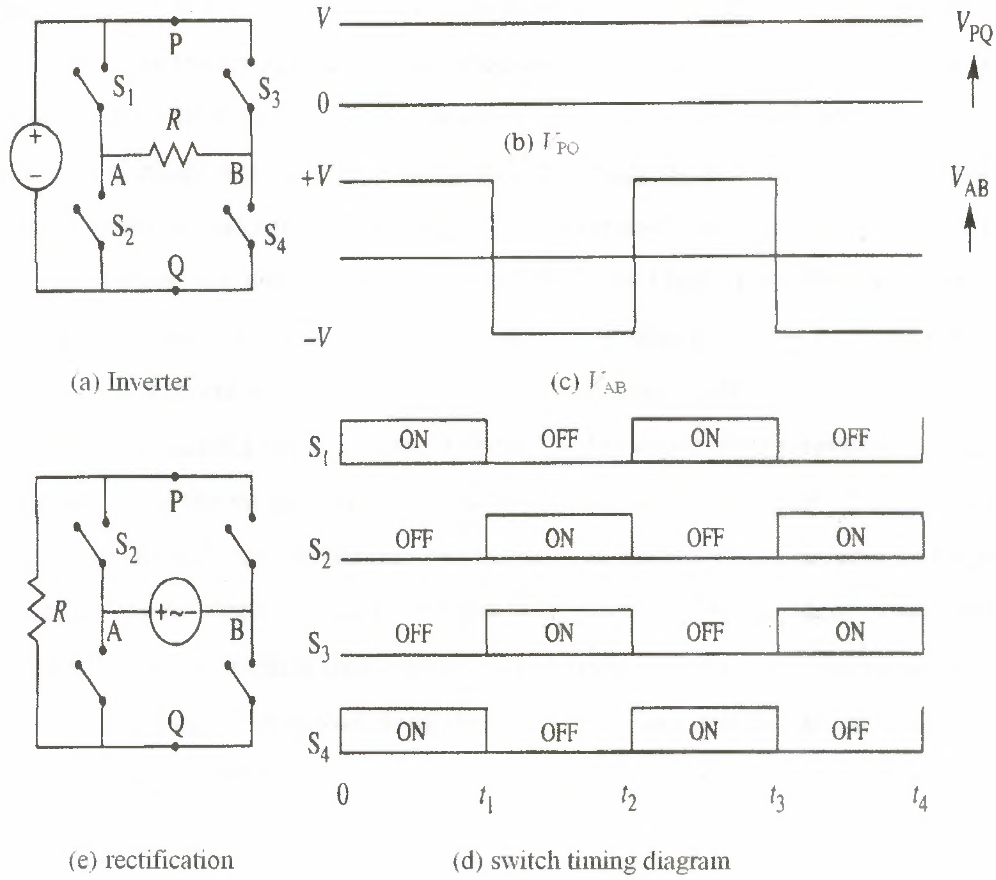


Figure 1.2 Power conversion by switching

The above examples are only intended to highlight the role of static switches in the power conversion process. The objective of this book is to present the principles underlying power conversion by the use of static switches and the techniques employed for controlling output parameters such as voltage, current, power, frequency and waveform. We shall, in a progressive sequence, present all the important types of power converters that have proved useful in the application areas of electric power. We shall also present important application areas, and this will bring out how converter schemes and control strategies can be tailored to meet specific needs. We begin this by describing the static semiconductor power switches themselves in the present chapter.

1.4 TYPES OF STATIC SWITCHES

1.4.1 UNCONTROLLED STATIC SWITCH -- THE POWER DIODE

The simplest static switch is the diode. A power diode is a two-terminal device whose circuit symbol is shown in Fig. 1.3(a). If a diode is present in an electrical circuit in such a way that its anode (terminal A) has a positive potential with respect to its cathode (terminal K), it is said to be forward-biased. An ideal diode conducts when forward-biased, with negligible voltage drop across it and a forward current, shown as I_F in Fig. 1.3(a), flows through it. If, however, it is reverse-biased, ideally, it does not conduct. A real diode will have a small forward voltage drop across it when it conducts and a small reverse leakage current when it is reverse-biased. For the purposes of our present description, we shall treat the power diode as ideal. Therefore an ideal diode, considered as a static switch, turns ON automatically whenever the external circuit can send a forward current through it. It turns OFF automatically whenever the external circuit attempts to send a reverse current though it by impressing a reverse voltage. The power terminals of the switch are A and K. It has no control terminal through which we can control its ON and OFF switching operations. The switch block reverse voltages, but has no capability to block forward voltages. We can describe the ideal power diode as an uncontrolled static switch that turns ON and turns OFF by itself, depending on the polarity of the voltage. For a switch of this kind, we shall use the diode circuit symbol itself, shown in Fig. 1.3(a).

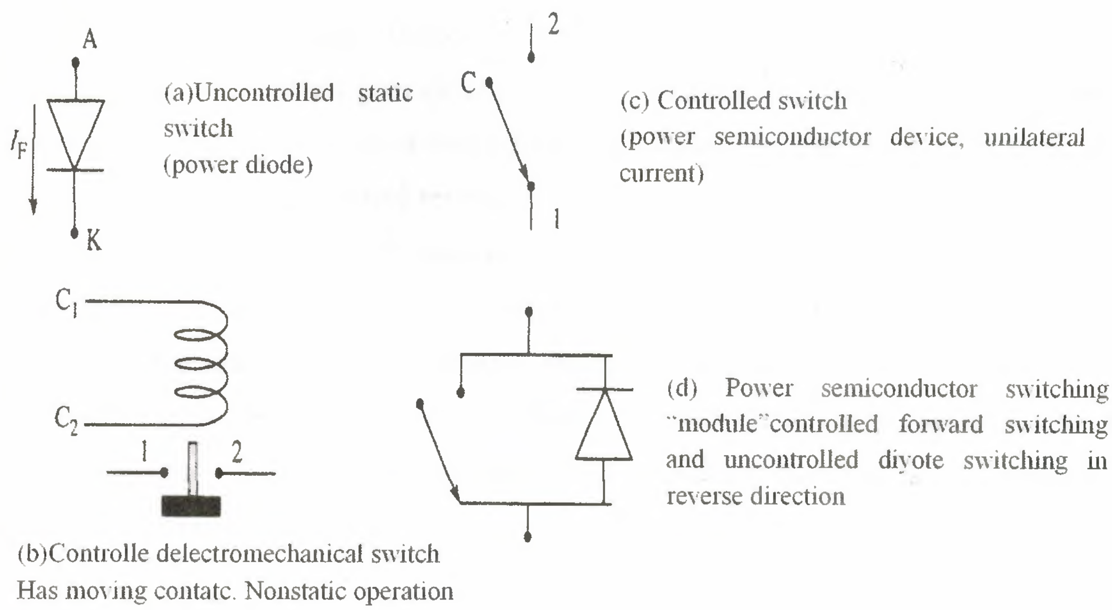


Figure1.3 Types of switches

1.4.2 CONTROLLED SWITCH

A controlled switch is one that could be turned ON and OFF by activating and deactivating a control circuit. Figure 1.3(b) shows a nonstatic (with moving contacts) switch of this kind. It has a control coil whose terminals are labeled C_1 and C_2. The power terminals of the switch are labeled 1 and 2.

To turn ON this switch, we send a current through the control coil, which will cause the plunger to move and connect the power terminals. The following aspects should be noted here.

1. The switch has four terminals—two for the power circuit and two for the control circuit.

2. The control circuit is electrically isolated from the power circuit.

The electrical isolation between the power terminals and the control circuit is very often a requirement in static power converters. The control circuit block in a static converter provides the electrical signals to perform the switching operations of the static switches. The control circuit often consists of low voltage electronic components such as analog and digital integrated circuits, working from a low voltage power supply with respect to ground. The control circuit will be damaged if large voltages with respect to ground are impressed on it. In a static power converter, the power terminals of a static switch may reach high voltages with respect to ground. Also, during the switching, these potentials change by large magnitudes in each switching cycle. It is important to ensure that these large voltages, as well as the fast changes that occur in them, do not disturb the control circuit. Unfortunately, power semiconductor switching devices presently available do not provide any isolation at all between the power and control terminals. In fact, in the case of every power semiconductor switch we shall present in this chapter, one of the control terminals is common with one of the power terminals. All are therefore three-terminal devices. A typical example is the power transistor. The power terminals are the collector and the emitter. The third terminal, namely the base, and the emitter are the control terminals. We turn ON the switch by sending the control current between the base and emitter. When electrical isolation between power and control circuits is a requirement, we have to use an external isolating device, typically a pulse transformer, to couple the control circuit to the control terminals of the switching device.

The circuit symbol that we shall use to represent an ideal three-terminal controlled unidirectional static switch is shown in Fig. 1.3(c). The power terminals of the switch are labeled 1 and 2. The arrow shows the direction of ON state current through it. The control input is across terminals C and 1.

1.4.3 DIRECTIONAL PROPERTIES OF STATIC SWITCHES -- CURRENT DIRECTION

We have described the power diode as a static uncontrolled switch with only one direction for current flow. The general circuit symbol that we have chosen and shown in Fig. 1.3(c) is for a controlled switch, which is also unidirectional. Of the power semiconductor switches described in this chapter, the unidirectional ones, besides the diode, are (1) the bipolar power transistor, (2) the insulated gate bipolar transistor (IGBT), (3) the thyristor, also known as the silicon controlled rectifier (SCR), (4) the asymmetrical silicon controlled rectifier (ASCR), (5) the gate turn off thyristor (GTO) and (6) the MOS controlled Thyristor (MCT). The switches with bidirectional current capability are (1) the power MOSFET, (2) the reverse conducting thyristor and (3) the triac. Of these, the power MOSFET and the reverse conducting thyristor function as controlled switches in the forward direction and as uncontrolled switches in the reverse direction. The triac works as a controlled switch in both directions.

In several types of static power converter circuits that we shall study in subsequent chapters, the static switches have to be bidirectional, with control only for one direction. To represent such a switch, or a combination of two switches that will achieve the same function, we shall combine our circuit symbols of the uncontrolled switch and the controlled switch in the manner shown in Fig. 1.3(d). We can use a power MOSFET or a reverse conducting thyristor if such a function is to be achieved using a single device. Alternatively, we can use a switching "block" or "module" consisting of one of the unidirectional controlled switches and a power diode connected in "anti-parallel" with it. Such switching blocks or "modules", consisting of a unilateral power semiconductor device (which may typically be a bipolar transistor or an IGBT), and an "antiparallel" diode, are currently readily available from manufacturers of power semiconductor devices.

1.4.4 DIRECTIONAL VOLTAGE CAPABILITIES OF STATIC SWITCHES

A distinction must be drawn between directional current flow capability and directional voltage blocking capability for a static power semiconductor switch. For example, the bipolar power transistor is a static switch that can only switch current in the "forward" direction. This does not mean that it has ability to block reverse voltages. In fact, it has no significant capability to withstand reverse voltages, and will be permanently damaged if it is subjected to an appreciable magnitude of reverse voltage. It can only block forward voltages, and should be used in such a way that no significant reverse voltage ever comes across it. In contrast, the thyristor, which is also a unidirectional switch, has a symmetrical voltage blocking capability in that it can block approximately the same voltages in both forward and reverse directions. The asymmetrical SCR, which is otherwise similar to the thyristor, has only a very low reverse voltage blocking capability. Certain GTO types have symmetrical voltage blocking capability, whereas others do not have this feature and can only block forward voltages. In DC-to-DC converters and in DC/AC inverters generally, reverse voltage blocking capability is seldom required. In such converters, the controlled switching element is typically used with an anti-parallel diode. Therefore the maximum reverse voltage such a device will be called upon to withstand will be the forward voltage drop of its antiparallel diode when this diode conducts current. The directional voltage and current capabilities of commonly used semiconductor power switches are summarized in Appendix A.

1.4.5 TYPES OF SWITCHING CONTROL - CONTINUOUS OR LATCHING

There is a major difference between static switches in the manner in which the control terminal performs the switching operation. In some devices, such as the bipolar power transistor and the power MOSFET, after the turn ON switching is implemented by an input to the control terminal, this input should continue to be present, to keep the switch in the ON state. If the control input ceases, the switch will turn OFF. With such a switch, both the turn ON and the turn OFF operations can be implemented by the same control circuit. Turn ON is implemented by giving a voltage or current input to the control terminal. Turn OFF is achieved by terminating this input. This type of control may be described as "continuous". In devices like the thyristor, the control input to implement turn ON need be only a pulse of very short duration. Once the switch has turned ON, there is no further need for the turn ON control pulse to be present. Another example of this type of control is the gate turn OFF thyristor (GTO). The GTO is turned ON by a short positive pulse and turned OFF by a short negative pulse on its control terminal. This type of control may be described as "latching", because the device is latched into the required state by a pulse of short duration.

The thyristor, which is a latching device, has a serious limitation. Its control terminal (gate) has the ability only to control the ON switching operation. Once the device has been latched into the ON state, the gate loses control and the device behaves like a diode. Its OFF switching has to take place by reverse bias of the main terminals like a diode..

1.5 IDEAL AND REAL SWITCHES

To assess the performance of a switch, we look at two aspects of its behavior—static and dynamic. If the switch is either in its ON or OFF state, we call this a static condition. The dynamic condition is the transition from one static state to the other. We shall look at the limitations of power semiconductor switches from these two aspects.

1.5.1 STATIC PERFORMANCE

An ideal switch should have zero voltage across it in the ON state and zero current through it in the OFF state when it is blocking a voltage. The product of current and voltage, which gives the power dissipated in the switch, is zero in both conditions. This is the basic reason why a power conversion scheme based on switching is more efficient than other methods, because ideally, there is no internal power loss. Power semiconductor switches depart to some extent from the ideal—there is a small but finite voltage drop in the ON state and a small but finite "leakage current" in the OFF state. The power dissipated in a switch during its ON state is given (in W) by

$$P = v_f i_f$$

For an ON period duration t_1 to t_2 during which v_f and i_f may vary, the total energy (in J) dissipated in it will be

$$J = \int_{t1}^{t2} v_f \, i_f \, dt$$

Energy dissipated in the switch causes its temperature to rise. For satisfactory operation, the maximum temperature of the silicon pellet that constitutes the switching element has to be limited below the specified safe limit. The efficiency of power conversion is also lowered, because of the power dissipation in the switches.

The relationship between i_f and v_f for a semiconductor switching element is typically nonlinear. This relationship, when graphically plotted, is called the output characteristic or forward characteristic of the device. This characteristic, or relevant parameters from it, are generally available from the manufacturer's data sheet for the device.

The leakage current that flows in the OFF state may be a "forward" leakage or a "reverse" leakage, depending on whether the switch is blocking a forward or a reverse voltage. In both cases, there will be power dissipation in the device. However, in present-day devices, the leakage current is quite small and does not vary significantly with voltage. Under normal load current conditions, the power dissipation due to this leakage current is small in comparison with the power dissipation in the ON state. We shall therefore generally neglect the power dissipation due to OFF state leakage and the resulting loss of energy.

1.5.2 DYNAMIC PERFORMANCE

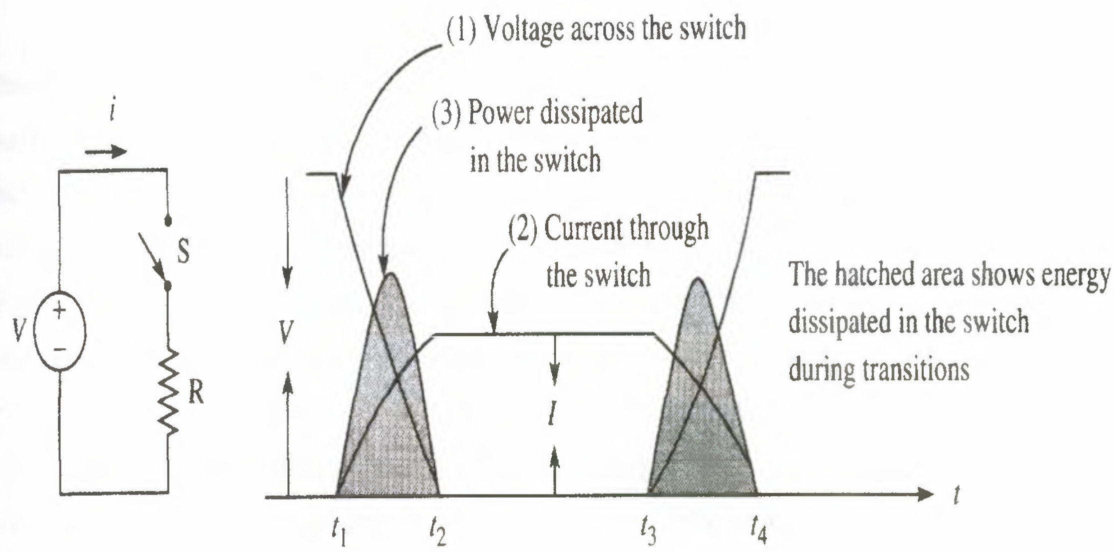
An ideal switch should change from the OFF state to the ON state, *instantaneously,* when the required switching control signal is applied to its control terminal. Similarly, the transition time for the turn OFF switching should also be zero. A real switch needs a finite t_{on} for ON switching and t_{off} for OFF switching. These finite switching times have two major consequences.

1. They limit the highest repetitive switching frequencies possible.
2. They introduce additional power dissipation in the switches themselves.

Of these, the second phenomenon needs further explanation.

Referring to Fig. 1.4(a), we assume that the switch shown there has ideal static characteristics, but nonideal dynamic performance. Figure 1.4(b) shows the waveforms of (1) the voltage drop across the switch and (2) the current through it, when a turn ON switching is implemented at $t = t_1$ and a turn OFF switching at $t = t_3$. Let us first look at the turn ON switching.

Prior to $t=t_1$ the switch is in the OFF state, and the blocking forward voltage across it is equal to the source voltage V. At $t=t_1$, the turn ON control signal arrives at its control terminal. During the turn ON transition, which takes a finite time, the instantaneous value of the voltage across the switch is assumed to be shown by the waveform in Fig. 1.4(b). During the same time, the current through the switch rises from zero to the static ON state magnitude I. The instantaneous value during the transition is assumed to be as shown by the current waveform.



(a) Circuit (b) Waveforms

Figure 1.4 Instantaneous voltages, currents and power in static switch during the switching transition

We notice that during the transition there is power dissipation taking place inside the switch, whose instantaneous value is given by the product of voltage and current. The instantaneous power is also plotted (3) as a function of time in Fig. 1.4(b). Depending on the nature of the current and voltage waveforms during the transition, the peak power can reach a relatively large magnitude. The energy dissipation in the switch is equal to the area shown hatched under the power waveform.

Similar considerations are valid for the turn OFF switching operation, which takes place from t_3 to t_4 in Fig. 1.4(b). During this transition, the voltage rises from zero to V and the current falls from / to zero. The transition times t_{on} and t_{off} are generally not equal in power semiconductor switches, t_{off} usually being larger. The total energy J (in J) dissipated in the switch in one switching cycle is given by the sum of the areas under the power waveform during t_{on} and t_{off}. The average switching power loss is therefore proportional to the switching frequency, and is given (in W) by

$$P_{switching} = f\,J$$

f being the converter switching frequency (in Hz).

1.5.3 TEMPERATURE RISE - USE OF HEAT SINKS

The immediate consequence of energy dissipation in a static switch is its temperature rise. The need for cooling arises to keep the switch temperature below its safe limit. The exact location where the heat is generated is in the silicon pellet that actually is the switch. A power semiconductor device is basically a very thin wafer of silicon in which the necessary impurity profiles are created during manufacture. Electrical contacts are made to appropriate areas of this thin wafer, and these constitute the terminals of the device, which are brought outside the casing in which the silicon pellet is housed. The power loss of the switch raises the temperature of the pellet. The temperature gradient thus created causes the heat power to flow outwards to the casing surface. To facilitate the easier flow of heat power to the ambient atmosphere outside, it is a common practice to mount the casing on a "heat sink." Heat sinks are made of metal, and provide a large surface area from which the heat power can pass by convection and radiation to the ambient atmosphere. The convection flow can be further enhanced, if needed, by using a fan to provide forced air cooling.

With the device mounted on a heat sink, the path of heat flow can be viewed as a series combination of the following individual paths: (1) from the junction (J) to the surface of the casing (C); (2) from the surface of the casing (C) to the outer surface of the heat sink (S); (3) from S to the ambient atmosphere (A), which we shall assume to be an external region sufficiently distant from the heat sink, at which the thermal gradient is negligible.

With constant power dissipation inside the pellet, thermal equilibrium conditions will be attained after a period of time. After this has happened, the temperatures and temperature gradients stay constant in such a way that all the power dissipated inside the device continually flows out into the ambient. Taking thermal power flow as proportional to temperature difference, we can represent this condition by an electrical analog in which electric current (in A) represents thermal power (in W) and potential difference (in V) represents temperature difference (in °C). Such an analogous electric circuit will consist of resistances (Ω = V/A) as the analog of "thermal resistance" (expressed as °C/W). On this basis, Fig. 1.8 shows the thermal power flow model for a static switch mounted on a heat sink. In this model, J represents the inside of the silicon pellet. Θ_{JC} is the thermal resistance (in °C/W) between J and the outer surface of the casing C. The other thermal resistances are labeled by appropriate subscripts. The model of Fig. 1.5 can be used to make estimates of the junction temperature rise when the power dissipation is known, or to estimate the maximum power dissipation possible for a specified junction temperature.

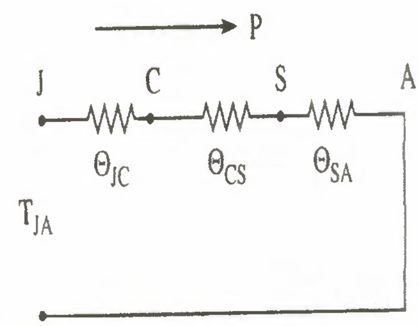


Figure 1.5 Model of heat flow under thermal equilibrium conditions

CHAPTER 2

POWER ELECTRONIC DEVICES

2.1 POWER SEMICONDUCTOR DEVICES

Since the first thyristor of silicon-controlled rectifier (SCR) was developed in late 1957, there have been tremendous advances in the power semiconductor devices. Until 1970, the conventional thyristors had been exclusively used for power control in industrial applications. Since 1970, various types of power semiconductor devices were developed and became commercially available. These can be divided broadly into five types: (1) power diodes, (2) thyristors, (3) power bipolar junction transistors (BJTs), (4) power MOSFETs, and (5) insulated-gate bipolar transistors (IGBTs) and static induction transistors (SITs). The thyristors can be subdivided into eight types: (a) forced-commutated thyristor, (b) line-commutated thyristor, (c) gate-turn-off thyristor (GTO), (d) reverse-conducting thyristor (RCT), (e) static induction thyristor (SITH), (f) gate-assisted turn-off thyristor (GATT), (g) light-activated silicon-controlled rectifier (LASCR), and (h) MOS-controlled thyristors (MCTs). Static induction transistors are also commercially available.

The ratings of commercially available power semiconductor devices are shown in Table 2.1, where the on-voltage is on-state voltage drop of the device at the specified current. The *v–i* characteristics and the symbols of commonly used power semiconductor devices are shown in Appendix A.

Table 2.1 Ratings of power semiconductor devices

	Type	Voltage/current rating	Upper frequency (Hz)	Switching time (μs)	On-state resistance (Ω)
Diodes	General purpose	5000 V/5000 A	1k	100	0.16m
	High speed	3000 V/1000 A	10k	2-5	1m
	Schottky	40 V/60 A	20k	0.23	10m
Forced-turned-off thyristors	Reverse blocking	5000 V/5000 A	1k	200	0.25m
	High speed	1200 V/1500 A	10k	20	0.47m
	Reverse blocking	2500 V/400 A	5k	40	2.16m
	Reverse conducting	2500 V/1000 A	5k	40	2.1m
	GATT	1200 V/400 A	20k	8	2.24m
	Light triggered	6000 V/1500 A	400	200-400	0.53m
TRIACs		1200 V/300 A	400	200-400	3.57m
Self-turned-off thyristors	GTO	4500 V/3000 A	10k	15	2.5m
	SITH	4000 V/2200 A	20k	6.5	5.75m
Power transistors	Single	400 V/250 A	20k	9	4m
		400 V/40 A	20k	6	31m
		630 V/50 A	25k	1.7	15m
	Darlington	1200 V/400 A	10k	30	10m
SITs		1200 V/300 A	100k	0.55	1.2
Power MOSFETs	Single	500 V/8.6 A	100k	0.7	0.6
		1000 V/4.7 A	100k	0.9	2
		500 V/50 A	100k	0.6	0.4m
IGBTs	Single	1200 V/400 A	20k	2.3	60m
MCTs	Single	600 V/60 A	20k	2.2	18m

The applications and frequency range of power devices. A superpower device should;

(1)have a zero on-state voltage,

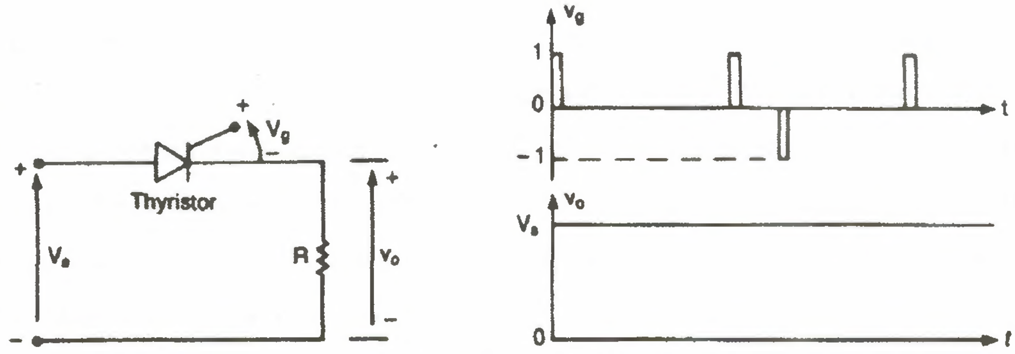
(2) withstand an infinite off-state voltage,

(3) handle an infinite current,

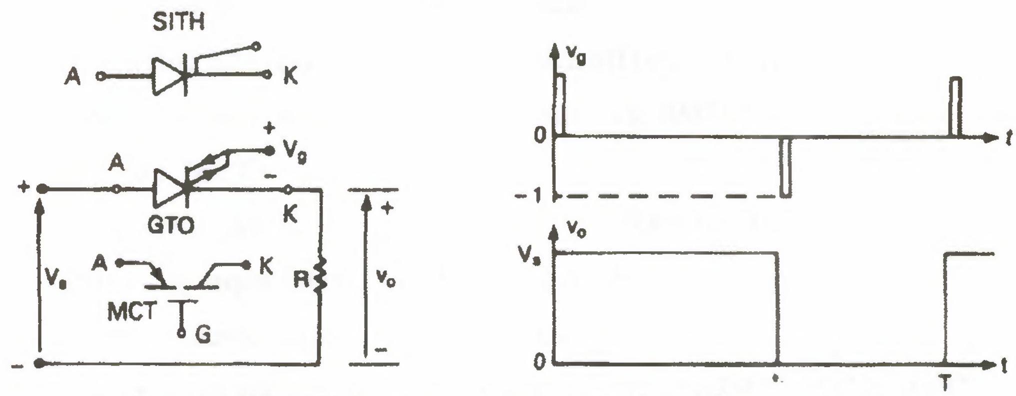
(4) turn "on" and "off" in zero time, thereby having infinite switching speed.

2.2 CONTROL CHARACTERISTICS OF POWER DEVICES

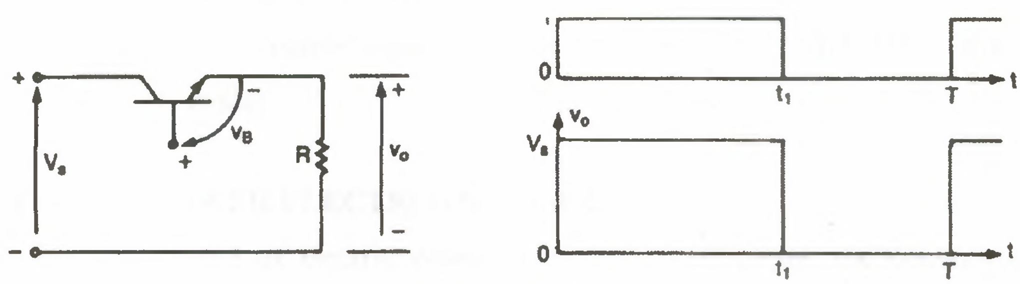
The power semiconductor devices can be operated as switches by applying control signals to the gate terminal of thyristors (and to the base of bipolar transistors).



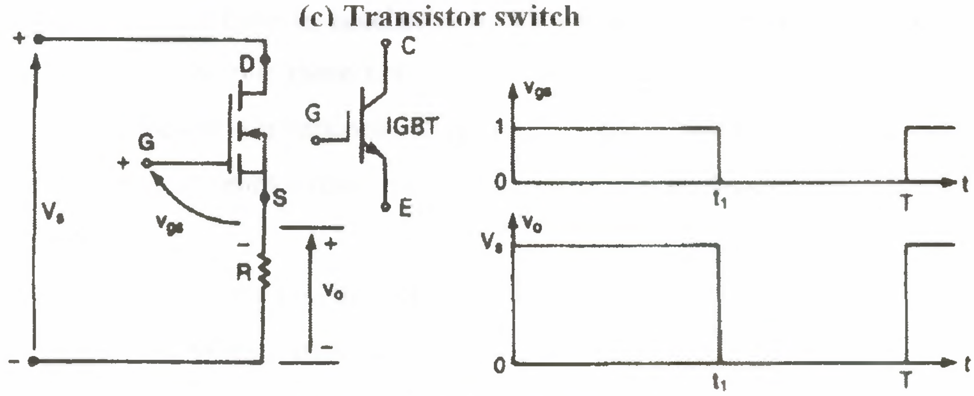
(a)Thyristor switch



(b) GTO/MCT/SITH switch (For MCT, the polarity of V_g is reversed as shown)



(c) Transistor switch



(d) MOSFET/IGBT switch

Figure 2.1 Control characteristics of power switching devices

The required output is obtained by varying the conduction time of these switching devices. Figure 2.1 shows the output voltages and control characteristics of commonly used power switching devices. Once a thyristor is in a conduction mode, the gate signal of either positive or negative magnitude has no effect and this is shown in Fig. 2.1(a). When a power semiconductor device is in a normal conduction mode, there is a small voltage drop across the device. In the output voltage waveforms in Fig. 2.1, these voltage drops are considered negligible.

The power semiconductor switching devices can be classified on the basis of:

1. Uncontrolled turn on and off (e.g., diode)
2. Controlled turn on and uncontrolled turn off (e.g., SCR)
3. Controlled turn on and off characteristics (e.g., BJT, MOSFET, GTO, SITH, IGBT, SIT, MCT)
4. Continuous gate signal requirement (BJT, MOSFET, IGBT, SIT)
5. Pulse gate requirement (e.g., SCR, GTO, MCT)
6. Bipolar voltage-withstanding capability (SCR, GTO)
7. Unipolar voltage-withstanding capability (BJT, MOSFET, GTO, IGBT, MCT)
8. Bidirectional current capability (TRIAC, RCT)
9. Unidirectional current capability (SCR, GTO, BJT, MOSFET, MCT, IGBT, SITH, SIT, diode)

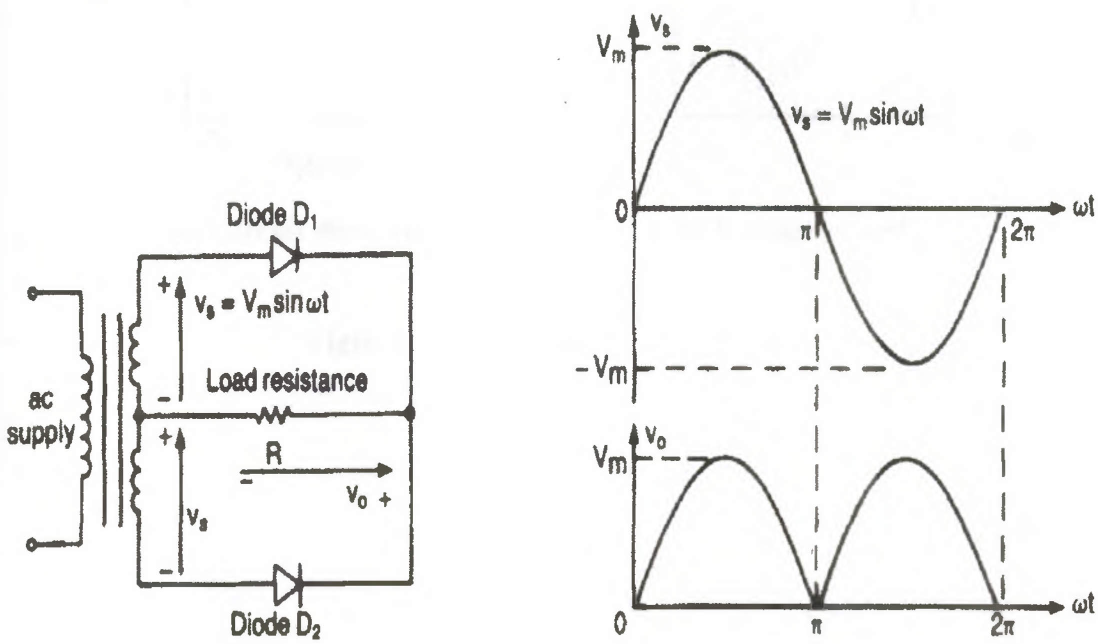
2.3 TYPES OF POWER ELECTRONIC CIRCUITS

For the control of electric power or power conditioning, the conversion of electric power from one form to another is necessary and the switching characteristics of the power devices permit these conversions. The static power converters perform these functions of power conversions. A converter may be considered as a switching matrix. The power electronics circuits can be classified into six types:

1. Diode rectifiers
2. ac-dc converters (controlled rectifiers)
3. ac-ac converters (ac voltage controllers)
4. dc-dc converters (dc choppers)
5. dc-ac converters (inverters)
6. Static switches

The devices in the following converters are used to illustrate the basic principles only. The switching action of a converter can be performed by more than one device. The choice of a particular device will depend on the voltage, current, and speed requirements of the converter.

1) Rectifiers. A diode rectifier circuit converts ac voltage into a fixed dc voltage and is shown in Fig. 2.2. The input voltage to the rectifier could be either single-phase or three-phase.

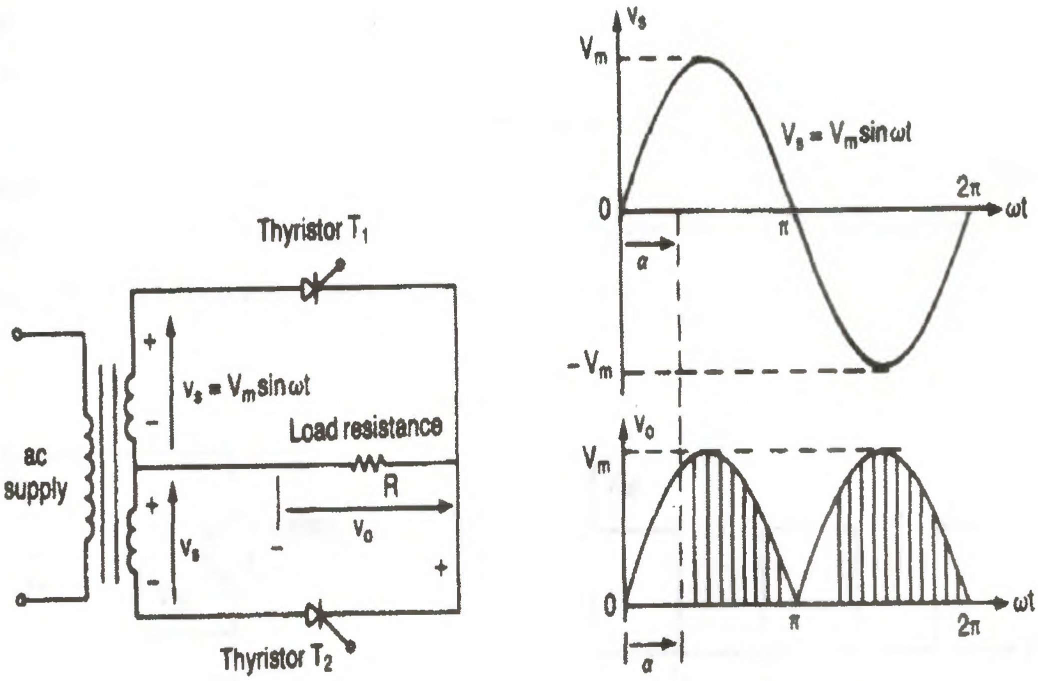


(a) Circuit diagram (b) Voltage waveforms

Figure 2.2 Single-phase rectifier circuit

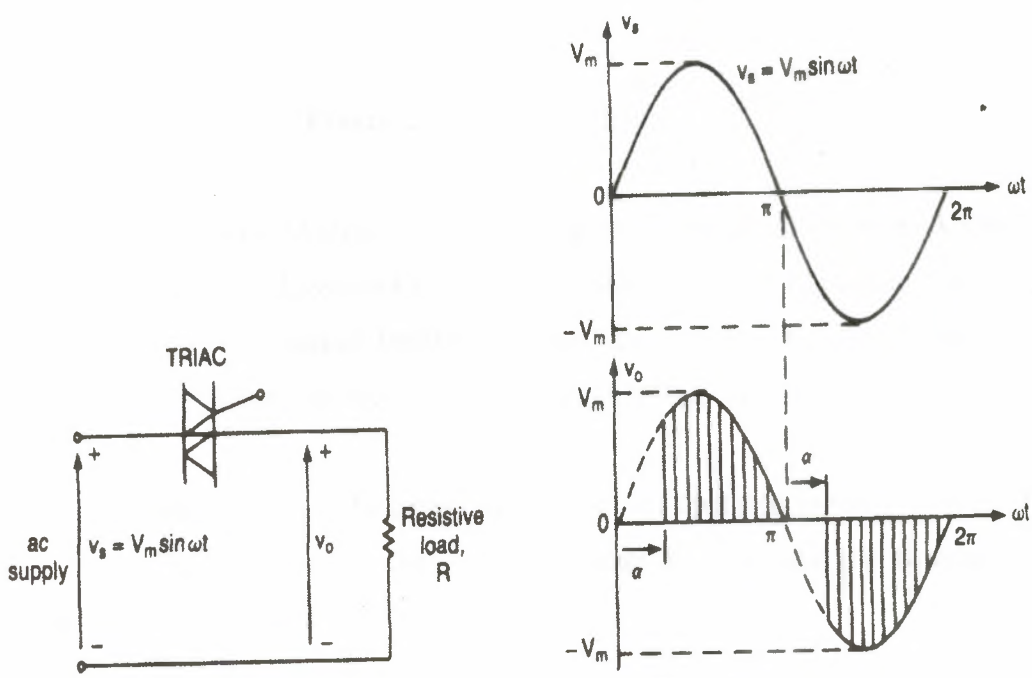
2) Ac-dc converters. A single-phase converter with two natural commutated thyristors is shown in Fig. 2.3. The average value of the output voltage can be controlled by varying the conduction time of thyristors or firing delay angle, α. The input could be a single or three-phase source. The converters are also known as *controlled rectifiers*.

3) Ac-ac converters. These converters are used to obtain a variable ac output voltage from a fixed ac source and a single-phase converter with a TRIAC is shown in Fig. 2.4. The output voltage is controlled by varying the conduction time of a TRIAC or



(a) Circuit diagram (b) Voltage waveforms

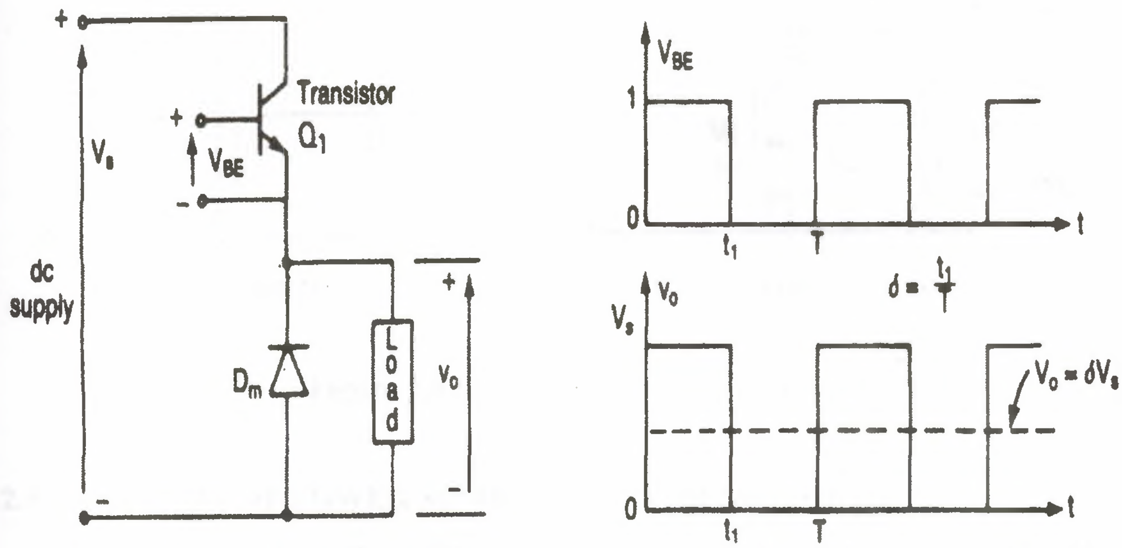
Figure 2.3 Single-phase ac-dc converter



(a) Circuit diagram (b) Voltage waveforms

Figure 2.4 Single-phase ac-ac converter firing delay angle, α. These types of converters are also known as *ac voltage controllers*.

4) Dc-dc converters. A dc-dc converter is also known as a *chopper* or *switching regulator* and a transistor chopper is shown in Fig. 2.5. The average output voltage is controlled by varying the conduction time *t*, of transistor Q_1. If T is the chopping period, then $t_1 = \delta T$. δ is called as the *duty cycle* of the chopper.

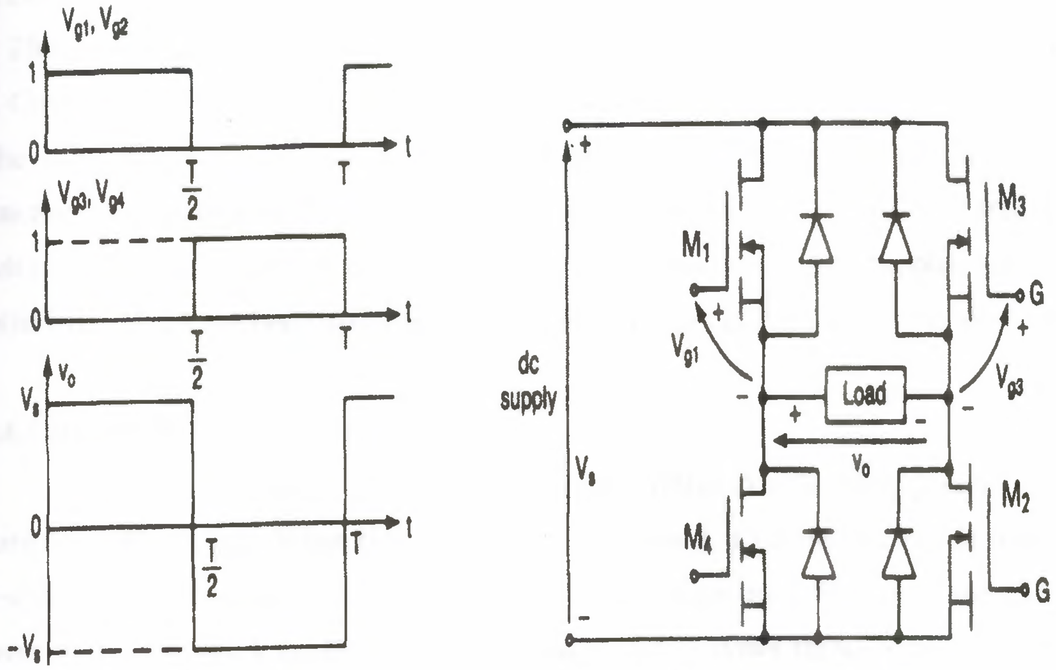


(a) Circuit diagram (b) Voltage waveforms

Figure 2.5 Dc-dc converter

5) Dc-ac converters. A dc-ac converter is also known as an *inverter*. A single-phase transistor inverter is shown in Fig. 2.6. If transistors M_1 and M_2 conduct for one-half period and M_3 and M_4 conduct for the other half, the output voltage is of alternating form. The output voltage can be controlled by varying the conduction time of transistors.

6) Static switches. Since the power devices can be operated as static switches or contactors, the supply to these switches could be either ac or dc and the switches are called as *ac static switches* or *dc switches*.



(a) Circuit diagram (b) Voltage waveforms

Figure 2.6 Single-phase dc-ac converter

2.4 OVERVİEW OF POWER SEMICONDUCTOR SWITCHES

The increased power capabilities, ease of control, and reduced costs of modern power semiconductor devices compared to those of just a few years ago have made conveners affordable in a large number of applications and have opened up a host of new converter topologies for power electronic applications. In order to clearly understand the feasibility of these new topologies and applications, it is essential that the characteristics of available power devices be put in perspective. To do this, a brief summary of the terminal characteristics and the voltage, current, and switching speed capabilities of currently available power devices are presented in this chapter.

If the power semiconductor devices can be considered as ideal switches, the analysis of converter topologies becomes much easier. This approach has the advantage that the details of device operation will not obscure the basic operation of the circuit. Therefore, the important converter characteristics can be more clearly understood. The summary of device characteristics will enable us to determine how much the device characteristics can be idealized.

Presently available power semiconductor devices can be classified into three groups according to their degree of controllability:

1. *Diodes.* On and off states controlled by the power circuit.

2. *Thyristors.* Latched on by a control signal but must be turned off by the power circuit.

3. *Controllable switches.* Turned on and off by control signals.

The controllable switch category includes several device types including bipolar junction transistors (BJTs), metal-oxide-semiconductor field effect transistors (MOSFETs), gate turn off (GTO) thyristors, and insulated gate bipolar transistors (IGBTs). There have been major advances in recent years in this category of devices.

2.4.1 DIODES

Figures 2.7(a) and 2.7(b) show the circuit symbol for the diode and its steady-state *i-v* characteristic. When the diode is forward biased, it begins to conduct with only a small forward voltage across it, which is on the order of 1 V. When the diode is reverse biased, only a negligibly small leakage current flows through the device until the reverse breakdown voltage is reached. In normal operation, the reverse-bias voltage should not reach the breakdown rating.

In view of the very small leakage currents in the blocking (reverse-bias) state and the small voltage in the conducting (forward-bias) state as compared to the operating voltage and currents of the circuit in which the diode is used, the *i-v* characteristics for the diode can be idealized, as shown in Fig. 2.7(c). This idealized characteristic can be used for analyzing the convener topology but should not be used for the actual design, when, for example, heat sink requirements for the device are being estimated.

At turn-on, the diode can be considered an ideal switch because it turns on rapidly compared to the transients in the power circuit. However, at turn-off, the diode current reverses for a reverse-recovery time t_{rr}, as is indicated in Fig. 2.8, before falling to zero. This reverse-recovery (negative) current is required to sweep out the excess carriers in the diode and allow it to block a negative polarity voltage. The reverse-recovery current can lead to overvoltages in inductive circuits. In most circuits, this reverse current does not affect the converter input/output characteristic and so the diode can also be considered as ideal during the turn-off transient.

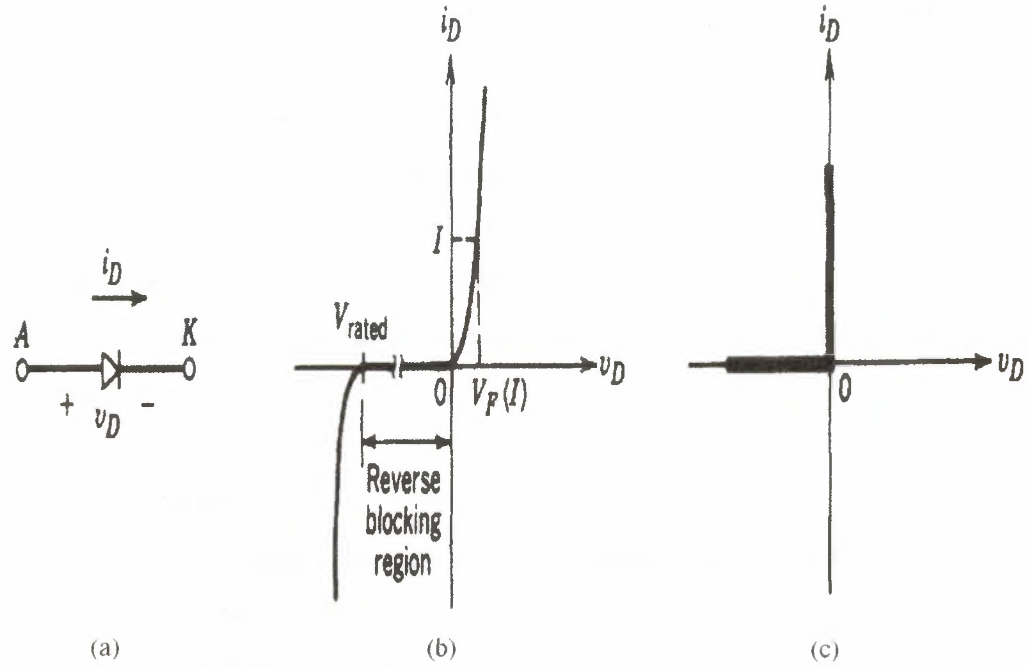


Figure 2.7 Diode: (a) symbol, (b) *i-v* characteristics, (c) idealized characteristics

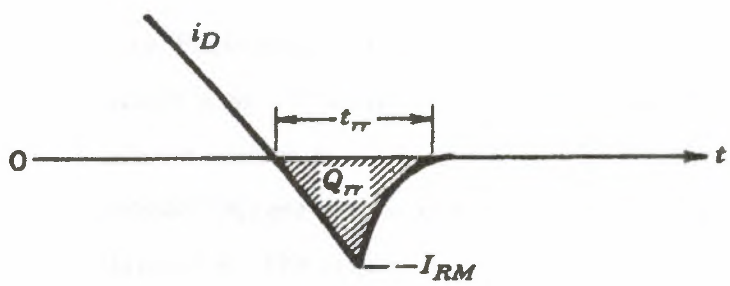


Figure 2.8 Diode turn-off

Depending on the application requirements, various types of diodes are available:

1. *Schottky diodes. These* diodes are used where a low forward voltage drop (typically 0.3 V) is needed in very low output voltage circuits. These diodes are limited in their blocking voltage capabilities to 50-100 V.

2. *Fast-recovery diodes. These* are designed to be used in high-frequency circuits in combination with controllable switches where a small reverse-recovery time is needed. At power levels of several hundred volts and several

hundred amperes, such diodes have t_{rr} ratings of less than a few microseconds.

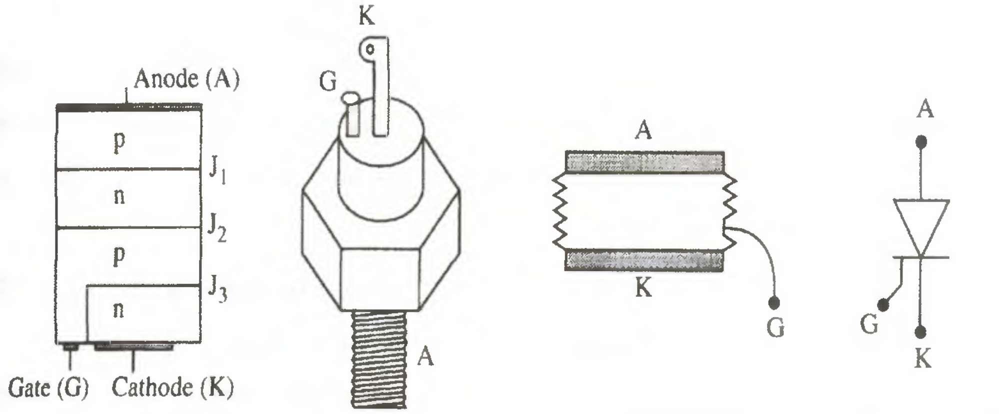
3. *Line-frequency diodes.* The on-state voltage of these diodes is designed to be as low as possible and as a consequence have larger t_{rr} which are acceptable for line-frequency applications. These diodes are available with blocking voltage ratings of several kilovolts and current ratings of several kiloamperes. Moreover, they can be connected in series and parallel to satisfy any voltage and current requirement.

2.4.2 THE THYRISTOR

The thyristor, also known as the silicon controlled rectifier (SCR), was the first solid state power semiconductor device to be developed to function as a controlled static switch, with large current and voltage capability. It was the advent of the thyristor that gave start to a new era of major developments in static power conversion and control, which has made rapid strides in recent years, making "Power Electronics" a recognized technology in its own right.

2.4.2.1 Junction Structure, Packaging, Circuit Symbol

The junction structure is shown in Fig. 2.9(a). This is a four-layer structure with three internal junctions shown labeled as J_1, J_2, and J_3. The device has three terminals. The "anode" (A) and "cathode" (K) are the power terminals of the switch. Control input is between the "gate" (G) and K. The anode metallic contact is on the outer *p* layer shown on top. The cathode contact is on the outer *n* layer. The inner *p* layer is the gate layer. During the fabrication of the device, the regions where the gate electrical contact is to be made are masked, before the cathode *n* layer is formed. Since the gate *p* layer lies sandwiched between two *n* layers, it is impossible to make external electrical contact with the entire gate layer area. Actually, the gate electrical contact is limited to a relatively small fraction of the total wafer area, the rest of it being allocated for the cathode contact. The geometrical pattern of the gate contact area on the surface of the pellet significantly affects the switching characteristics of the device. We shall have occasion to highlight this later on. We shall see that the absence of direct electrical contact between the gate terminal and a major part of the gate p region is responsible for limiting the role of the gate to the turn ON function only.



(a) Junction structure (b) Stud-type package (c) Disc-type package (d) circuit symbol

Figure 2.9 The thyristor structure, types of packaging and circuit symbol

When a forward voltage (positive voltage polarity at the anode terminal) exists across the main terminals of the thyristor, a short current pulse from gate to cathode will "fire" the thyristor, that is, trigger it into the ON state. Once the thyristor is fired, the gate has no further control over the current flow through the device. During the subsequent conduction, it behaves like a diode. It cannot be turned OFF by a reverse current pulse on the gate.

The two commonly available types of casings in which thyristor pellets are packaged are shown in Figs. 2.9(b) and (c). The stud-type package is similar to the stud-type package for power diodes, except that there is an additional terminal—the gate. In a thyristor, the stud is always the anode. The cathode and gate leads, whose internal contacts with the pellet are on the side opposite to the anode, are brought out with insulation from the casing, as shown on the top in (b). The stud is used for mounting the device on a metallic heat sink. The disc-type package, shown in (c), is similar to the corresponding diode package, except that a gate lead is present. This comes out on the side through the ceramic housing that separates the flat anode and cathode terminals. In some designs, an additional cathode lead (the gate return) is also provided as a thin flexible cable, along with the gate lead, so that these twin leads can be conveniently connected to the external gate control circuit. Power connections are made through pressure contacts with the flat surfaces of the anode and cathode, using special mounting hardware. The flat surfaces provide large contact area, both for electric currents and for heat flow. For better cooling, it is advisable to use two separate heat sinks, insulated from each other, one on each side of the device.

Figure 2.9(d) shows the circuit symbol for the thyristor. This is derived from that for the diode, with the addition of the gate terminal. The gate terminal location near the cathode is in conformity with the internal geometry and the fact that the firing control input is always between the gate and the cathode.

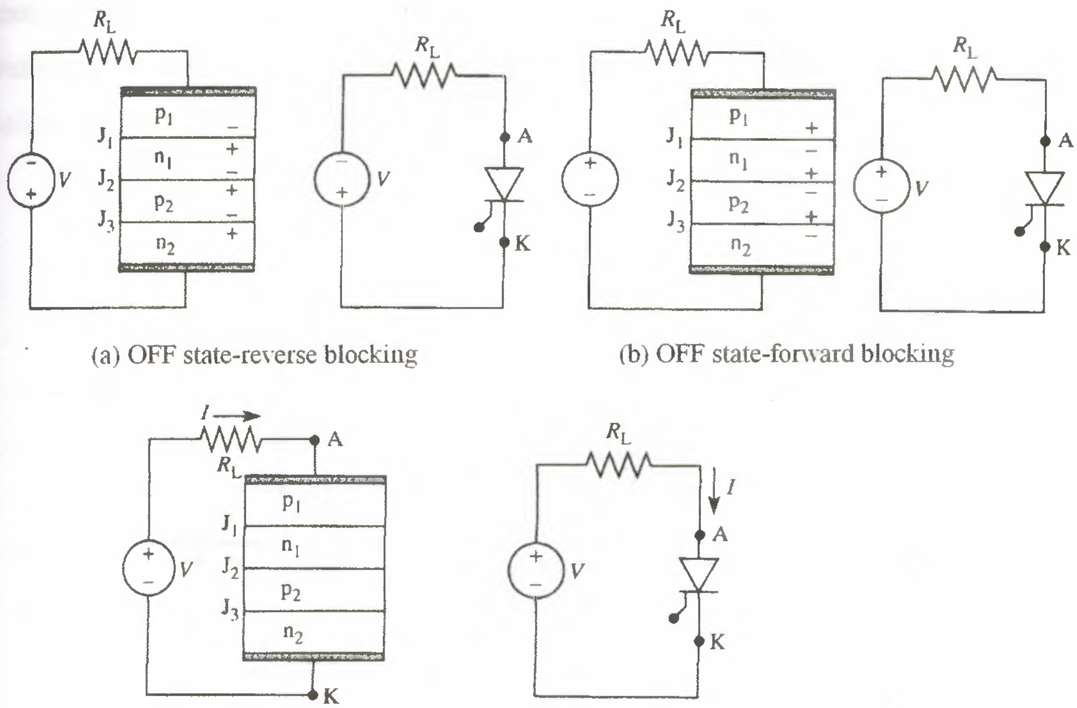
2.4.2.2 Operating States of the Thyristor

The thyristor can exist in one of three alternative states in circuit operation. Two of these are OFF states—the reverse blocking OFF state and the forward blocking OFF state. The third is the forward conducting ON state. It can stay in each of these states without an electrical input being present on the gate terminal. The gate serves only to implement the transition from the forward blocking OFF state to the forward conducting ON state. Therefore in our present discussion using Fig. 2.10, we are not showing any electrical connection to the gate terminal.

Figures 2.10(a), (b) and (c) show the three operating states of the thyristor. In each condition, figures are drawn—one showing the junction structure and the second showing the circuit with the circuit symbol. The thyristor is shown connected in series with a resistance to a voltage source with polarity appropriate to each condition.

In (a), the source polarity is such as to reverse-bias the thyristor. With such a polarity, the thyristor can only exist in the OFF state. The reverse blocking voltage between A and K is distributed serially across the three junctions. These junction voltage polarities are marked in (a). It will be seen that in this condition, J_1 and J_3 are reverse-biased and J_2 is forward-biased. Because of the reverse-biased junctions in series, the thyristor cannot conduct, except for the inevitable small leakage current, which we shall neglect. It would appear that the reverse voltage capability of the thyristor will be decided by the total reverse voltage capability of J_1 and J_3. In practical thyristors, the reverse voltage breakdown limit of J_3 is very small, usually well below 10 V. Therefore the reverse voltage rating of a thyristor is almost entirely determined by the breakdown limit of the junction J_1.

In (b), the source voltage polarity is such as to forward-bias the thyristor. Therefore the junction voltage polarities are opposite to those in (a). Now J_2 is reverse-biased, and the thyristor still cannot conduct because of this. We now have the forward blocking OFF state. The forward blocking voltage capability of a thyristor is therefore



(a) OFF state-reverse blocking

(b) OFF state-forward blocking

(c) ON state-forward conducting

Figure 2.10 Operating states of the thyristor switch

determined by the breakdown limit of the junction J_2. In practical thyristors, the forward blocking voltage rating (decided by J_2) is about the same as the reverse blocking voltage rating (decided by J_1). The thyristor is a "symmetrical voltage blocking" device. The symmetrical voltage blocking ability is a result of the structure for the following reason.

During the fabrication of the device, the junctions J_1 and J_2 are created simultaneously by a single diffusion operation, during which *p* type impurity atoms are diffused into both sides of a *n* type silicon wafer of high purity. The cathode *n* layer is formed subsequently by a second diffusion, this time only on one side of the wafer. During this second diffusion, some areas on the cathode side are also masked, to provide space for the metal deposition for the gate electrode. From this, it will be seen that J_1 and J_2 have similar characteristics and about the same breakdown limit. These breakdown limits are large, because of the high purity of the starting *n* type wafer. Since J_3 has higher impurity levels on both sides, its breakdown limit is much smaller.

2.4.2.3 Turn ON Switching, Two-Transistor Analog

The turn ON switching of a thyristor is best explained using the "two-transistor" analogy. Figure 2.11(a) shows the thyristor junction structure as a composite of a *pnp* transistor T_1 and an *npn* transistor T_2, by visualizing an imaginary plane through the pellet as shown by the broken line.

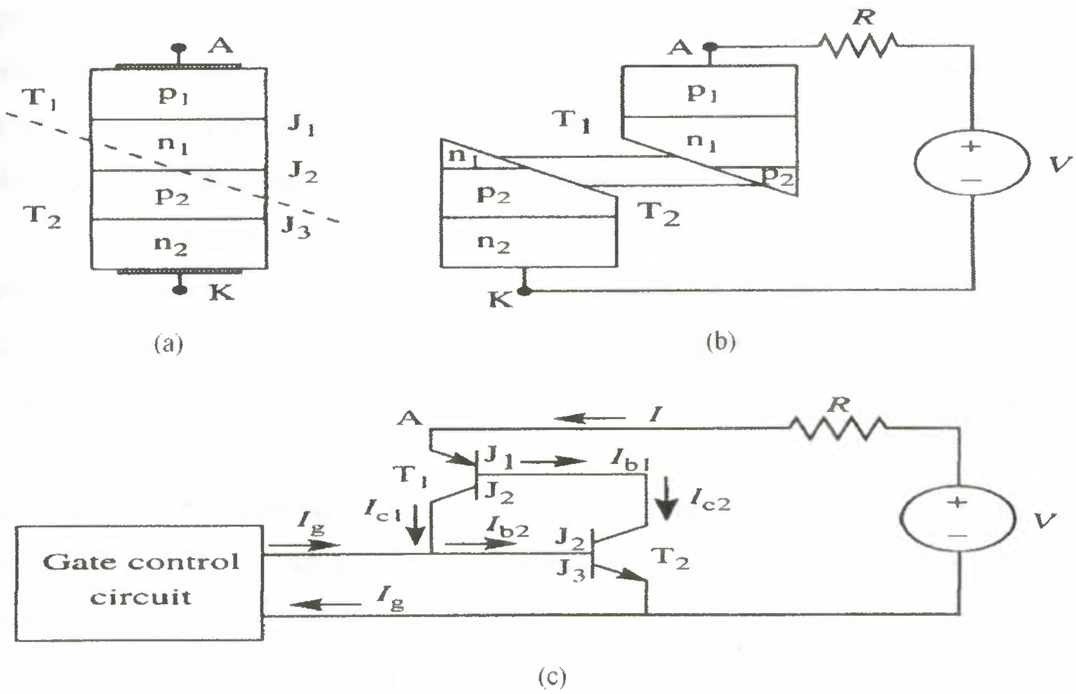


Figure 2.11 Two-transistor equivalent of the thyristor structure to explain turn ON switching

For greater clarity, the two transistors are shown physically separated in (b), but with the common layers connected together. The thyristor is shown forward-biased by an external source, which makes the anode positive with respect to the cathode. In the case of the *pnp* transistor, the anode layer p_1 functions as the emitter. In the case of the *npn* transistor, the cathode layer n_2 functions as the emitter. On this basis, the circuit is redrawn in (c) using the appropriate transistor symbols. Included in (c) are the voltage source V and a load resistor R. Also shown is a gate control circuit, whose output is connected between the gate and the cathode. This serves to send a gate current I_g in the direction shown, when desired.

Initially I_g is zero. Both transistors are OFF. If we now send a small gate current, this serves as the base current I_{b2} for the transistor T_2. Therefore a collector current I_{c2}

results. Inspection of the circuit shows that I_{c2} serves as the base current I_{b1} for transistor T_1. Because of I_{b1}, a collector current I_{c1} is initiated, I_{c1} serves as additional base current for T_2, causing further increase in I_{c2}. This in turn causes further increase of the base current of T_2; and therefore of I_{c1}. In this way, a regenerative current build up process takes place, and both T_1 and T_2 drive each other into the saturated ON state. This happens in a matter of a few microseconds. Once turned ON, the two transistors mutually supply each other's base current, and there is no need for an external gate current to maintain the ON state. The thyristor stays in the ON state with a small forward voltage drop, which is usually in the neighborhood of 2 V for a high power device.

In fact, the gate current is needed only until the current builds up to a certain level, after which the regenerative process will continue even without the external gate current, and the turn ON switching will be completed. The minimum value of the current in the external circuit after which the turn ON switching will be completed without further gate supply being necessary is called the latching current. We notice that the gate serves only to implement the turn ON switching transition. After turn ON, the gate loses control and it is not possible to implement turn OFF switching by means of a reverse gate current.

In practical thyristors there is a minimum current necessary to maintain the device in the ON state. If we decrease V or increase R, the thyristor will turn OFF when the current tends to fall below this minimum level. The minimum current necessary to keep the thyristor in the ON state is called the "holding current." The holding current is lower than the latching current in practical thyristors. In an ideal thyristor, we shall assume the holding current to be zero. Such a thyristor behaves as a diode after it has been turned ON. The gate regains control only after the thyristor has turned OFF, typically by a reversal of the polarity of the anode-to-cathode voltage.

2.4.2.4 Why Turn OFF is Impossible by Reverse Gate Pulse

We shall take a closer look at the reason why a thyristor cannot be turned OFF by a reverse gate current pulse. We do this for two reasons. First, examination of this question gives us a better understanding of an important parameter of a thyristor—its *di/dt* limit. Second, this prepares us for the understanding of another important static power switching device, the gate turn off thyristor (GTO) which we will consider later, and which has a similar junction structure, but can in fact be turned OFF by a reverse gate current pulse.

In Fig. 2.12, the single thyristor T is viewed as consisting of many in parallel, such as T_1, T_2, and so on. Of these, only T_1 is shown to have an external gate terminal, it being assumed that T_1 is the one nearest to the gate electrode. Therefore, when a gate current pulse I_{g1} arrives to implement turn ON switching, the first of these thyristors to turn ON is T_1. Thereby, a current I_1 is created in this thyristor. Part of this current flows through the gate layer of T_2 and thus provides the gate current I_{g2} to turn ON T_2. In a similar way, T_2 turns ON T_3 by providing its gate current I_{g3}. In this way, all the thyristors that constitute the device turn ON, one after the other, beginning with the one nearest to the gate electrode layer. Once this has taken place, the gate is extremely ineffective in influencing the current in the elementary thyristors which are in remote locations from the gate electrode. The main reason why the gate is unable to turn OFF the device is the fact that the gate electrode layer is not in close proximity to a large part of the cathode area.

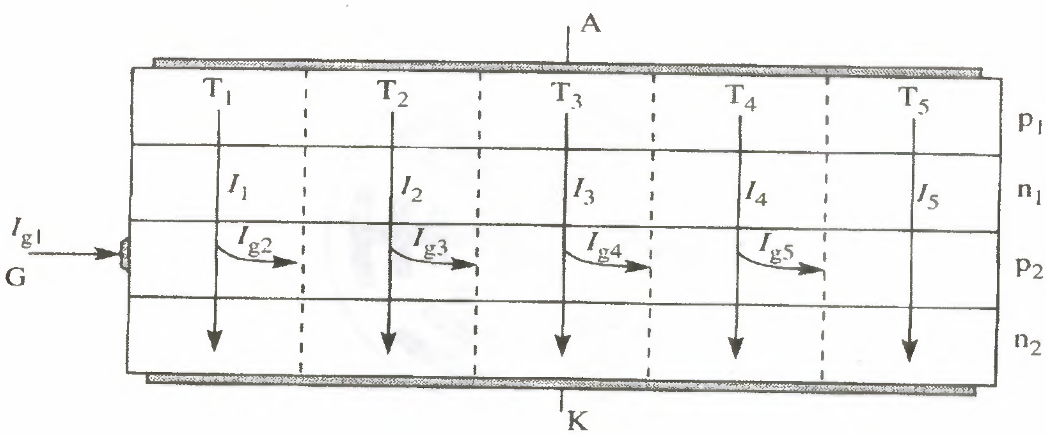


Figure 2.12 A single thyristor viewed as several individual cells in parallel.

The above consideration also shows that there is a finite time needed for the switching ON operation to spread across the entire area of the silicon wafer. This means that the thyristor switch behaves as though it consisted of several switches connected in parallel, all of which do not turn ON simultaneously. One turns ON, this causes the next to turn ON, and in this manner the switching progresses until every one comes ON. If we now look at the external power circuit, the current that flows does not depend on how many of the switches have turned ON at a given instant of time. Therefore, if the current in the external circuit rises at too fast a rate in comparison with the rate at which switching progresses across the area of the thyristor pellet, excessive local current density can occur in regions close to the gate electrode and cause permanent damage to the device. In other words, the thyristor has a *di/dt* limit that should not be exceeded to avoid damage to it. This is one of the parameters specified by the manufacturer in the data sheet of the thyristor.

Interdigitated gate geometry. In the practical design of thyristors, one technique often used to increase the *di/dt* rating of the device is to use a gate electrode pattern that results in the gate electrode boundaries being within short distance of any part of the cathode area. This is achieved by distributing the gate electrode area over the pellet and interleaving the gate and cathode regions on it. Such a gate electrode geometry is referred to as an interdigitated gate structure. Figure 2.13 shows an example of an interdigitated gate pattern.

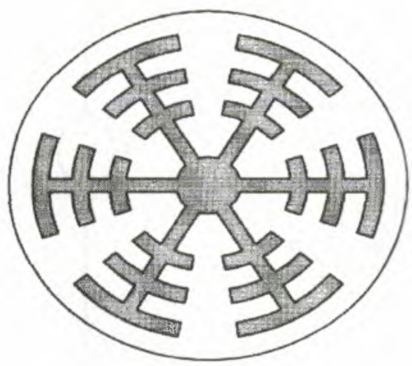


Figure 2.13 An interdigitated gate-cathode geometry to achieve a high *di/dt* capability. (Gate areas are shown shaded; the remainder is the cathode area)

2.4.2.5 Rate of Rise of Forward Voltage, *dv/dt* Rating

For reliable operation, a static thyristor switch should turn ON only when a gate pulse is applied to it, and for no other reason. One physical reason that can cause a thyristor to turn ON without a gate pulse is an excessive rate of rise of forward voltage (dv/dt). How this can happen is explained with reference to Fig. 2.14. The thyristor shown there is in the forward-biased OFF state. No gate connection is shown. Junction J_2 is reverse-biased and is blocking the source voltage. A reverse-biased pn junction, because of the charge distribution on either side of the junction boundary plane, behaves like a charged capacitance. If this capacitance is denoted by C, the associated charge will be

$$Q=CV$$

If the voltage V increases at a rate dV/dt, there will be an associated charging current of this capacitor, which is dependent on dV/dt. The higher the value of dV/dt, the larger will be the magnitude of this charging current. This charging current flows from the n_1 zone to the p_2 zone across the junction J_2, and is denoted by i_c in Fig. 2.14. The current i_c can cause the turn ON switching of the thyristor in exactly the same manner as a gate current pulse injected into the p_2 zone through the gate terminal, if it exceeds the threshold value necessary for turn ON. This means that the thyristor has a maximum dV/dt limit beyond which it will turn ON without the presence of a gate pulse.

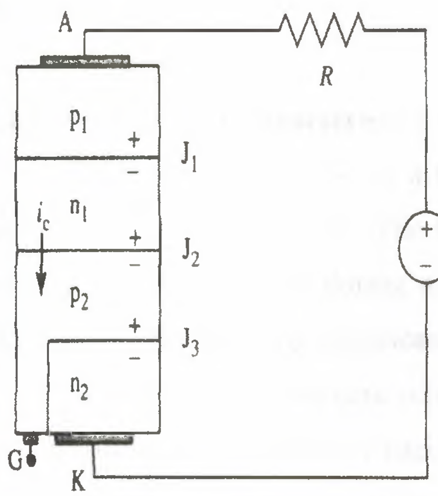


Figure 2.14 *dv/dt* limit of a thyristor due to junction capacitance charging current

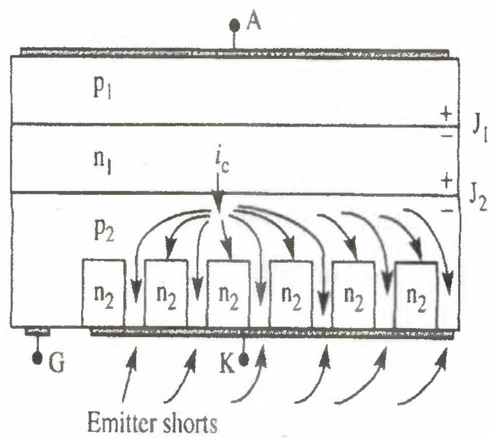


Figure 2.15 Shorted emitter construction of a thyristor to achieve a high *dv/dt* rating

When devices with high dv/dt are required, the technique used to achieve this is called the shorted emitter construction. This may be explained with reference to Fig. 2.15. This shows the cathode (n_2 region) as made up of several n_2 islands instead of one continuous layer. The cathode electrode metal deposition, however, covers not only the n_2 islands, but also the p_2 areas between these n_2 islands. Therefore the cathode electrode metal layer short-circuits the p_2 and n_2 areas at the boundaries between these areas. In the two-transistor equivalent of a thyristor given earlier in Fig. 2.11, the p_2 layer functions as the base of the *npn* transistor, and the n_2 layer functions as the emitter. Therefore in this arrangement we are creating local short circuits between the emitter and some areas of the base. These areas are labeled as "emitter shorts" in Fig. 2.15. The capacitance current that is caused when there is a dv/dt is also shown in the figure labeled i_c.

We notice that these emitter shorts provide an alternative path through which ;c can flow to the cathode terminal, without crossing into the cathode n_2 zone. Therefore part of i_c is bypassed directly to the cathode terminal through the emitter shorts and does not contribute as a gate current to cause turn ON switching. In this way, the shorted emitter construction serves to raise the dv/dt threshold for turn ON switching. It should be noted that when we send a gate current pulse to fire the thyristor, some of this current is also bypassed through the emitter shorts. Therefore a thyristor with the shorted emitter construction has a comparatively higher gate current threshold for turn ON switching.

2.4.2.6 Switching Characteristics -- Turn ON Time

The turn ON time t_{ON} of a thyristor may be defined by reference to the circuit and waveforms of Fig. 2.16,. The figure shows typical waveforms of the gate voltage and the thyristor current during a turn ON switching transition. The timing instants shown have the following significance.

t_1 Instant at which the gate voltage rises to 10% of the final value

t_2 Instan*t* at which the thyristor current rises to 10% of the final ON state value

t_3 Instant at which the thyristor current rises to 90% of the final ON state value

Using the above as reference instants, we define the following:

$$t_d = t_2 - t_1 = \text{delay time}$$

$$t_r = t_3 - t_2 = \text{rise time}$$

The turn ON time is defined as

$$t_{ON} = t_d + t_r$$

The turn ON time is usually in the range of a few microseconds. The geometry of the gate structure has a significant effect on t_{ON}. An interdigitated gate reduces the time needed for the turn ON switching to spread over the entire cathode area. A shorter turn ON time is therefore usually associated with a higher *di/dt* rating.

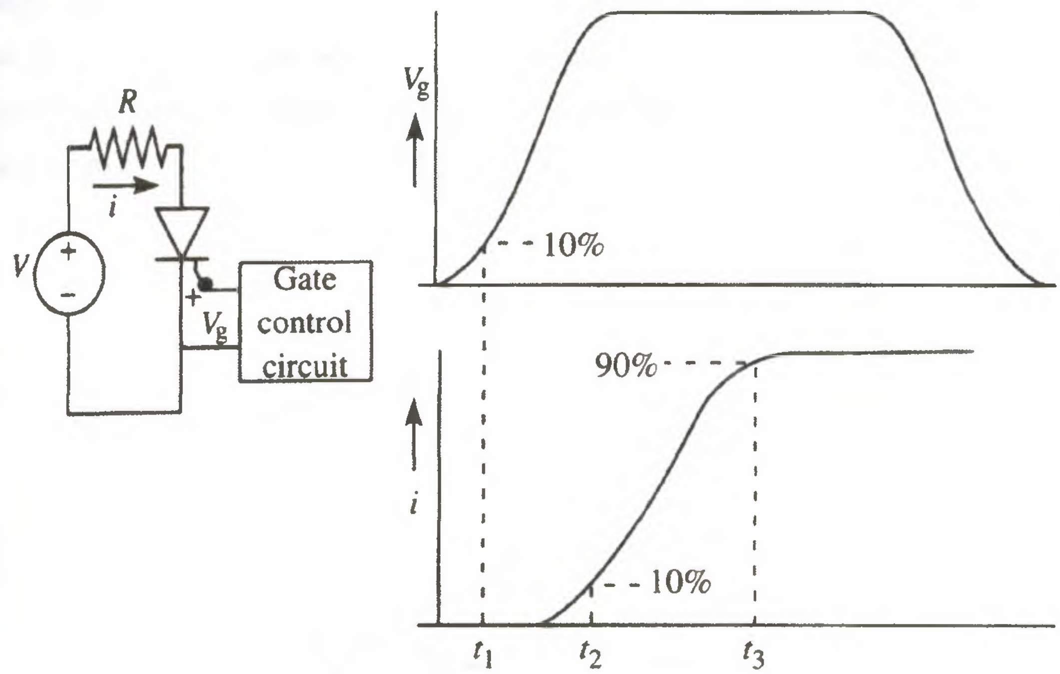


Figure 2.16 Definition of turn ON time for a thyristor

2.4.2.7 Switching Characteristics -- Turn OFF Time

The turn OFF switching of a thyristor occurs when a reverse voltage is made to appear across the main terminals of the switch, in the same manner as a diode gets turned OFF. The corresponding waveforms for a thyristor are shown in Fig. 2.17(a). Comparison will show that both are similar until the instant labeled t_3 in both the figures. This is the instant at which each device is able to block the full reverse voltage.

In the case of a thyristor, the turn OFF switching is considered complete only when the device has also regained its ability to block forward voltages. So the question is whether the device is capable of blocking a forward voltage from the instant t_3.

To find the answer to this, we can experiment by applying a forward voltage across the device immediately after t_3, without any input to the gate terminal. It will then be found that the device, in fact, turns ON without a gate pulse, as if it were a diode.

The thyristor actually needs a finite time after t_3 to regain its ability to block forward voltages. To ascertain the minimum time required for this, we can repeat the experiment several times, each time increasing the time delay in very small steps after t_3, before applying the forward voltage. We do this until we arrive at the condition when the device just stops turning ON when forward-biased. Alternatively, we can begin with a long delay initially and repeat the test, progressively decreasing the time delay in small steps each time, until the device starts turning ON when forward-biased. Whichever way we do the test, it is this minimum delay condition that is shown in Fig. 2.17(a). The current and voltage waveforms of Fig. 2.17(a) are used for defining the turn OFF time of the thyristor.

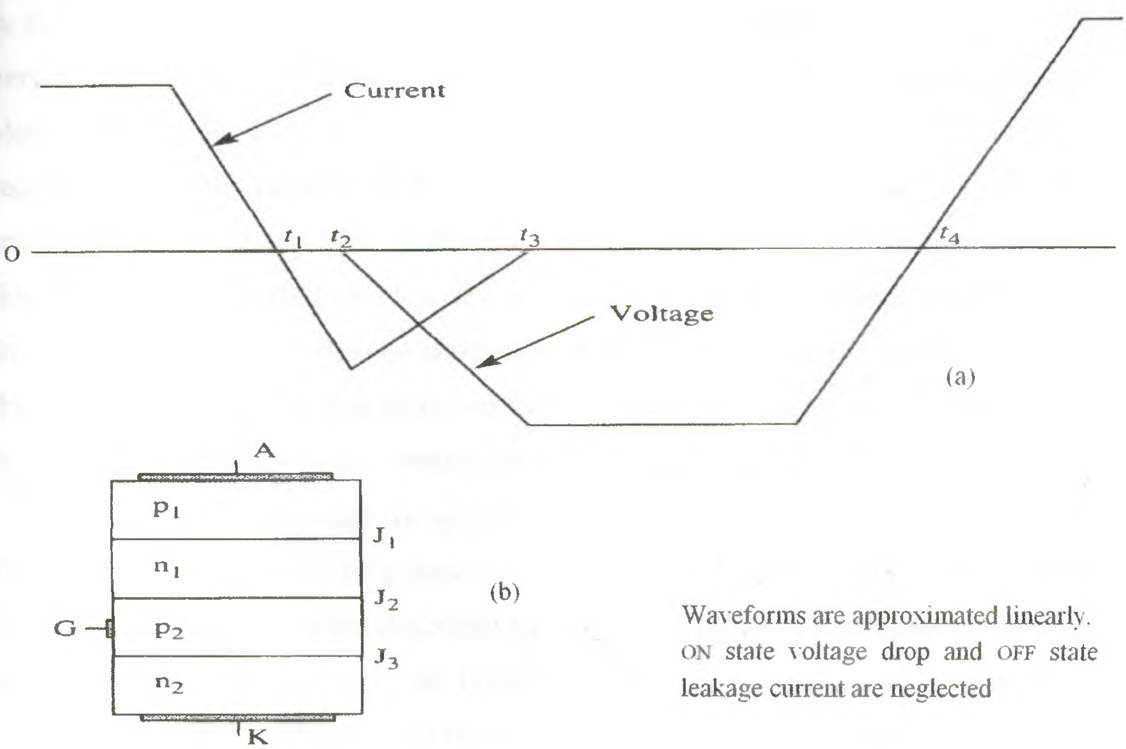


Figure 2.17 Turn OFF time definition for a thyristor

By definition, t_{OFF} is the time interval from the instant t_1 to t_4 in Fig. 2.17(b). t_1 is the instant at which the forward current has fallen to zero during the turn OFF switching transition. t_4 is the instant at which the voltage across the thyristor crosses zero and the device begins to get forward-biased. This is the earliest instant at which the device is able to block forward voltages. It is at this instant that the gate terminal has regained control over subsequent turn ON switching.

The reason for the finite time delay necessary for the device to attain forward blocking ability, after recovering reverse blocking capability, may be explained by reference to the junction structure of the thyristor shown in 2.17(b). The junction J_1 is primarily responsible for blocking reverse voltages. Therefore, when excess minority carriers on either side of this junction have disappeared, the thyristor has recovered the reverse voltage blocking capability. This happens when the reverse recovery current transient has ended at t_3. But the junction that provides the forward voltage blocking ability is J_2. It takes a finite time for the excess minority carriers on either side of this junction to disappear after the reverse recovery current transient. Actually, a reverse current for J_1 is a forward current for J_2. The excess minority carriers on the n_1 side of J_2 will be holes injected from the p_2 zone. Those on the p_2 side will be electrons from the n_1 side, but these will be relatively few because, in practical devices, the n_1 layer has a very high purity level. Therefore the primary reason for the delay to regain the forward blocking ability is the time needed for the excess holes in the n_1 zone to disappear by recombination. By speeding up the recombination process, it is possible to achieve a reduction in the turn OFF time. A technique used for this purpose in the manufacture of fast thyristors is to diffuse gold as an impurity. Gold atoms have been found to create "recombination centers," thereby shortening the lifetime of minority carriers. However, the use of gold diffusion also increases the ON state voltage drop, so that a trade off in this respect is necessary in the manufacture of fast thyristors.

If a forward voltage is applied before the excess holes have completely disappeared then these will be pushed across J_2. This will constitute a gate current in the same way as the *dv/dt* current described earlier. The thyristor will therefore turn ON in a manner similar to the *dv/dt* turn ON. From this, it is apparent that the *dv/dt* at the instant t_4, in Fig. 2.17(a) should affect the turn OFF time of the thyristor. In other words, if our *dv/dt* during the testing to determine the instant t_4 is lower, t_4 will occur earlier and vice versa. The *dv/dt* in this context is called the reapplied *dv/dt.* The reapplied *dv/dt is* one of the test conditions to be stated if the turn OFF time of a thyristor is to be specified in an exact manner, because t_{OFF} will be shorter if reapplied *dv/dt* is lower, and vice versa. The usual *dv/dt* specification of a thyristor is accurate only if the thyristor has been OFF for a long enough time for the junction J_2 to completely recover.

2.4.2.8 Thyristor Classification According to Switching Times and Thyristor Selection According to Converter Types

Based on switching times, thyristors are often listed under two categories. Fast switching thyristors (with short t_{OFF}) belong to one class, while slower thyristors belong to the other. The user has to choose the category, according to the type of converter where they are to be used. In this context, we may state here that thyristor converters are classified under two types: (1) line commutated converters and (2) force commutated converters. Line commutated converters generally work at relatively low power line frequencies such as 50 or 60 Hz. In such converters, the relatively longer turn OFF times of the slower category thyristors have negligible adverse effect on the converter performance. It is therefore always preferable to choose the slower category of thyristors, both from the point of view of cost and that of performance, because the faster switching times of the other category involve a trade off in other benefits, such as forward voltage drop. For this reason, this class of thyristors are often separately listed as "line commutated" types.

In general, force commutated converters operate at higher repetitive switching frequencies, and therefore require fast switching thyristors. Force commutated converters employ special "force commutation circuits" for the specific purpose of implementing the turn OFF switching of the thyristors by forcing a reverse voltage across the thyristor to be turned OFF, at the appropriate instant in the converter switching cycle. The sizes of the circuit elements needed in the force commutation circuits depend on how long the thyristor to be turned OFF has to be maintained in reverse bias, that is, on the turn OFF time of the thyristor. Fast thyristors are invariably chosen for force commutated converters, even if the repetitive switching frequency of the converter is low, because in this way the size and cost of the force commutation circuits will be less.

2.4.2.9 Gate Circuit Requirements for Thyristor

In this section, we discuss two aspects of gate control: (1) gate pulse level and (2) gate pulse duration.

2.4.2.9.1 GATE PULSE LEVEL

The threshold gate current pulse amplitude needed to fire (that is, turn ON) a thyristor is stated in the data sheet of the device. It depends on the thyristor voltage and current ratings, and is typically in the range of about fifty to several hundred milliamperes for medium and large size devices. Since the gate cathode junction is being forward-biased by the gate pulse, the gate-to-cathode voltage level is small, well below 5 V, typically around 1.2 V. In practice, the gate current pulse amplitude actually used for firing a thyristor is considerably higher than the threshold value, and its rise time is as short as possible, to ensure fast turn ON switching.

The reverse voltage capability of the gate cathode junction is very low. It is therefore important to carefully examine the gate control circuit and ensure that there is no possibility of a large reverse voltage occurring, even under any abnormal circuit conditions. An expensive high power thyristor can be permanently damaged by the application of a reverse gate voltage, which may be low, but still above the breakdown limit.

2.4.2.9.2 GATE PULSE DURATION

Gate firing of a thyristor will be successful only if the gate pulse lasts at least till the thyristor current rises to the latching level. Looking at this requirement solely from the point of view of the thyristor, the minimum time needed is small, typically in the neighborhood of $10\mu s$ for a large power device. However, in practical converters, there are external power circuit conditions that will necessitate the use of much wider pulses. This could be one or both of the following.

1. External circuit conditions can cause a delay in the current rise. A situation of this kind was considered where the presence of a large inductance in the power circuit delayed the rise of the current to the latching level, necessitating a wide gate pulse.
2. There can be a delay in the forward-biasing of the thyristor.

A gate firing pulse can be effective only if the thyristor is forward-biased. In some converter circuits, the exact instant in the converter switching cycle when a thyristor becomes forward-biased varies according to load conditions, and it becomes very difficult to design the gate firing circuit to provide a properly timed narrow pulse. It is more convenient to provide a wide pulse that commences at the earliest instant at which the thyristor may turn ON. In this way, even if the thyristor is not forward-biased when the gate pulse commences, the gate pulse will still be present when it does become forward-biased and is ready to turn ON.

2.4.2.10 Timing Control and Firing of Thyristors

The gate control circuit of a thyristor converter is designed to provide the gate firing pulses to the thyristors at the appropriate instants of time. The switching control unit of a power electronic converter generally has a timing circuit in it.

This timing circuit generates the timing pulses at the correct instants of time at which each power device has to be switched. The timing circuits for different types of converters will be described when we consider the converters themselves in subsequent chapters. The timing circuit for a switching element normally provides a pulse whose duration is the ON time of the device. In general, such a pulse from the timing circuit is unsuitable for being directly applied to the control terminal of the switching device for the following reasons.

1. A latching device like a thyristor only needs a pulse of short duration, and not for the entire duration of the ON time. Therefore unnecessary gate power dissipation can be avoided by using a very short pulse that starts at the rising edge of the timing pulse.
2. It may need further amplification to be able to provide the required current and power to successfully turn ON the power switching device.
3. It is invariably necessary to provide electrical isolation between the switching control circuit and the power circuit of the converter. The power switching elements will generally be working at high and variable potentials, from which the control circuits will have to be isolated.

In the case of a thyristor, the first of the above three requirements can be met by using a monostable mulitivibrator chip that is triggered by the rising edge of the timing pulse. The multivibrator chip can be programmed to output a pulse of the required width for the thyristor.

To raise the power capability of a voltage pulse, we can use a "driver." Drivers are available as integrated circuits. They reproduce the input pulse at the output terminals with greatly increased current capability. Alternatively, we can use discrete elements as shown in Fig. 2.18 for pulse amplification.

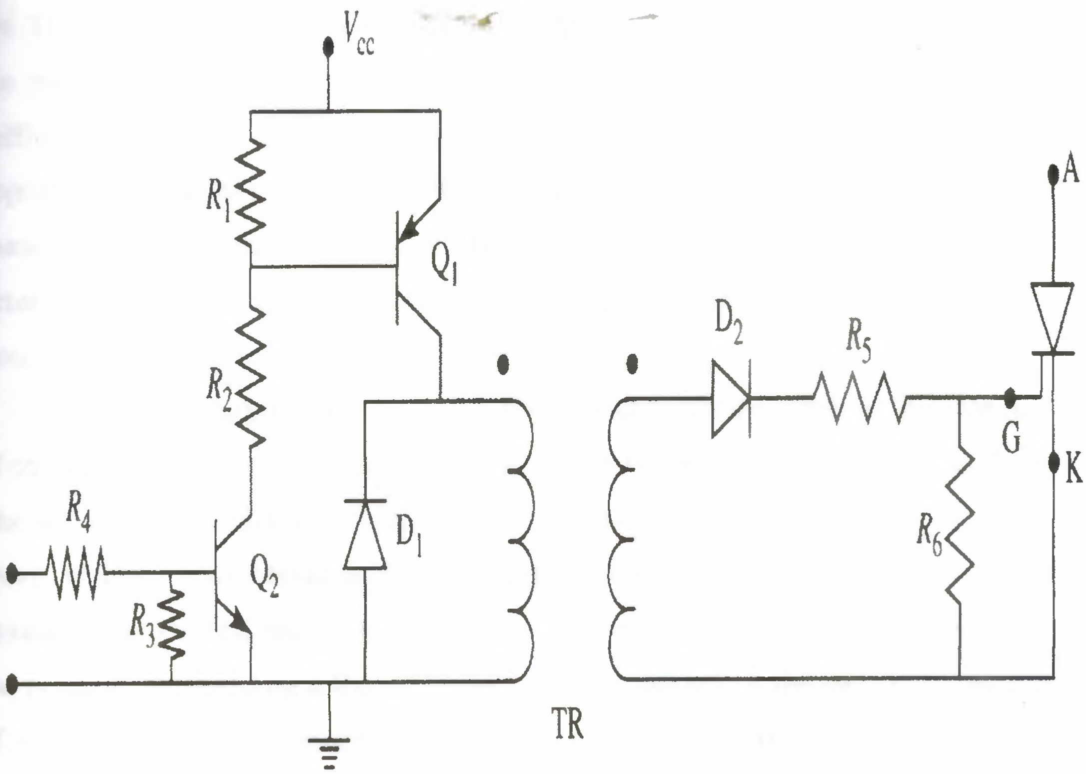


Figure 2.18 Gate firing circuit for a thyristor

A pulse transformer can be used to provide electrical isolation. But careful design of the pulse transformer is usually needed to faithfully reproduce the input pulse at its output terminals. Opto-isolators are also used for electrical isolation. An opto-isolator chip consists of a light emitting diode (LED) that throws light on to a photodiode or phototransistor, causing it to conduct. If an opto-isolator is used for electrical isolation, it is normally required to have a floating (isolated) power source on the gate side, for pulse amplification, because it is impossible to transmit any appreciable power through the opto-isolator. A gate firing circuit for a thyristor using discrete transistors for pulse amplification and a pulse transformer for isolation is shown in Fig. 2.18. This circuit assumes that the input pulse available is of the required width. Therefore the monostable multivibrator chip is not included.

In this arrangement, the turn ON pulse to be amplified is applied to the base of the transistor Q_2, causing it to turn ON. R_1 and R_3 are high resistances to improve performance by providing ohmic paths between the respective base and emitter. These may be ignored for the purpose of our present description. When Q_2 turns ON, it provides an amplified base drive current for the pnp transistor Q_1 and causes it to turn

ON. The value of the base drive current for Q_1 is decided by the value chosen for R_2 and the power supply voltage, labeled V_{cc} in the figure, provided that the input to Q_2 is sufficient to turn it fully ON. When Q_1 turns ON, it applies a pulse of amplitude approximately equal to V_{cc} to the primary of the pulse transformer. This pulse has the same duration as the input pulse. But it has a large current capability, which is determined by the base current provided to Q_1. In this circuit, the transistors are functioning as switches.

The secondary coil of the pulse transformer feeds the pulse to the gate terminal of the thyristor. The resistances R_5 and R_6 serve to limit the gate voltage and current. The diode D_2 prevents any reverse voltage on the gate from the transformer. R_6 also serves to provide an ohmic path from the gate to the cathode of the thyristor. This is a desirable feature, and makes the gate less sensitive to stray voltages. The diode D_1 is for the purpose of providing a freewheeling path through which the current in the primary of the transformer will freewheel when Q_1 turns OFF. This will avoid an excessive voltage occurring across this transistor due to the transformer inductance. Careful design of the pulse transformer is essential for the satisfactory working of this circuit. It has to have the required pulse width capability. Also, the unidirectional pulses in the primary will have a DC component that can lead to magnetic saturation of the core.

2.4.2.11 Thyristor Ratings and Protection

Unlike the power diode, the thyristor has two categories of voltage ratings— the forward blocking and the reverse blocking. These ratings are generally equal.

The current ratings are specified in a manner similar to what we described for power diodes. As in the case of all power semiconductor devices, current ratings are generally valid only if cooling is provided by the use of suitable heat sinks, to keep the device temperature within the specified limit. As in the case of power diodes, average, r.m.s., repetitive peak as well as surge current ratings are usually specified separately, especially for devices with large ratings. Power semiconductor category fuses may be used with each individual thyristor in a converter, to provide short-circuit protection to it. The main parameter on the basis of which the fuse is selected is the I^2t rating. To afford protection, the I^2t rating of the fuse must be less than the I^2t rating of the thyristor. Its continuous current rating should match that of the thyristor. The voltage rating of the selected fuse will be based on the maximum circuit voltage that can occur across it, in the event of a fuse blow.

Other major specifications of the thyristor, which are directly related to its safety, from the power circuit side, are the *dv/dt* and the *di/dt* ratings. These are expressed in terms of V / μs and A / μs respectively. A *dv/dt* failure causes an erratic turn ON of the device, and, depending on the circuit conditions, this may result in a short circuit in the system, with the possibility of damage to the device and other components. Exceeding the *di/dt* limit can directly damage the device because of excessive local current concentration in an area of the thyristor pellet.

Switching aid circuits, also called snubber circuits, are used for the purpose of limiting the stress on static semiconductor switching devices during switching transitions A snubber circuit is invariably used to protect a thyristor from excessive stresses. Figure 2.19(a) shows the commonly used *R-C* snubber. This snubber circuit functions in a similar way as for a power diode, by limiting the over-voltage resulting from the reverse recovery current transient. Besides, by suitable choice of values for *R* and *C*, it can also serve to mitigate the *dv/dt* stress. But the snubber can be made more effective in limiting *dv/dt* by the addition of a diode as shown in Fig. 2.19(b).

To limit *di/dt*, when circuit conditions are such that there is a danger of exceeding the specification, an inductance may be added as shown in Fig. 2.19(c). This is usually a coil of a few turns, capable of carrying the full thyristor current. In some designs, this coil is wound on a small ferrite ring. This reduces the size of the coil for the same inductance value. Besides, the core gets saturated when the current exceeds its saturating level, and so the effective inductance falls to a low value from then.

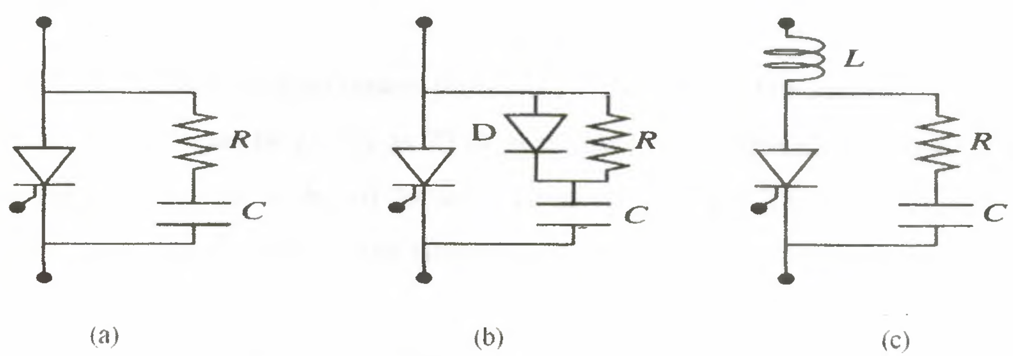


Figure 2.19 Snubber circuits for thyristors

2.4.3 POWER BIPOLAR JUNCTION TRANSISTORS AND POWER DARLINGTONS

2.4.3.1 Types and Ratings

These are high power versions of conventional small signal junction transistors, which are widely used for signal processing, both as discrete devices and in integrated circuits. High power discrete devices with individual current ratings of several hundred amperes and voltage ratings of several hundred volts are presently available, and such devices are widely used as static switches in power electronic converters. Power devices are available both in the *npn* and the *pnp* format. But the available current and voltage ratings are higher for *npn* devices. Because of the greater mobility of electrons compared with holes, *npn* devices can be fabricated on a smaller chip area to provide the same performance as an equivalent *pnp* device.

In static power converters, junction power transistors are invariably used as switches. These switches do not have any significant ability to block reverse voltages, and should be used in such a way that they are only required to block forward voltages. They are current controlled devices. This means that the operation of the switch is specified by the current input at its control terminal. There is a minimum threshold current input I_B necessary at the control terminal (which is labeled "base" in a junction transistor), for a given ON state current I_C through the main terminal (labeled the "collector terminal"). This minimum requirement to ensure the proper ON state is specified by the parameter h_{FE} for these devices, which is defined as

$$h_{FE} = I_C / I_B$$

The h_{FE} values for high power transistors are relatively low compared with low power devices, and may be as low as 20 or even less. This means that to switch 200 A using a transistor that has an h_{FE} of 20, we shall need to input at least 200 / 20 = 10 A at its base terminal. This is still a large current requirement from a control circuit point of view.

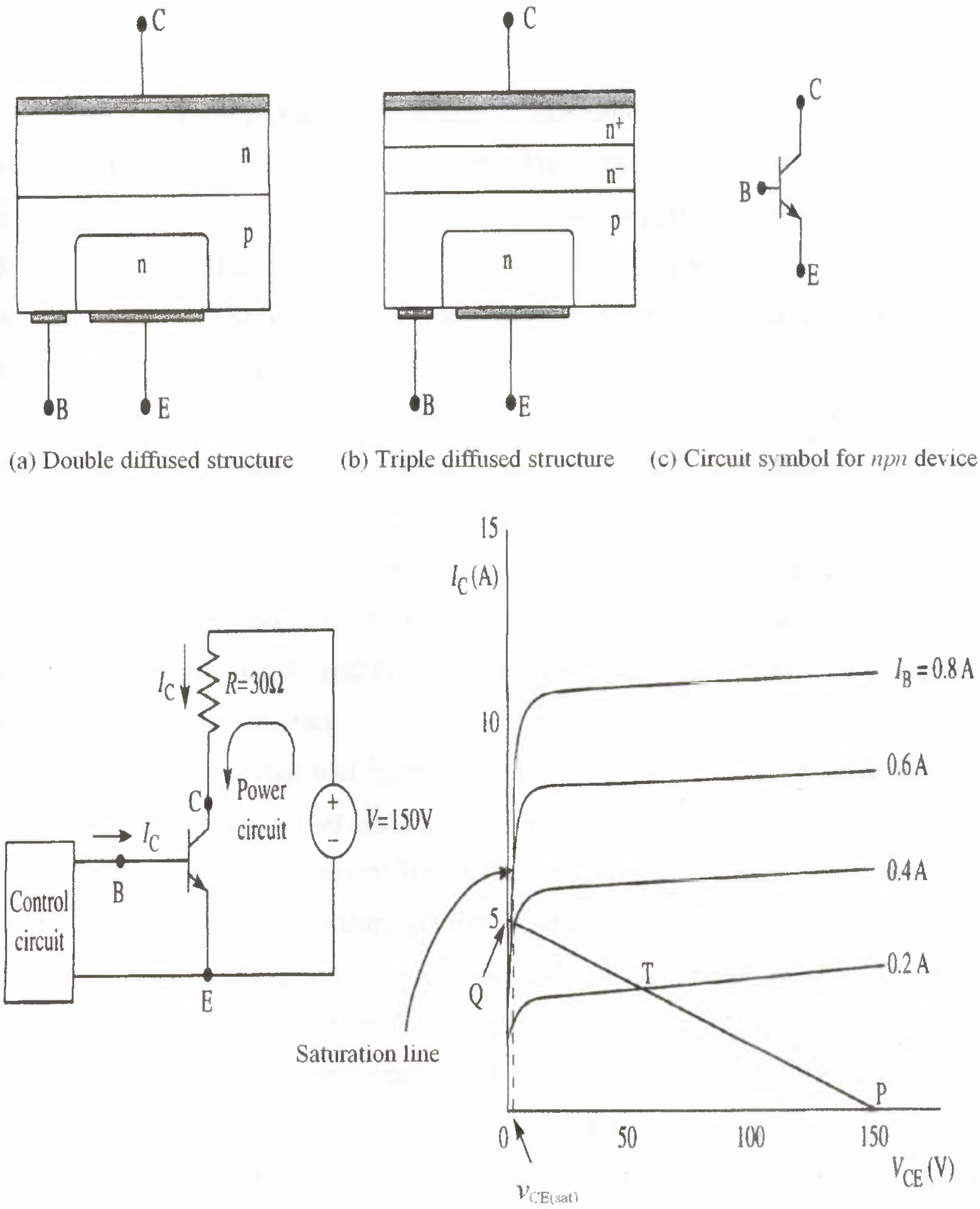
This difficulty can be alleviated by using the "Darlington" arrangement. This scheme employs two transistors, one of which is the main transistor, the other being a smaller one. They are interconnected, in the manner to be explained later, so that the smaller "drive" transistor provides the base current to the main transistor. To turn the Darlington switch ON, it is only necessary to provide a very much smaller input at the base of the drive transistor, to enable it to provide a higher base current to the main transistor. A Darlington pair can be assembled from suitably chosen discrete transistors. However, integrated power Darlingtons are presently manufactured and are widely used. In these, the two interconnected transistors are fabricated on the same silicon pellet.

2.4.3.2 Junction Structure, Static Characteristics

Figure 2.20(a) shows the junction structure of a double diffused *npn* power transistor. In the fabrication of this device, the starting material is an *n* type silicon wafer. First a *p* layer is formed, by diffusion of impurities, on one side. This is the base layer. A second diffusion, after masking the base terminal area, creates an *n* zone, which is the emitter layer. Electrical contacts are made by forming layers of metal, by vapor deposition, which are indicated by thickened lines in Fig. 2.20(a). The three external terminals of the device, which are collector (C), base (B) and emitter (E), make contact with the appropriate zones in the pellet through these metal deposition areas, as shown in Fig. 2.20(a). The circuit symbol for a power transistor is the same as for a low power device, and is shown in Fig. 2.20(c) for an *npn* type. For a *pnp*, the arrow should be reversed. Figure 2.20(b) shows the "triple diffused" construction employed in some high power transistors. In this, an additional low resistivity *n* region labeled the n^+ layer is formed on top of the collector layer by means of a third diffusion. This is done primarily to provide a low resistance ohmic contact between the collector and the collector metal layer with good mechanical properties. It does not create any fundamental difference in the operating principle of the transistor.

When a transistor is used as a controlled switch, the control current input is provided at the base terminal. The control circuit is connected between the base and emitter. The power terminals of the switch are the collector and the emitter. This is shown in Fig. 2.20(d). In this circuit, by sending an appropriate current into the base terminal the "load" resistance R is connected across the DC supply voltage V. The manner in which the switching is achieved may be explained with reference to the "output" characteristics of the transistor shown in Fig. 2.20(e). The output characteristic is a plot of the current I_C through the switch versus the voltage V_{CE} across it for a fixed value of the control terminal current I_B. In Fig. 2.20(e), we have a family of output characteristics for different discrete values of I_B.

OFF state. If I_B is made zero, the value of I_C is negligibly small. This is the OFF state of the switch. This is also called the cut-off condition of the transistor. For negative values of V_{BE} also, there is no base current, and the transistor remains OFF. But the reverse voltage capability of the base-emitter junction is quite small, and it is important to ensure that this is not exceeded.



(d) Transistpr power switch (e) Output characteristics

Figure 2.20 *npn* bipolar junction power transistor

ON state. Let us assume that in Fig. 2.20(d), $V = 150$V. Let $R = 300\ \Omega$ initially. Let us also decide to keep I_B at 0-6 A. The voltage V_{CE} across the switch and the current I_B through it must be given by a point on the characteristic for $I_B = 0.6$ A. To locate this point, we used a second relationship resulting from the application of Kirchhoff's law to the power circuit loop, which gives

$$V_{CE} = V - I_C R$$

This relationship is given by a straight line called the "load line." The load line corresponding to R=30 Ω is shown as PQ in Fig.2.20(e). It is drawn by choosing two points on it that satisfy the above equation. These are usually chosen as the one for $I_B = 0$ for which $V_{CE} = V$ (point P) and the other for $V_{CE} = 0$ for which $I_C = V / R$ (point Q). The intersection of the output characteristic and the load line gives us the current through the switch and the voltage across it.

We notice that the characteristics for all the different values of I_B are overlapping on the left-hand side, along a near-vertical line, indicated in the figure as the "saturation" line. The load resistance of 30 Ω gives us a load line that intersects the $I_B = 0.6$ A characteristic in the saturation region. The voltage across the transistor is seen to be very small, and is denoted by the symbol $V_{CE(sat)}$ in the figure. This is the ON state of the transistor switch, and $V_{CE(sat)}$ is the unavoidable small forward voltage drop across the switch in its ON state.

Let us now see what will happen if we progressively reduce the base current I_B. We notice from the figure that until the base current falls to about 0.4 A, there is no noticeable fall in the load current. But further reduction in I_B causes the intersection point to move away from the saturation line, resulting in a large increase in the voltage across the transistor and a corresponding decrease in the voltage across the load. For example, for $I_B = 0.2$ A, the intersection point (labeled T in the figure) gives $v_{CE} = 60$ V and a current of 3 A. The transistor is no longer in the saturated ON state. Such a condition is to be avoided, because clearly the switch is very far from perfect. More importantly, there will be excessive power dissipation in the transistor, which can result in its damage. In our example, the collector power dissipation is seen to be 60 V X 3 A = 180 W. Compared with this, the power dissipation in the saturated ON state is very small. From the figure, the current is seen to be approximately 5 A and the voltage $V_{CE(sat)}$ approximately 2.5 V, resulting in a power dissipation of about 12.5 W.

The above consideration highlights the need to ensure a saturated ON state, by providing adequate base drive current, for the safe and satisfactory operation of the switch. To determine the minimum base current to ensure the saturated ON state, we use a parameter specified as h_{FE} in the data sheet of the transistor:

$$h_{FE} = I_C + I_B$$

From this, the minimum base current needed to ensure a saturated ON state will be given by $I_B = I_C / h_{FE}$. For values of the base current higher than this, the transistor will be saturated. Often it will be advisable to use a somewhat higher value of base current than that indicated by the above formula, as a safety feature, to take care of possible increases in I_C above the anticipated value. When this is done, we say that the circuit is designed to work normally with a "forced h_{FE}" that is less than the specified h_{FE} of the transistor.

Proportional drive. The fact that the minimum base current drive needed to ensure the saturated ON state of the transistor switch depends on the ON state current. In practical converters, the ON state current through the switch may vary according to load conditions. Therefore, if we employ a fixed base current drive, this should be sufficient for the highest ON state current to be expected. This implies that the base will be over-driven whenever the ON state current is less than the maximum value.

A major disadvantage of over-driving the base is the increase in the transition time for turn OFF switching. This happens because excessive base current will cause excessive injection of minority carriers into the base region of the transistor, from the emitter side. Because of this, the collector current will persist for a longer time, until the excess minority carriers are removed, during turn OFF switching. To overcome this difficulty, circuit designers some times use "proportional drive." In such a scheme, the base current is automatically increased or decreased according to the magnitude of the collector current.

2.4.3.3 Safe Operating Area (SOA)

When a transistor functions in an electrical circuit, we can define its "operating point" at any given instant of time by means of the voltage V_{CE} across it and the current I_C through it. The operating point so defined can be graphically located by a point on the

I_C versus VCE plane such as in Fig. 2.20(e). Whenever there is a change of V_{CE} or I_C, or both, the operating point moves to a different location on this plane. The transition will be along a curve on the I_C versus V_{CE} plane, whose path will be determined by the instantaneous values of I_C and V_{CE} during the change. To ensure safe operation of the transistor without damage to it, all the operating points should be within finite boundaries on the I_C versus V_{CE} plane during transitions between operating points, which may occur during switching or for other reasons. This is called the Safe Operating Area (SOA). The boundaries of the SOA are usually specified by the manufacturer of the device, for stated conditions of working. Figure 2.21(a) shows a typical safe operating area. We shall examine the nature of the SOA and the parameters that determine each of the boundary lines.

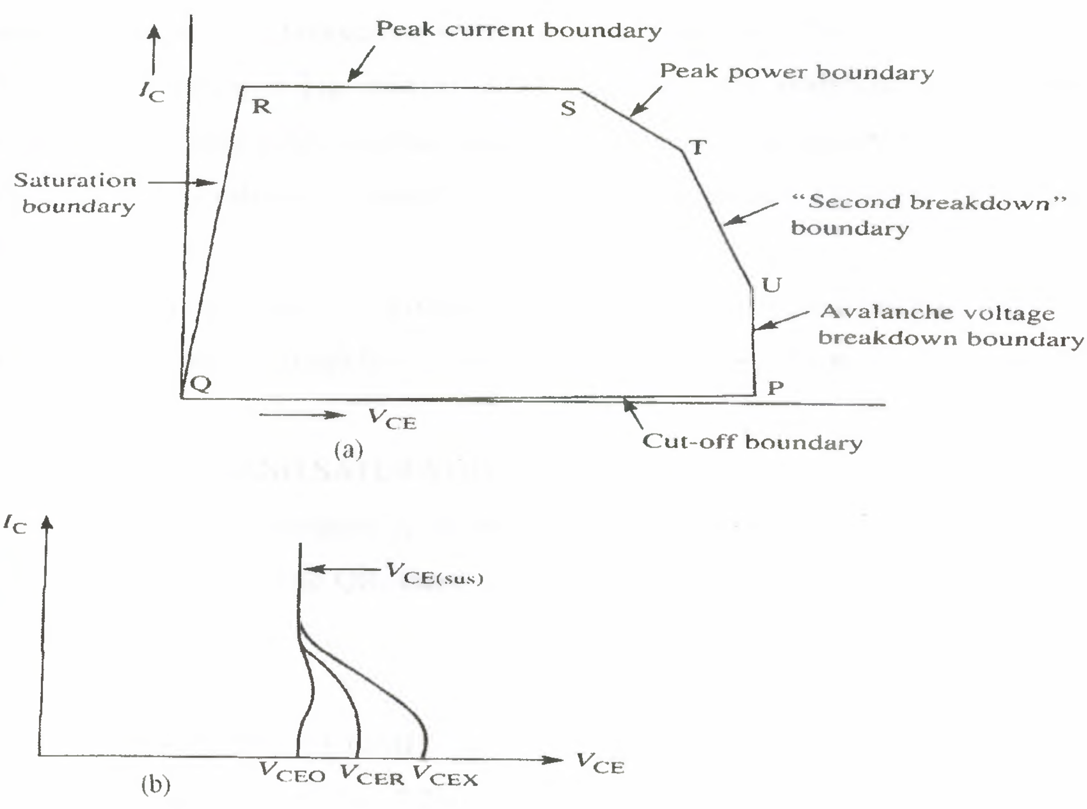


Figure 2.21 (a) Reverse-biased safe operating are for a junction power transistor (b) Avalanche breakdown limits

2.4.3.3.1 MAXIMUM VOLTAGE - AVAILABLE BREAKDOWN LIMIT

A transistor has a maximum collector-to-emitter voltage V_{CE} that it can withstand, above which avalanche breakdown at the collector junction will occur. This determines the maximum voltage limit P in the SOA in Fig. 2.21 and the vertical boundary line PU. An indication of the maximum voltage capability is also provided in the data sheet of the transistor, by a parameter labeled as the "sustaining voltage" ($V_{CE(SUS)}$). The significance of $V_{CE(SUS)}$ may be explained as follows.

If the voltage V_{CE} across the transistor is increased progressively, the voltage at which avalanche breakdown commences will be determined by the manner in which the base terminal is connected. If the base terminal is left open (which means that the external impedance between the base and emitter is infinite), breakdown commences at a lower value of V_{CE} in comparison with the case where the base terminal is shorted to the emitter (zero external impedance between base and emitter). The breakdown voltage for the case of base terminal open is labeled V_{CEO} and that for the case of base terminal shorted to emitter is labeled V_{CEX}. For an intermediate value R of external base-to-emitter impedance, the breakdown voltage will be V_{CER}, intermediate between V_{CEO} and V_{CEX}. This is shown in Fig. 2.21(b). In all cases, after the avalanche breakdown has occurred, the voltage tends to remain constant at or near a particular value, as shown in Fig. 2.21(b). This value is the specified $V_{CE(SUS)}$ of the transistor. $V_{CE(SUS)}$ is usually about the same as V_{CEO}.

The value of $V_{CE(SUS)}$ sets one of the operating limits of the transistor switch. The following example illustrates how a large V_{CE} can occur during turn OFF switching.

2.4.3.3.2 CUT OFF AND SATURATION BOUNDARIES

Since normal operation is above the cut-off line PQ in Fig. 2.21(a) and to the right of the saturation line QR, these two lines constitute two other boundaries of the SOA.

2.4.3.3.3 PEAK CURRENT LIMIT

The lines RS in Fig. 2.21(a) corresponding to the maximum permissible collector current constitutes another boundary of the SOA.

2.4.3.3.4 MAXIMUM POWER

Neglecting the small base power dissipation, the total power dissipation in the transistor is equal to the collector power dissipation given by $p_{max} = V_{CE} I_C$. The maximum permissible value $p_{max} = V_{CE} I_C$ constitutes the boundary of the SOA indicated as ST in Fig. 2.21(a).

When used as a static switch, the peak current boundary RS and the maximum power boundary ST are very important during the turn OFF switching transition. The turn OFF switching of a power transistor is usually achieved by applying a reverse voltage to the base terminal. This reverse voltage is for the purpose of speeding up the turn OFF transition by "sucking out" the excess minority carriers from the base region. Therefore, this reverse voltage should last at least till the turn OFF switching transition is completed. So, in practice, these two boundaries of the SOA and the last boundary, which we shall consider next, are applicable under conditions of reverse-biased base voltage.

2.4.3.3.5 SECOND BREAKDOWN

In addition to the five boundaries of the SOA already described, there is another one, shown as TU in Fig. 2.21(a) and labeled "second breakdown." This is a phenomenon that can occur in a junction power transistor when voltage current and power dissipation are high, but still below the levels indicated by the limits discussed earlier. If we assume that during the turn OFF switching transition the collector current is uniformly distributed over the collector junction area, the power distribution will also be uniform over this area. If, on the other hand, the current distribution is nonuniform, local hot spots can occur due to excessive power dissipation, in locations in the junction area that experience high current densities. A failure of the device due to such occurrence of local hot spots is described as "second breakdown." On an oscillogram, the failure of a transistor can be identified as due to second breakdown by the collapse of V_{CE} to a low value. In the case of avalanche breakdown, the voltage across the device does not collapse, but stays at the $V_{CE(SUS)}$ level. While avalanche breakdown need not necessarily result in permanent damage, second breakdown always does.

2.4.3.4 Switching times

Typical waveforms of collector current during turn ON and turn OFF transitions are shown in Fig. 2.22. The instants of time marked therein have the following significance:

t_0 the instant at which the turn ON switching is initiated by the arrival of the base current pulse

t_1 the instant at which the collector current has risen to 10% of its final value

t_2 the instant at which I_C reaches 90% of its final value

t_3 the instant at which turn OFF switching is initiated

(This is typically done by the application of a small reverse voltage on the base, to speed up the transition, resulting in a reverse base current pulse of short duration due to the excess minority carriers in the base region. This reverse base current lasts until the excess carriers are swept out of the base region. Notice that the collector current continues without any significant decrease, for a short time after t_3, because of these stored excess minority carriers.)

t_4 the instant at which the collector current has fallen to 90% of its ON state value

t_5 the instant at which the collector current has fallen to 10% of its ON value

The time delays stated in a typical data sheet of the device are defined as follows:

t_r "rise time" = $t_2 - t_1$

t_s "storage time" = $t_4 - t_3$

t_f "fall time" = $t_5 - t_4$

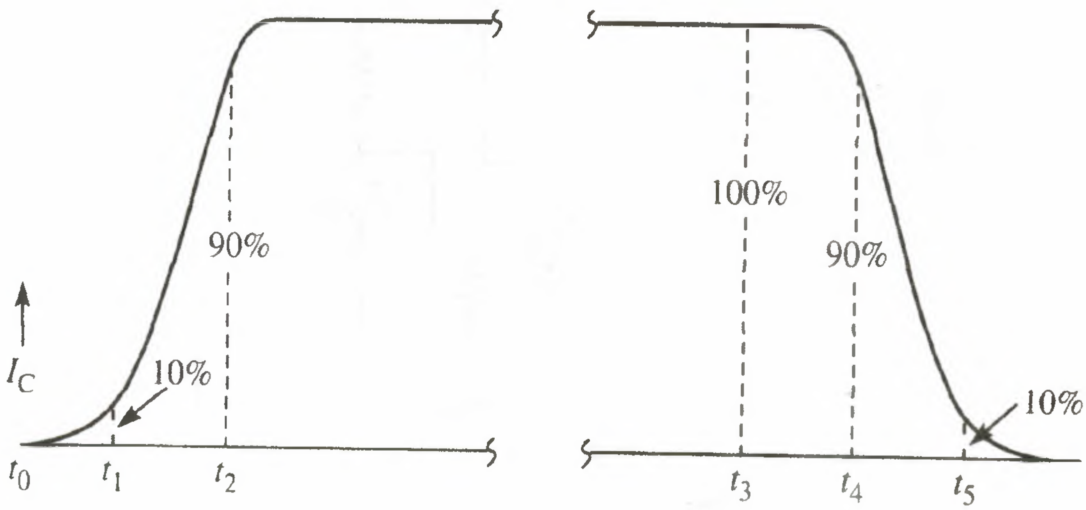


Figure 2.22 Power transistor switching times

2.4.3.5 Base Drive Circuits for Power Transistor Switches

A well-designed base drive circuit should provide adequate base current to guarantee a saturated ON state under all conditions of collector current that can occur during operation. Also, a fast rising base current waveform will ensure fast turn ON switching. During turn OFF switching, a reverse base current of sufficient amplitude will result in the reduction of the storage time and therefore faster switching. For this purpose, it will be necessary to apply a reverse voltage pulse. It is not necessary to maintain this reverse voltage after the reverse current has fallen to zero, when the excess minority carriers in the base zone have been removed. It is important to remember that the reverse voltage capability of the base emitter function is very small, and any reverse voltage should be well below the rated value to avoid damage to the transistor.

An example of a base drive circuit for a power transistor switch is given in Fig. 2.23. The base is driven from the secondary coil C_3 of a three-winding transformer. The transformer has two primary coils labeled C_1 and C_2. To implement the turn ON switching, a positive voltage is applied at the terminal labeled A in Fig. 2.23.

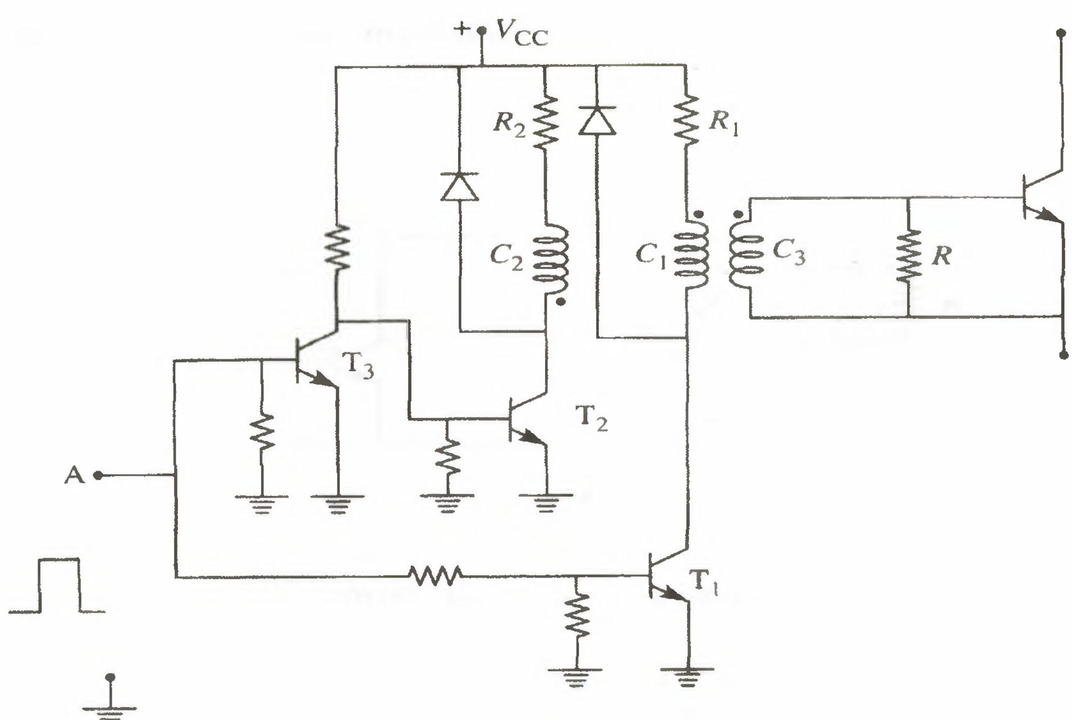


Figure 2.23 Example of a base drive circuit for a junction power transistor

This causes the transistor T_1 to turn ON, energizing the primary coil C_1. At this time, the primary C_2 is turned OFF by transistor T_2. This happens because T_3 turns ON and connects the base of T_2 to ground. When the voltage pulse at A goes to the zero-voltage level, both T_1 and T_3 turn OFF. When T_1 turns OFF, the primary coil C_1 is disconnected from the power supply. The turning OFF T_3 causes the primary coil C_2 to be energized by the turning ON of T_2. Therefore a negative voltage appears on the secondary coil C_3. This negative voltage should be well below the rated reverse voltage of the base emitter junction. In this circuit, the reverse voltage on the base terminal may be maintained during the entire OFF period of the transistor depending on the pulse width capability of the pulse transformer. The pulse transformer provides isolation between the switching control circuit and the power circuit of the transistor. This pulse transformer must have the necessary pulse width capability to ensure the continued presence of the output pulse, during the entire duration of the ON time of the transistor.

2.4.3.6 Switching Aid Circuits (Snubber Circuits)

A typical switching aid circuit used with a junction power transistor is shown in Fig. 2.24. This circuit is for the purpose of limiting the operating point within the safe operating area (SOA) during turn OFF switching.

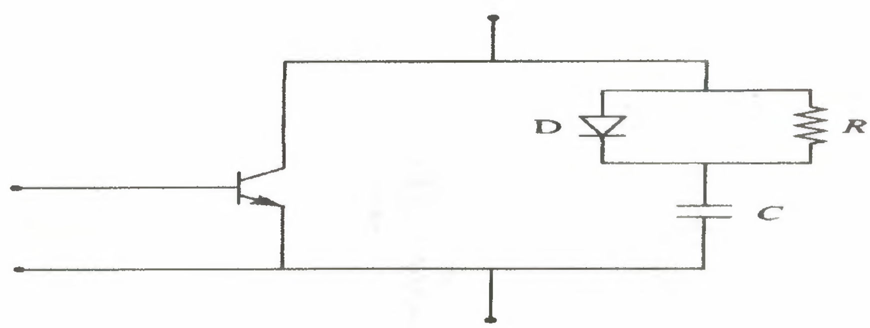


Figure 2.24 Turn OFF snubber circuit for junction transistor

The circuit consists of a capacitor C, a diode D and a resistor R. When the transistor is in the ON state, the voltage across it, and therefore across the switching aid circuit, is nearly zero. The purpose of the capacitor-diode combination is to slow down the rate of rise of voltage across the switch during the turn OFF switching transition. This happens because during this time the diode turns ON and the capacitor starts charging. In the OFF state of the transistor, the capacitor remains charged to the full blocking voltage. It discharges during the next turn ON switching of the transistor. The resistor R is for the purpose of limiting the peak value of the discharge current through the transistor. Each time the transistor is turned ON, the total energy stored in the capacitor is dissipated in the resistor. Therefore the power dissipation in R is proportional to the switching frequency and to the square of the blocking voltage.

2.4.3.6.1 TURN ON SWITCHING.

Turn ON switching into a highly inductive load does not normally create a power dissipation problem. Referring to Fig. 2.25, during the turn ON transition, the voltage v_{CE} is given by

$$v_{CE} = V - (L\,(di\,/\,dt\,) + R\,i\,)$$

The inductance slows down the rate of rise of current through the transistor during the transition time. For this reason, the voltage transition from V to zero will be completed well before the current rises to significant levels. Therefore the internal power dissipation during the transition will be small. For such a condition of working a "turn ON" snubber may not be necessary.

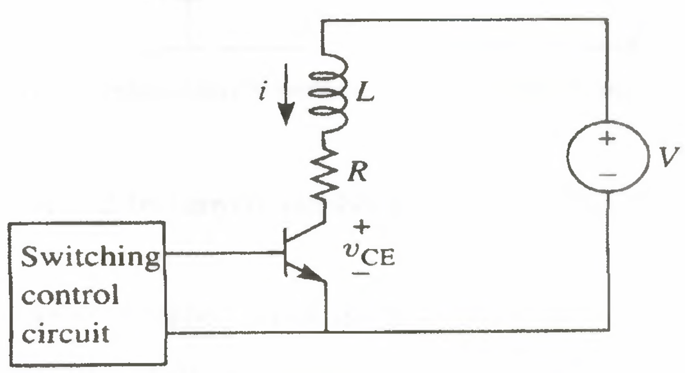
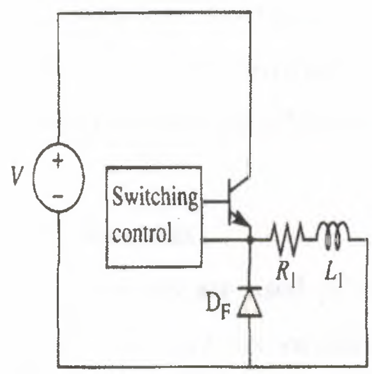


Figure 2.25 Turn OFF switching of an inductive load

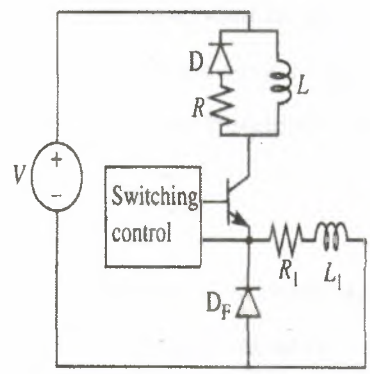
There are, however, other situations, which occur very often in practice, in which a turn ON snubber will be necessary. This is illustrated by the circuit shown in

Fig. 2.26.

In the circuit of Fig. 2.26(a), the transistor switch is turned ON and turned OFF at a chosen repetitive frequency. This is for the purpose of controlling the current in the "$R_1 - L_1$" load circuit, which can be done by controlling the ON time in the switching period. During the ON time of the switch, the voltage V is applied to the load, and during the OFF time, the current freewheels (continues to flow due to the stored energy in L_1) through the diode D_F. Let us focus attention on the instant at which the transistor is turned ON, when the current is already freewheeling through D_F. Because D_F has a finite recovery time, initially, when the transistor switch is turned ON, the full voltage V comes across it, without any circuit element in the loop to limit the current. This is therefore a situation in which the voltage and current magnitudes can take the device outside its SOA and cause damage to it. This type of situation occurs in the majority of static DC/DC converters and DC/AC inverters. Unless there are external circuit elements to limit the growth of current, it will become necessary to use a "turn ON" snubber.



(a) Transistor switch without turn ON snubber



(b) Switch with turn ON snubber

Figure 2.26 Turn ON sunbber for a junction transistor

A typical turn ON snubber circuit is shown with the transistor switch in Fig. 2.26(b). This consists of a small inductance L, *a* diode D and a resistor R. During turn ON switching, the diode D is OFF and the inductance limits the rate of rise of current in the transistor. When the transistor is turned OFF, the current in L freewheels through the diode, and the energy stored in L is dissipated in R. Since L is needed only for a short time initially during the recovery time of D_F, some designers use a saturable core inductance, consisting of a few turns of wire on a toroidal magnetic core. The core gets saturated when the current rises to a certain level, and afterwards the inductance becomes negligible. Therefore such a coil will present the needed inductance at the beginning only of the turn ON switching. In Fig. 2.26(b), only the turn ON section of the snubber circuit is shown, in the interests of clarity. But the total switching aid circuit may actually consist of both the turn ON and the turn OFF snubbers.

2.4.3.6.2 SNUBBER POWER LOSS

The snubber circuits help to protect the transistor switch by limiting its operation within the SOA. But there is a price to be paid for this, in the form of additional power loss. In the turn OFF snubber, the energy stored in the snubber capacitor C and dissipated in each switching cycle is $1/2CV^2$. In the turn ON snubber, the energy stored in the snubber inductance and dissipated in each cycle is $1/2LI^2_{peak}$. Since these energy losses occur in every switching cycle, the average power losses in the snubber circuits increase directly in proportion to the converter switching frequency. Some snubber circuits that incorporate the recovery of stored energy from the snubber circuit elements, instead of dissipating it in resistances, have been developed and are described in the literature. The additional circuit complexity involved in using such schemes may be justified in the interests of improving converter efficiency when the switching frequencies are high.

2.4.3.7 Power Darlingtons

Power Darlingtons are used primarily for the purpose of reducing the control current requirement for turn ON switching. Figure 2.27 shows how the two junction transistors that constitute the Darlington switch are interconnected. T_M is the main power transistor. T_A is the auxiliary transistor, of lower power, which provides the base current to the main transistor. At the present time, integrated power Darlingtons, in which both the transistors and their interconnections are fabricated on the same silicon chip, are available for large ratings. Externally, they have only three terminals, as in single transistors.

Referring to Fig. 2.27, the base current I_{BA} needed to maintain the saturated ON state for the auxiliary transistor with a collector current I_{CA} will be given by the relationship

$$I_{CA} = h_{FE(A)} I_{BA}$$

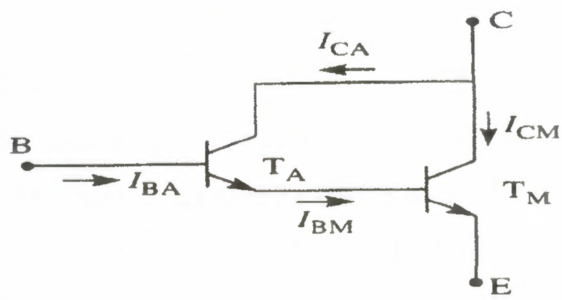


Figure 2.27 Power darlington

The corresponding emitter current, which is the base current of the main transistor, will be

$$I_{BM} = (1 + h_{FE(A)})\, I_{BA}$$

The corresponding collector current I_{CM} will be

$$I_{CM} = h_{FE(M)} I_{BM} = h_{FE(M)}\ (1 + h_{FE(A)})\, I_{BA}$$

Therefore the total load current in the external circuit will be

$$I = I_{CA} + I_{CM} = (h_{FE(A)} + h_{FE(M)} + h_{FE(A)}\, h_{FE(M)}) I_{BA}$$

Since the h_{FE}s are relatively large numbers, their product will be very much larger, and therefore we may approximate the above expression as

$$I = h_{FE(A)}\, h_{FE(M)}\, I_{BA}$$

This means that the Darlington switch has an overall current gain approximately equal to the product of the current gains of the individual transistors constituting the switch. This explains why the Darlington switch needs only a very much smaller control current for its operation, in comparison with a single power transistor, to switch the same load current. However, the switching times of power Darlingtons are somewhat longer than comparable single transistors, as may be expected from the fact that two transistors need to switch in a Darlington.

2.4.3.8 Unijunction Transistor

The unijunction transistor (UJT) is commonly used for generating triggering signals for SCRs. A basic UJT-triggering circuit is shown in Fig. 2.28(a). A UJT has three terminals, called the emitter E, base-one B_1, and base-two B_2. Between B_1 and B_2 the unijunction has the characteristics of an ordinary resistance. This resistance is the interbase resistance R_{BB} and has values in the range 4.7 to 9.1 kΩ. The static characteristics of a UJT are shown in Fig. 2.28(b).

When the dc supply voltage V_s is applied, the capacitor C is charged through resistor R since the emitter circuit of the UJT is in the open state. The time constant of the charging circuit is $\tau_1 = RC$. When the emitter voltage V_E, which is the same as the capacitor voltage v_C, reaches the peak voltage V_P, the UJT turns on and capacitor C will discharge through R_{B1} at a rate determined by the time constant $\tau_2 = R_{B1}C$. τ_2 is much smaller than τ_1. When the emitter voltage V_E decays to the valley point V_v, the emitter ceases to conduct, the UJT turns off, and the charging cycle is repeated. The waveforms of the emitter and triggering voltages are shown in Fig. 2.28(c).

The waveform of the triggering voltage V_{B1} is identical to the discharging current of capacitor C_1. The triggering voltage V_{B1} should be designed to be sufficiently large to turn on the SCR. The period of oscillation, T, is fairly independent of the dc supply voltage V_S, and is given by

$$T = 1/f \approx RC \ln (1/\ 1\text{-}\eta)$$

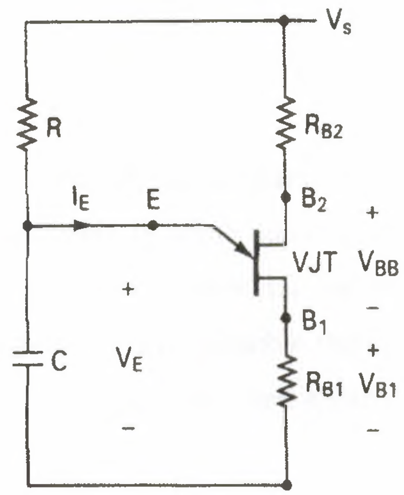
where the parameter η is called the intrinsic *stand-off ratio*. The value of η lies between 0.51 and 0.82.

Resistor R is limited to a value between 3 kΩ and 3 MΩ. The upper limit on R is set by the requirement that the load line formed by R and V_s intersects the device characteristics to the right of the peak point but to the left of the valley point. If the load line fails to pass to the right of the peak point, the UJT will not turn on. This condition will be satisfied if $V_s - I_p R > V_p$. That is,

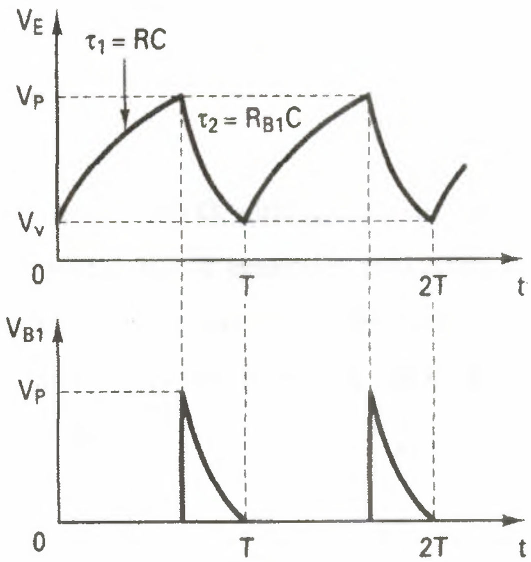
$$R < (V_s - V_p) / I_p$$

At the valley point $I_E = I_V$ and $V_E = V_V$ so that the condition for the lower limit on R to ensure turning off is $V_s - I_v R < V_p$ .That is,

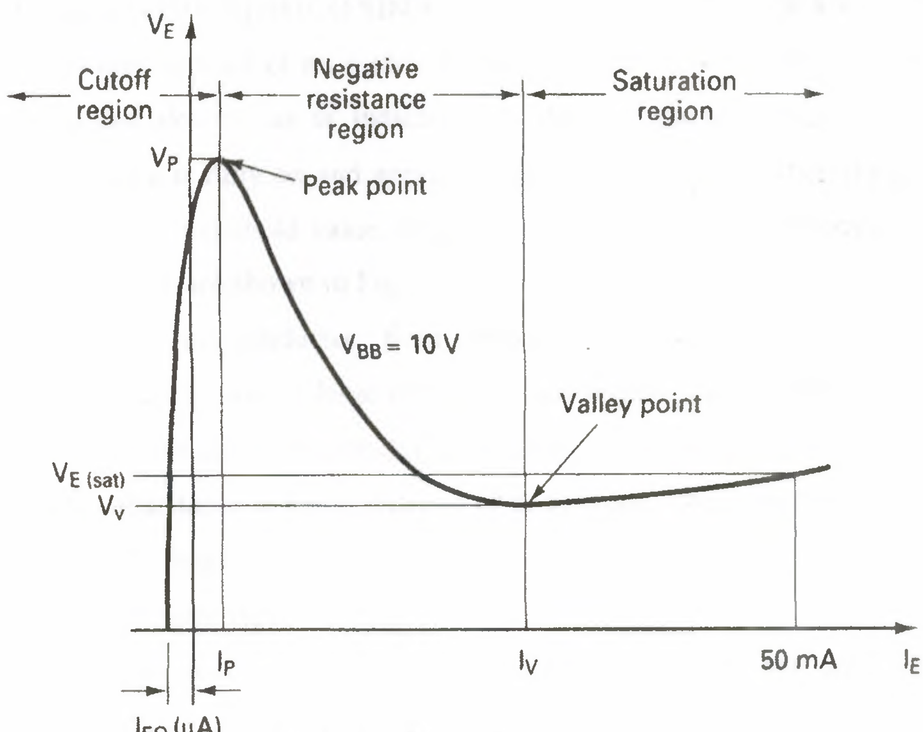
$$R > (V_s - V_v) / I_v$$



(a) Circuit



(c) Waveforms



(b) Static characteristics

Figure 2.28 UJT triggering circuit

The recommended range of supply voltage V_s is from 10 to 35 V. For fixed values of η, the peak voltage V_p will vary with the voltage between the two bases, V_{BB}. V_p is given by

$$V_p = \eta\ V_{BB} + V_D\,(= 0.5\text{V}) \approx \eta\ V_S + V_D\,(= 0.5\text{ V})$$

where V_D is the one-diode forward voltage drop. The width t_g of triggering pulse is

$$t_g = R_{B1}\,C$$

In general, R_{B1} is limited to a value below 100 Ω, although values up to 2 or 3 kΩ are possible in some applications. A resistor R_{B2} is generally connected in series with base-two to compensate for the decrease in V_P due to temperature rise and to protect the UJT from possible thermal runaway. Resistor R_{B1} has a value of 100 Ω or greater and can be determined approximately from

$$R_{B2} = 10^4 / \eta\ V_S$$

2.4.4 METAL-OXIDE-SEMICONDUCTOR FIELD EFFECT TRANSISTORS

The circuit symbol of an *n*-channel MOSFET is shown in Fig. 2.29(a). It is a voltage-controlled device, as is indicated by the *i-v* characteristics shown in Fig. 2.29(b). The device is fully on and approximates a closed switch when the gate-source voltage is below the threshold value, $V_{GS(\text{th})}$. The idealized characteristics of the device operating as a switch are shown in Fig. 2.29(c).

Metal-oxide-semiconductor field effect transistors require the continuous application of a gate-source voltage of appropriate magnitude in order to be in the on state. No gate current flows except during the transitions from on to off or vice versa when the gate capacitance is being charged or discharged. The switching times are very short, being in the range of a few tens of nanoseconds to a few hundred nanoseconds depending on the device type.

The on-state resistance $r_{DS(\text{on})}$ of the MOSFET between the drain and source increases rapidly with the device blocking voltage rating. On a per-unit area basis, the on-state resistance as a function of blocking voltage rating BV_{DSS} can be expressed as

$$r_{DS(\text{on})} = k\,\mathrm{BV}_{DSS}$$

where k is a constant that depends on the device geometry. Because of this, only devices with small voltage ratings are available that have low on-state resistance and hence small conduction losses.

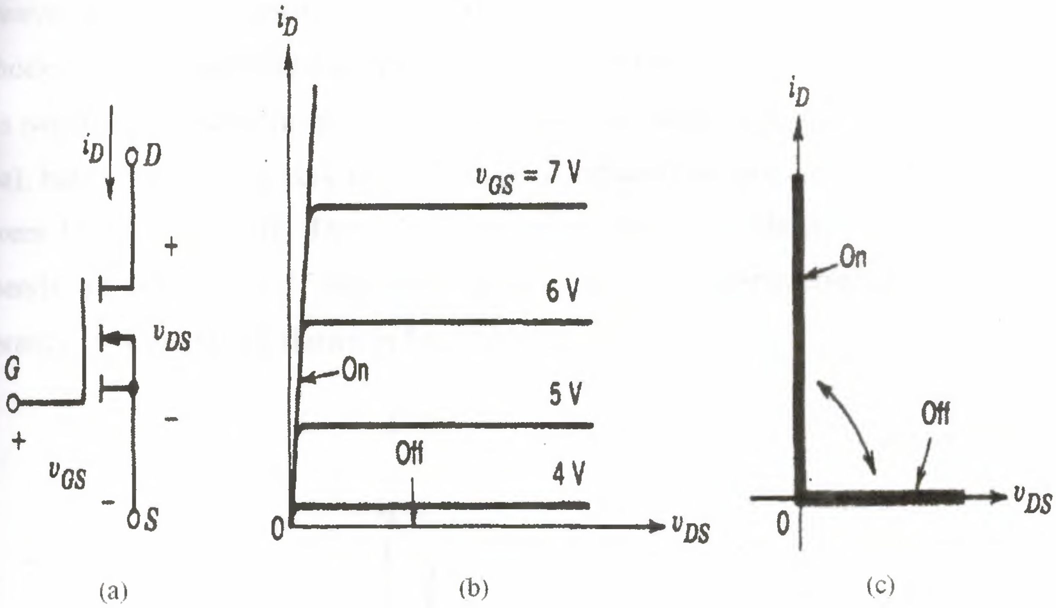


Figure 2.29 *N*-channel MOSFET: (a) symbol ,(b) *i-v* characteristics, (c) idealized characteristics

However, because of their fast switching speed, the switching losses can be small. From a total power loss standpoint, 300-400V MOSFETs compete with bipolar transistors only if the switching frequency is in excess of 30-100 kHz. However, no definite statement can be made about the crossover frequency because it depends on the operating voltages, with low voltages favoring the MOSFET.

Metal-oxide—semiconductor field effect transistors are available in voltage ratings in excess of 1000 V but with small current ratings and with up to 100 A at small voltage ratings. The maximum gate-source voltage is ±20 V, although MOSFETs that can be controlled by 5-V signals are available.

Because their on-state resistance has a positive temperature coefficient, MOSFETs are easily paralleled. This causes the device conducting the higher current to heat up and thus forces it to equitably share its current with the other MOSFETs in parallel.

2.4.5 GATE-TURN-OFF THYRISTORS

The circuit symbol for the GTO is shown in Fig. 2.30(a) and its steady-state *i-v* characteristic is shown in Fig. 2.30(b).

Like the thyristor, the GTO can be turned on by a short-duration gate current pulse, and once in the on-state, the GTO may stay on without any further gate current. However, unlike the thyristor, the GTO can be turned off by applying a negative gate-cathode voltage, therefore causing a sufficiently large negative gate current to flow. This negative gate current need only flow for a few microseconds (during the turn-off time), but it must have a very large magnitude, typically as large as one-third the anode current being turned off. The GTOs can block negative voltages whose magnitude depends on the details of the GTO design. Idealized characteristics of the device operating as a switch are shown in Fig. 2.30(c).

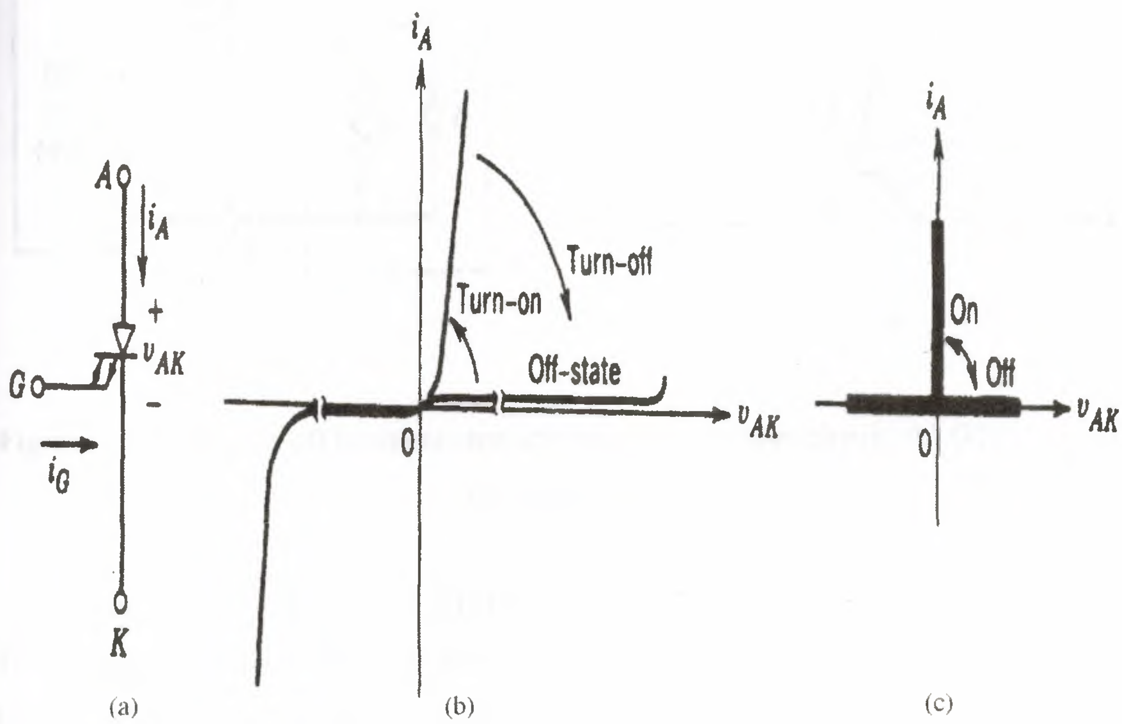
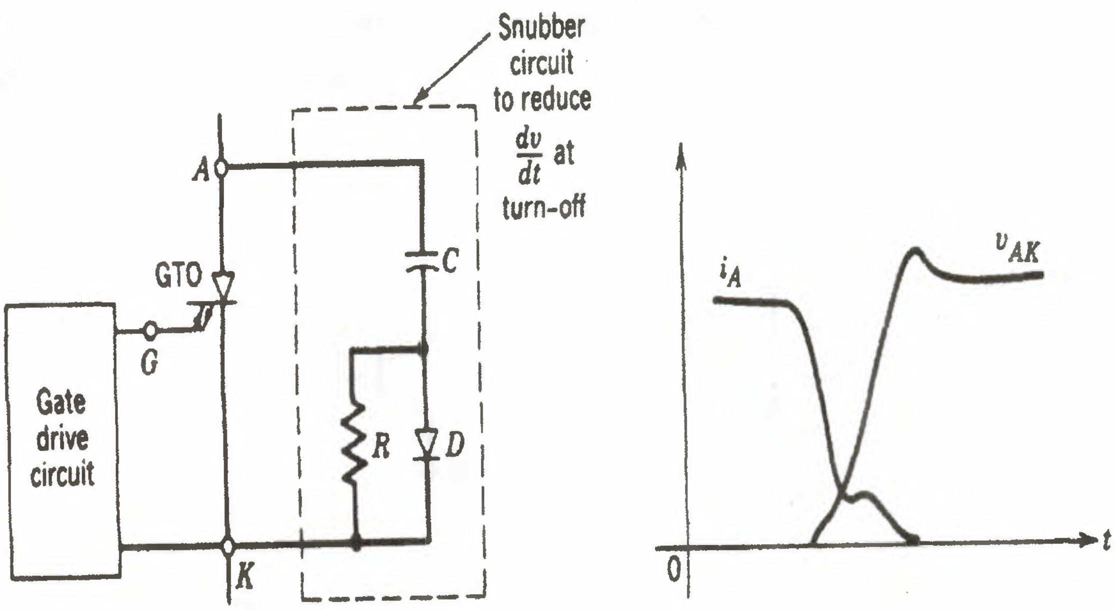


Figure 2.30 A GTO: (a) symbol, (b) i-v characteristics, (c) idealized characteristics

Even though the GTO is a controllable switch in the same category as

MOSFETs and BJTs, its turn-off switching transient is different. This is because presently available GTOs cannot be used for inductive turn-off unless a snubber circuit is connected across the GTO (see Fig. 2.31(a)). This is a consequence of the fact that a large *dv/dt* that accompanies inductive turn-off cannot be tolerated by present-day GTOs. Therefore a circuit to reduce *dv/dt* at turn-off that consists of *R*, *C*, and *D*, as shown in Fig. 2.31(a), must be used across the GTO. The resulting waveforms are shown in Fig. 2.31(b), where *dv/dt* is significantly reduced compared to the *dv/dt* that would result without the turn-off snubber circuit.



(a) (b)

Figure 2.31 Gate turn-off transient characteristics: (a) snubber circuit, (b) GTO turn-off characteristics

The on-state voltage (2- 3 V) of a GTO is slightly higher than those of thyristors. The GTO switching speeds are in the range of a few microseconds to 25 μs. Because of their capability to handle large voltages (up to 4.5 kV) and large currents (up to a few kilo-amperes), the GTO is used when a switch is needed for high voltages and large currents in a switching frequency range of a few hundred hertz to 10 kHz.

2.4.6 INSULATED GATE BIPOLAR TRANSISTORS

The circuit symbol for an IGBT is shown in Fig. 2.32(a) and its *i-v* characteristics are shown in Fig. 2.32(b). The IGBTs have some of the advantages of the MOSFET, the BJT, and the GTO combined. Similar to the MOSFET, the IGBT has a high impedance gate, which requires only a small amount of energy to switch the device. Like the BJT, the IGBT has a small on-state voltage even in devices with large blocking voltage ratings (for example, V_{on} is 2-3 V in a 1000 V device). Similar to the GTO, IGBTs can be designed to block negative voltages, as their idealized switch characteristics shown in Fig. 2.32(c) indicate.

Insulated gate bipolar transistors have turn-on and turn-off times on the order of 1 μs and are available in module ratings as large as 1700 V and 1200 A. Voltage ratings of up to 2- 3 kV are projected.

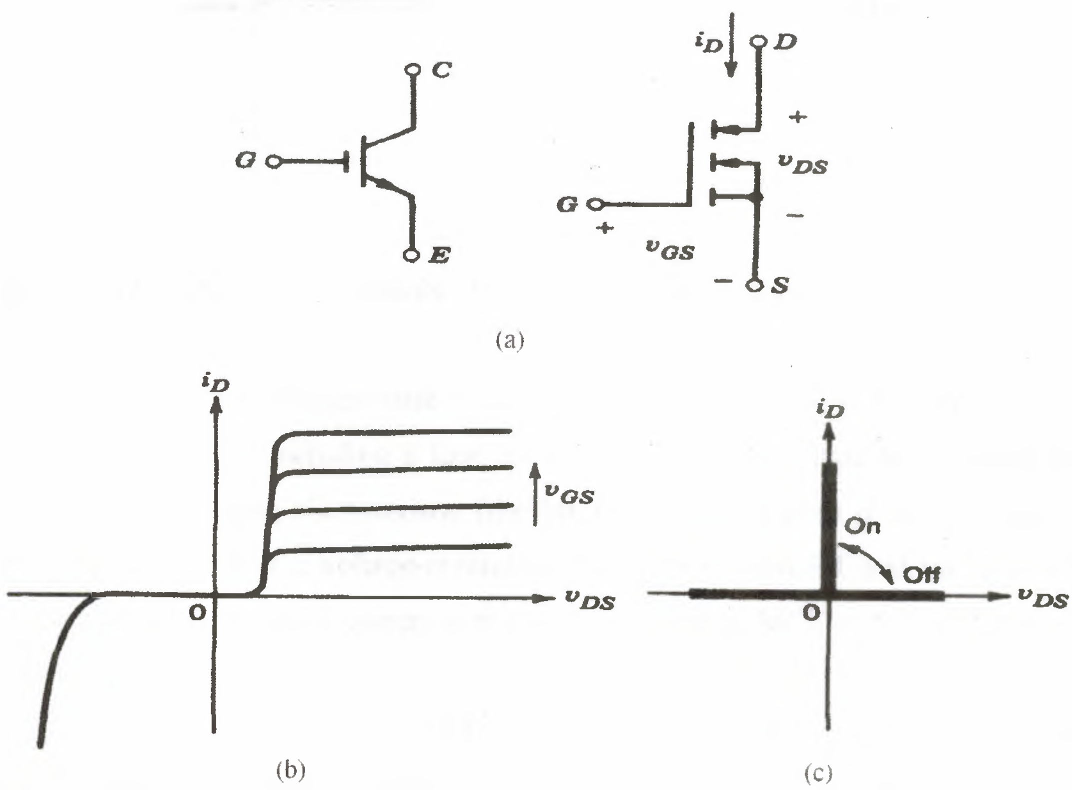


Figure 2.32 An IGBT : (a) symbol, (b) *i-v* characteristics, (c) idealized characteristics

2.4.7 MOS-CONTROLLED THYRISTORS

The MOS-controlled thyristor (MCT) is a new device that has just appeared on the commercial market. Its circuit symbol is shown in Fig. 2.33(a), and its *i-v* characteristic is shown in Fig. 2.33(b). The two slightly different symbols for the MCT denote whether the device is a P-MCT or an N-MCT. The difference between the two arises from the different locations of the control terminals.

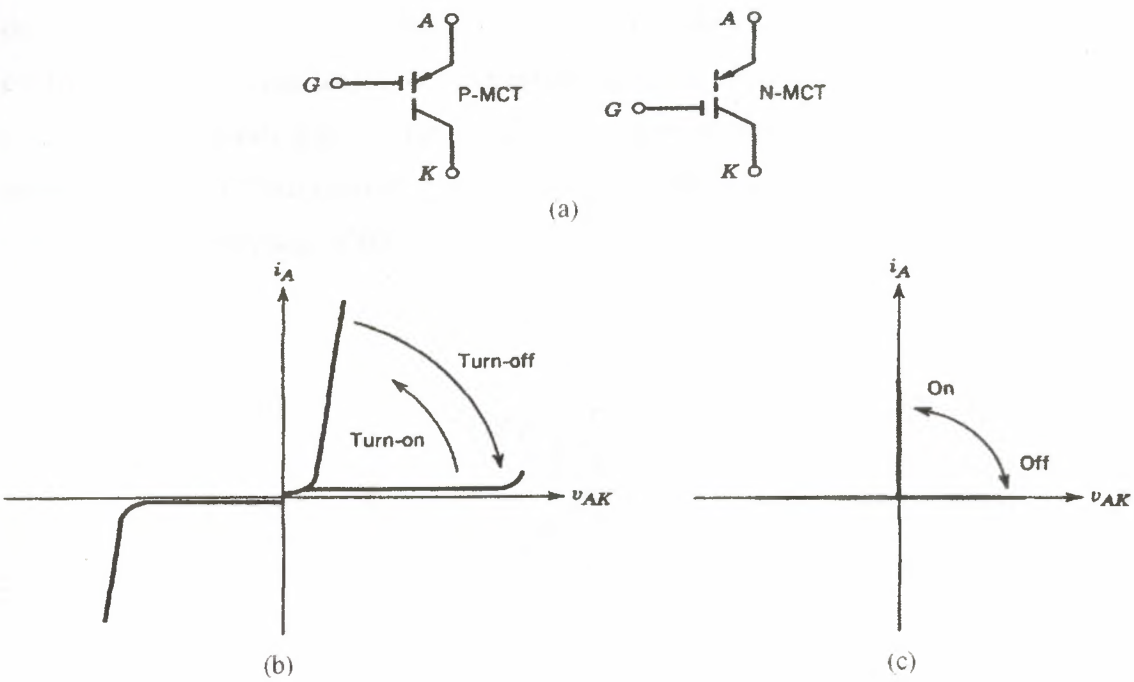


Figure 2.33 An MCT : (a) symbols, (b) *i-v* characteristics, (c) idealized characteristics

From the *i-v* characteristic it is apparent that the MCT has many of the properties of a GTO, including a low voltage drop in the on state at relatively high currents and a latching characteristic (the MCT remains on even if the gate drive is removed). The MCT is a voltage-controlled device like the IGBT and the MOSFET, and approximately the same energy is required to switch an MCT as for a MOSFET or an IGBT.

The MCT has two principal advantages over the GTO, including much simpler drive requirements (no large negative gate current required for turn-off like the GTO) and faster switching speeds (turn-on and turn-off times of a few microseconds). The MCTs have smaller on-state voltage drops compared to IGBTs of similar ratings and are presently available in voltage ratings to 1500 V with current ratings of 50 A to a few hundred amperes. Devices with voltage ratings of 2500-3000 V have been demonstrated in prototypes and will be available soon. The current ratings of individual MCTs are significantly less than those of GTOs because individual MCTs cannot be made as large in cross-sectional area as a GTO due to their more complex structure.

2.4.8 DESIRED CHARACTERISTICS IN CONTROLLABLE SWITCHES

As mentioned in the previous sections, several types of semiconductor power devices including BJTs, MOSFETs, GTOs, and IGBTs can be turned on and off by control signals applied to the control terminal of the device. These devices we term *controllable switches* and are represented in a generic manner by the circuit symbol shown in Fig. 2.34. No current flows when the switch is off, and when it is on, current can flow in the direction of the arrow only.

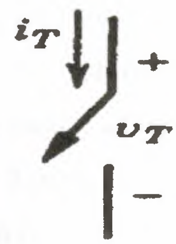


Figure 2.34 Generic controllable switch

The ideal controllable switch has the following characteristics:

1. Block arbitrarily large forward and reverse voltages with zero current flow when off.
2. Conduct arbitrarily large currents with zero voltage drop when on
3. Switch from on to off or vice versa instantaneously when triggered.
4. Vanishingly small power required from control source to trigger the switch.

Real devices, as we intuitively expect, do not have these ideal characteristics and hence will dissipate power when they are used in the numerous applications already mentioned. If they dissipate too much power, the devices can fail and, in doing so, not only will destroy themselves but also may damage the other system components.
Power dissipation in semiconductor power devices is fairly generic in nature; that is, the same basic factors governing power dissipation apply to all devices in the same manner.

The converter designer must understand what these factors are and how to minimize the power dissipation in the devices.

In order to consider power dissipation in a semiconductor device, a controllable switch is connected in the simple circuit shown in Fig. 2.35(a). This circuit models a very commonly encountered situation in power electronics; the current flowing through a switch also must flow trough some series inductance(s). The dc current source approximates the current that would actually flow due to inductive energy storage. The diode is assumed to be ideal because our focus is on the switch characteristics, though in practice the diode reverse-recovery current can significantly affect the stresses on the switch.

When the switch is on, the entire current I_o flows through the switch and the diode is reverse biased. When the switch is turned off, I_o flows through the diode and a voltage equal to the input voltage V_d appears across the switch, assuming a zero voltage drop across the ideal diode. Figure 2.35(b) shows the waveforms for the current through the switch and the voltage across the switch when it is being operated at a repetition rate or switching frequency of $f_s = 1/ T_s$,with T_s being the switching time period. The switching waveforms are represented by linear approximations to the actual waveforms in order to simplify the discussion.

When the switch has been off for a while, it is turned on by applying a positive control signal to the switch, as is shown in Fig. 2.35(b). During the turn-on transition of this generic switch, the current buildup consists of a short delay time $t_{d(on)}$ followed by the current rise time t_{ri}. Only after the current I_o flows entirely through the switch can the diode become reverse biased and the switch voltage fall to a small on-state value of V_{on} with a voltage fall time of t_{fv}. The waveforms in Fig. 2.35(b) indicate that large values of switch voltage and current are present simultaneously during the turn-on crossover interval $t_{c(on)}$, where

$$t_{c(on)} = t_{ri} + t_{fv}$$

The energy dissipated in the device during this turn-on transition can be approximated from Fig. 2.35(c) as

$$W_{c(on)} = \tfrac{1}{2} V_d I_o t_{c(on)}$$

where it is recognized that no energy dissipation occurs during the tum-on delay interval $t_{d(on)}$.

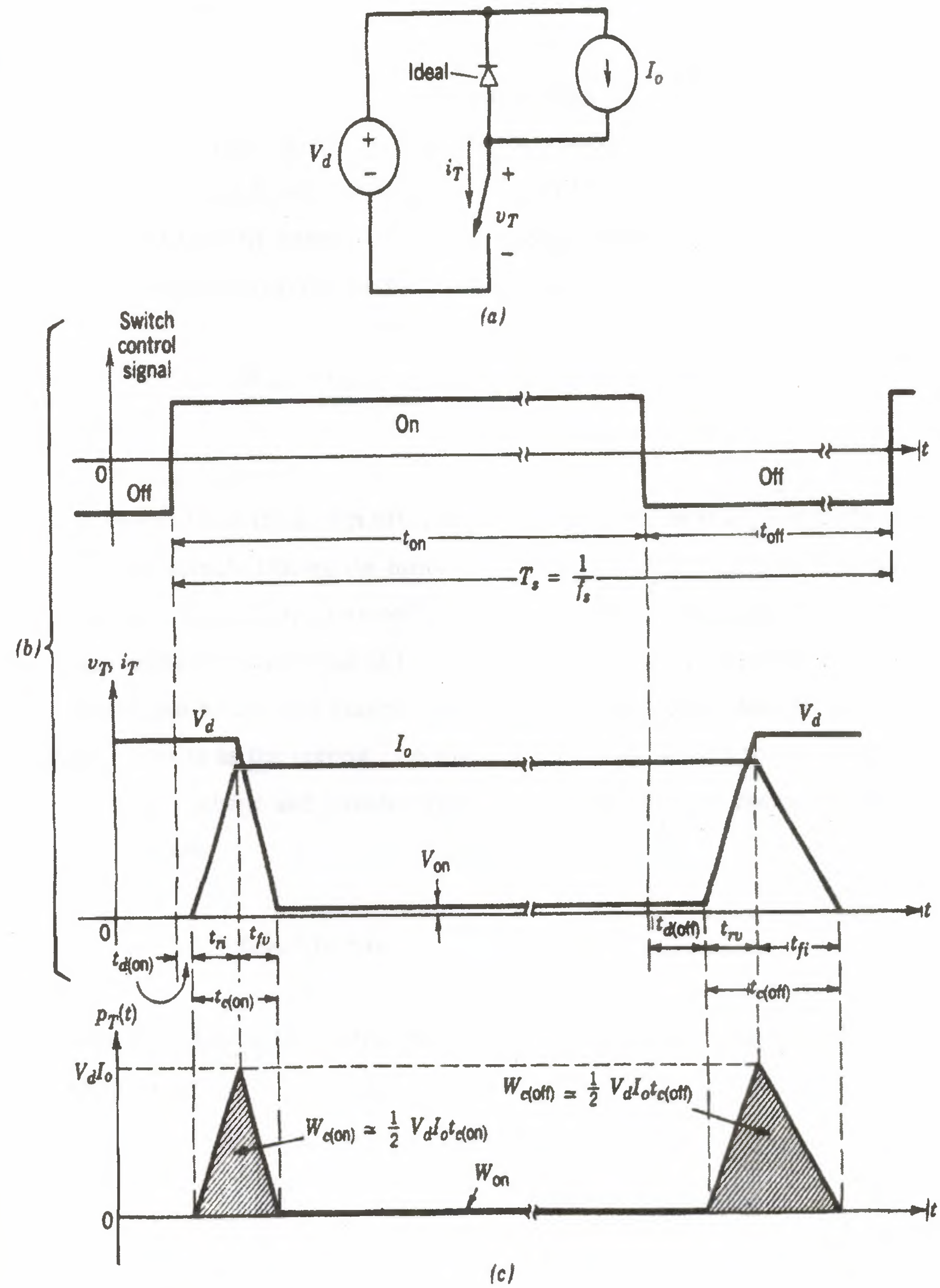


Figure 2.35 Generic-switch characteristic (linearized)

In Fig. 2.35 we are considering the below facts:

a) Simplified clamped-inductive-switching circuit

b) Switch waveforms

c) Instantaneous switch power loss

Once the switch is fully on, the on-state voltage V_{on} will be on the order of a volt or so depending on the device, and it will be conducting a current I_o. The switch remains in conduction during the on interval $t_{(on)}$, which in general is much larger than the turn-on and turn-off transition times. The energy dissipation $W_{(on)}$ in the switch during this on-state interval can be approximated as

$$W_{(on)} = V_{on} I_o t_{on}$$

where $t_{on} >> t_{c(on)}, t_{c(off)}$.

In order to turn the switch off, a negative control signal is applied to the control terminal of the switch. During the turn-off transition period of the generic switch, the voltage build-up consists of a turn-off delay time $t_{d(off)}$ and a voltage rise time t_{rv}. Once the voltage reaches its final value of V_d *(see* Fig. 2.35(a)), the diode can become forward biased and begin to conduct current. The current in the switch falls to zero with a current fall time t_{fi} as the current I_o commutates from the switch to the diode. Large values of switch voltage and switch current occur simultaneously during the crossover interval $t_{c(off)}$, where

$$t_{c(off)} = t_{rv} + t_{fi}$$

The energy dissipated in the switch during this turn-off transition can be written, using Fig. 2.35(c), as

$$W_{c(off)} = ½ V_d I_o t_{c(on)}$$

where any energy dissipation during the turn-off delay interval $t_{d(off)}$ is ignored since it is small compared to $W_{c(off)}$.

The instantaneous power dissipation $p_T(t) = v_T i_T$ plotted in Fig. 2.35(c) makes it clear that a large instantaneous power dissipation occurs in the switch during the turn-on and turn-off intervals. There are f_s, such turn-on and turn-off transitions per second. Hence the average switching power loss P_s, in the switch due to these transitions can be approximated as

$$P_s = \tfrac{1}{2}\, V_d I_o f_s (\, t_{c(on)} + t_{c(off)})$$

This is an important result because it shows that the switching power loss in a semiconductor switch varies linearly with the switching frequency and the switching times. Therefore, if devices with short switching times are available, it is possible to operate at high switching frequencies in order to reduce filtering requirements and at the same time keep the switching power loss in the device from being excessive.

The other major contribution to the power loss in the switch is the average power dissipated during the on-state P_{on} which varies in proportion to the on-state voltage. P_{on} is given by

$$P_{on} = V_{on} Io\, (t_{on} / T_s)$$

which shows that the on-stage voltage in a switch should be as small as possible.

The leakage current during the off state (switch open) of controllable switches is negligibly small, and therefore the power loss during the off state can be neglected in practice. Therefore, the total average power dissipation PT in a switch equals the sum of P_s and P_{on}.

Form the considerations discussed in the preceding paragraphs, the following characteristics in a controllable switch are desirable:

1. Small leakage current in the off state.
2. Small on-state voltage Von to minimize on-state power losses.
3. Short tum-on and tum-off times. This will permit the device to be used at high switching frequencies.
4. Large forward and reverse voltage-blocking capability. This will minimize the need for series connection of several devices, which complicates the control and protection of the switches. Moreover, most of the device types have a minimum on-state voltage regardless of their blocking voltage rating. A series connection

of several such devices would lead to a higher total on-state voltage and hence higher conduction losses. In most (but not all) converter circuits, a diode is placed across the controllable switch to allow the current to flow in the reverse direction. In those circuits, controllable switches are not required to have any significant re verse-voltage-blocking capability.

5. High on-state current rating. In high-current applications, this would minimize the need to connect several devices in parallel, thereby avoiding the problem of current sharing.

6. Positive temperature coefficient of on-state resistance. This ensures that paralleled devices will share the total current equally.

7. Small control power required to switch the device. This will simplify the control circuit design.

8. Capability to withstand rated voltage and rated current simultaneously while switching. This will eliminate the need for external protection (snubber) circuits across the device.

9. Large *dvldt* and *dildt* ratings. This will minimize the need for external circuits otherwise needed to limit *dvldt* and *dildt* in the device so that it is not damaged.

We should note that the clamped-inductive-switching circuit of Fig. 2.35(a) results in higher switching power loss and puts higher stresses on the switch in comparison to the resistive-switching circuit.

2.4.9 COMPARISON OF CONTROLLABLE SWITCHES

Only a few definite statements can be made in comparing these devices since a number of properties must be considered simultaneously and because the devices are still evolving at a rapid pace. However, the qualitative observations given in Table 2.1 can be made.

It should be noted that in addition to the improvements in these devices, new devices are being investigated. The progress in semiconductor technology will undoubtedly lead to higher power ratings, faster switching speeds, and lower costs. A summary of power device capabilities is shown in Fig. 2.36.

On the other hand, the forced-commutated thyristor, which was once widely used in circuits for controllable switch applications, is no longer being used in new converter designs with the possible exception of power converters in multi-MVA ratings. This is a pertinent example of how the advances in semiconductor power devices have modified converter design.

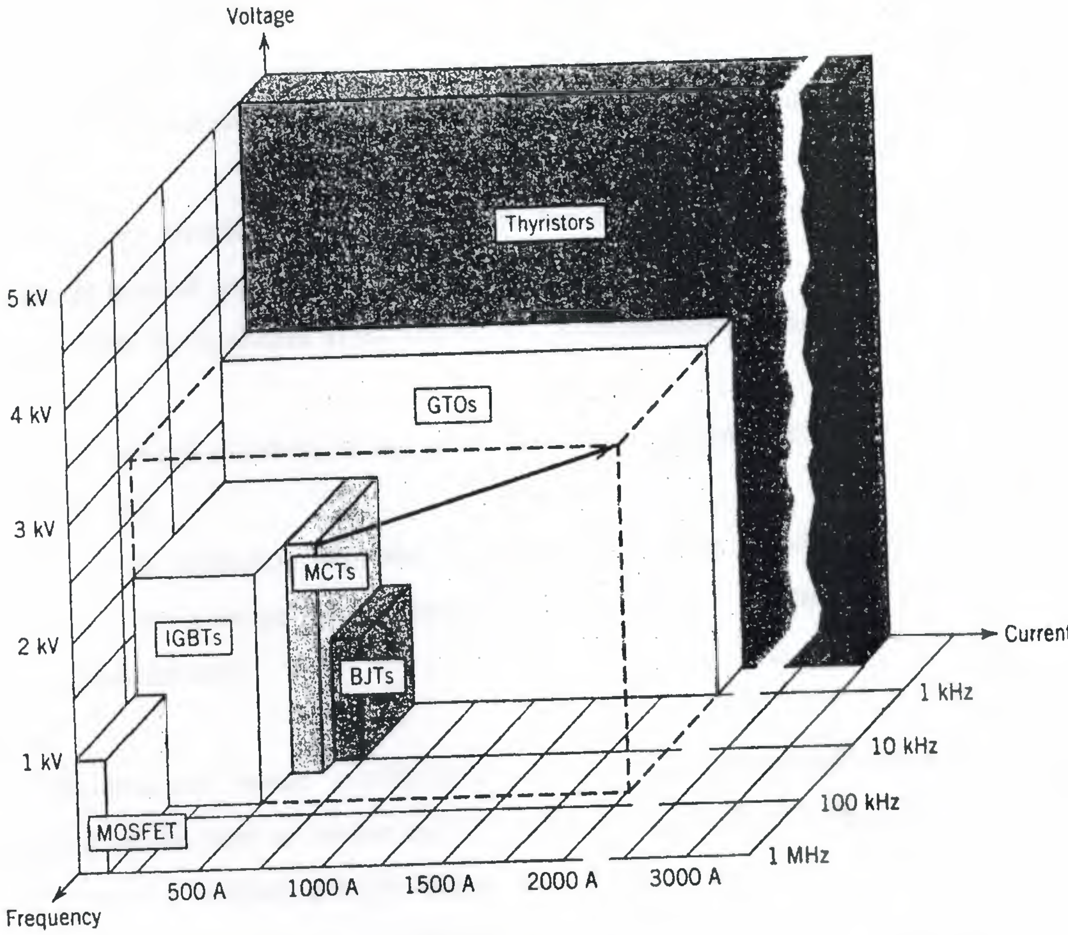


Figure 2.36 Summary of power semiconductor device capabilities

Table 2-1 Relative Properties of Controllable Switches

Device	Power Capability	Switching Speed
BJT / MD	Medium	Medium
MOSFET	Low	Fast
GTO	High	Slow
IGBT	Medium	Medium
MCT	Medium	Medium

2.4.10 DRIVE AND SNUBBER CIRCUITS

In a given controllable power semiconductor switch, its switching speeds and on-state losses depend on how it is controlled. Therefore, for a proper converter design, it is important to design the proper drive circuit for the base of a BJT or the gate of a MOSFET, GTO, or IGBT. The future trend is to integrate a large portion of the drive circuitry along with the power switch within the device package, with the intention that the logic signals, for example, from a microprocessor, can be used to control the switch directly.

Snubber circuits are used to modify the switching waveforms of controllable switches. In general, snubbers can be divided into three categories:

1. Turn-on snubbers to minimize large overcurrents through the device at turn-on.
2. Turn-off snubbers to minimize large overvoltages across the device during turn-off.
3. Stress reduction snubbers that shape the device switching waveforms such that the voltage and current associated with a device are not high simultaneously.

In practice, some combination of snubbers mentioned before are used, depending on the type of device and converter topology. The future trend is to design devices that can withstand high voltage and current simultaneously during the short switching interval and thus minimize the stress reduction requirement. However, for a device with a given characteristic, an alternative to the use of snubbers is to alter the converter topology such that large voltages and currents do not occur at the same time. These converter topologies, called resonant converters.

2.4.11 JUSTIFICATION FOR USING IDEALIZED DEVICE CHARACTERISTICS

In designing a power electronic converter, it is extremely important to consider the available power semiconductor devices and their characteristics. The choice of devices depends on the application. Some of the device properties and how they influence the selection process are listed here:

1. On-state voltage or on-state resistance dictates the conduction losses in the

device.

2. Switching times dictate the energy loss per transition and determine how high the operating frequency can be.

3. Voltage and current ratings determine the device power-handling capability.

4. The power required by the control circuit determines the ease of controlling the device.

5. The temperature coefficient of the device on-state resistance determines the ease of connecting them in parallel to handle large currents.

6. Device cost is a factor in its selection.

In designing a converter from the system viewpoint, the voltage and current requirements must be considered. Other important considerations include acceptable energy efficiency, the minimum switching frequency to reduce the filter and the equipment size, cost, and the like. Hence the device selection must ensure a proper match between the device capabilities and the requirements on the converter.

These observations help to justify the use of idealized device characteristics in analyzing converter topologies and their operation in various applications as follows:

1. Since the energy efficiency is usually desired to be high, the on-state voltage must be small compared to the operating voltages, and hence it can be ignored in analyzing converter characteristics.

2. The device switching times must be short compared to the period of the operating frequency, and thus the switchings can be assumed to be instantaneous.

3. Similarly, the other device properties can be idealized.

The assumption of idealized characteristics greatly simplifies the converter analysis with no significant loss of accuracy. However, in designing the converters, not only must the device properties be considered and compared, but the converter topologies must also be carefully compared based on the properties of the available devices and the intended application.

CHAPTER 3

VARIABLE POWER CIRCUIT

3.1 DEVICES USED IN VARIABLE POWER CIRCUIT

Variable Power Circuit was constructed by using different devices. Each device was performs different tasks in the circuit in coorporation with the others to obtain the results as possible as to near to expected results. Some source of errors are affects our experiment but not too much.

The devices were used in the construction of the circuit are given below with their types and ratings.

3.1.1. RESISTORS

The resistors are passive devices where they are used to drop voltage, limit current, attenuate signals, act as heaters, act as fuses, furnish electrical loads or divide voltages.

In variable power circuit 100 Ω, 390 Ω and 2.2 kΩ resistor were used to do one of the task in above purposes.

3.1.2. POTENTIOMETER

Potentiometer is used to control the level of the signal for the emitter of the unijunction transistor. The 470 kΩ potentiometer is used to adjust the voltage for the UJT by adjusting its resistance value where the UJT operates according to this voltage and creates related output.

3.1.3. UNIJUNCTION TRANSISTOR (UJT)

Unijunction transistor 2N2646 was used in the circuit as pulse generator for the SCR. The characteristics of the UJT was introduced in Chapter 2 and data sheet of 2N2646 was given in Appendix B.

3.1.4. SILICON CONTROLLED RECRIFIER (SCR)

TIC 106 SCR was used as a switch which is only fired by gate current from the UJT. The characteristics of SCR were largely mentioned in Chapter 2 in Thyristor section and data sheet of the TIC 106 was given in Appendix C.

3.1.5. ZENER DIODE

The 9 V Zener Diode was used the keep the voltage on 9V for the devices that are parallel to the zener diode.

3.1.6. LAMP

The 12 V 500 mA lamp was used as a load for the variable power circuit.

3.1.7. BRIDGE RECTIFIER

Bridge Rectifier consist of 4 equvalent diodes which rectifies the a.c. signal into d.c. signal without control in variable power circuit. B308C was used to rectify the 12 V a.c. voltage to required d.c. voltage level for the variablepower circuit.

3.1.8 CAPACITOR

The capacitors are useful for coupling and filter networks or circuits. In the circuit 0.02 μF (22n / 63 V) capacitor was used to do this tasks.

3.2. CIRCUIT OPERATION

Figure 3.1 (a) shows how the SCR can be used, in conjunction with a UJT pulse generator, as a variable power unit feeding a d.c. load. The voltage across ZD1, and thus across the UJT circuitry, is rough d.c. clipped at 9 V, so the power to the generator is automatically connected and disconnected in sympathy with the power line frequency. At the start of each new half cycle, the UJT circuitry starts a timing cycle, after a delay determined by R_5, generates positive (+ ve) pulse and fires SCR. Thus, the UJT gives delayed and variable firing of the SCR.

When the unit is set for minimum output power (in LP), the UJT gives maximum delay, so the SCR fires towards the very end of each half cycle, so only a small part of the total available power is fed to the load. At half maximum power, the UJT fires the SCR half way through each half cycle, so half of the maximum available power is fed to the load. At maximum power, the UJT triggers the SCR towards the start of each half cycle, so almost the full available power is developed on the load. The d.c. power to the load is thus fully variable via R_5, and, since the SCR is used as a switch, the system is highly efficient as a variable power source. The picture of the circuit is shown in Pic. 3.1.

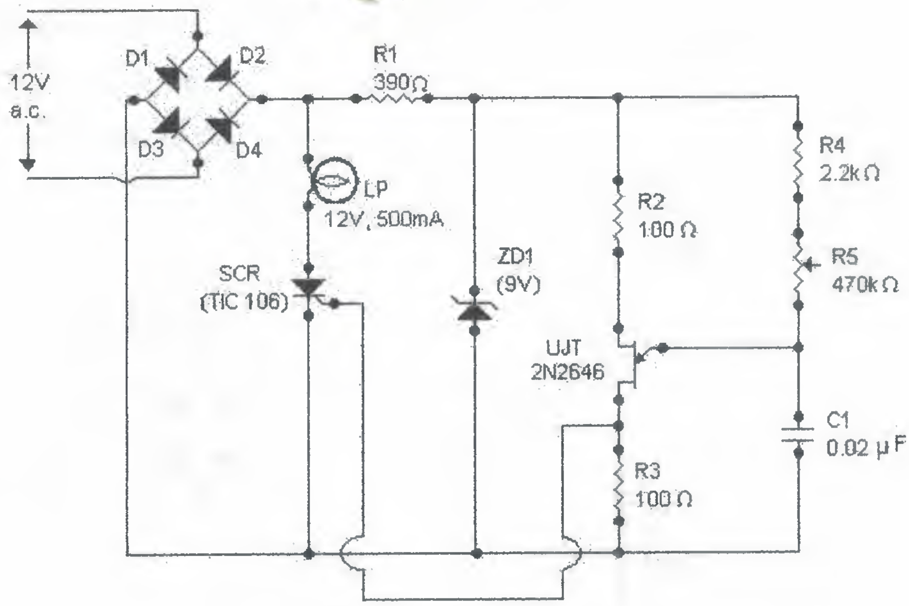
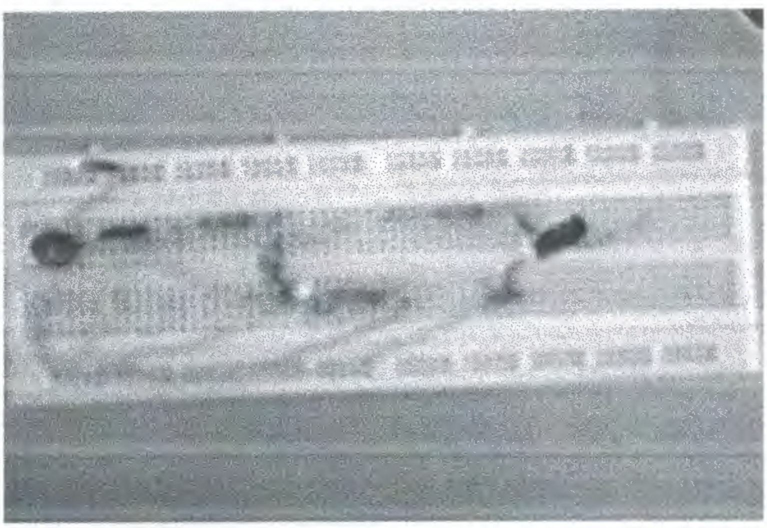


Figure 3.1 (a) Variable power unit, feeding a d.c. load



Picture 3.1 Variable power circuit

Finally, Fig. 3.1 (b) shows how similar circuit can be used to control an a.c. power load. This circuit is nearly same as the circuit we mentioned above except that the load is placed on the a.c side of the bridge rectifier.

In this case, as soon as the UJT triggers the SCR, almost the full supply voltage is developed across the load, so the voltage across SCR and ZD1 fall to near-ground potential. This is of no importance, however, since the SCR has already fired, and thus stays locked-on until its anode falls to full ground potential at the end of the each half cycle. The power to the load can thus be smoothly varied from near-zero to maximum via R_5, as in the case of the d.c. circuit.

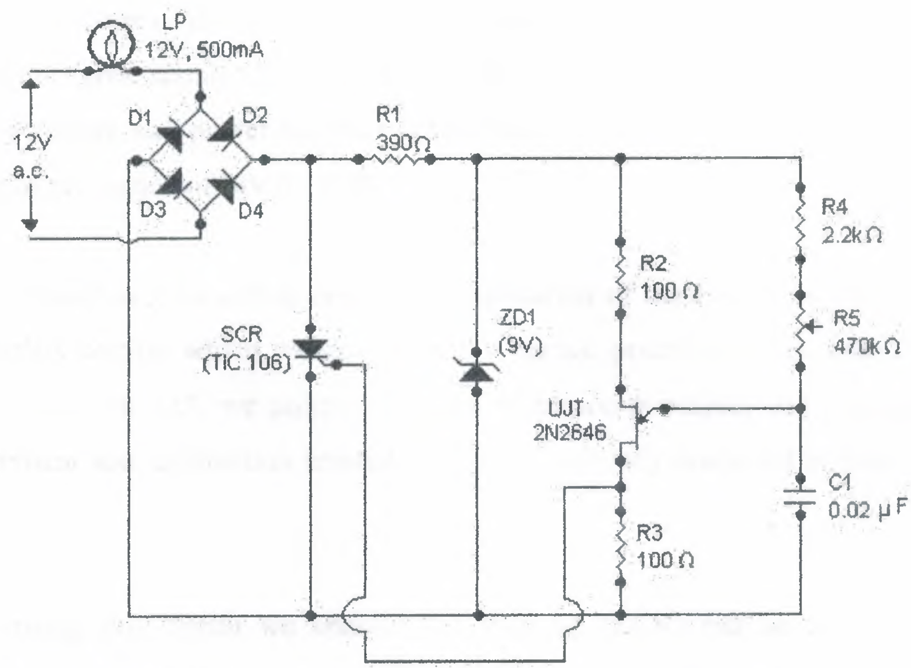


Figure 3.1 (b) Variable power unit, feeding an a.c. load

CONCLUSION

Power Electronics refers to control and conversion of electrical power by power semiconductor devices where these devices operates as switches. Advent of silicon controlled rectifiers abbreviated as SCRs, led to the developments in this area. Before these developments mercury-arc rectifiers were used which were part of industrial electronics and scope of the application was limited. Once the SCRs have been invented the application area spread to many field such as drives, aviation electronics, high frequency inverters and power supplies where they are using in the range from a few VA / watts to thousands of MVA / MW.

Variable-Power Circuit is one of the application of the Power Eectronics under power supplies section which we deal with this circuit practically. For analyzing and observing the circuit well, we utilized from the operation principles and characteristics of both thyristor and unijunction transistor which we already discussed in main section of this Thesis.

By using this circuit we showed the thyristor (SCR) can be used under the control of unijunction transistor, as a variable power unit feeding a dc load. By using the unijunction transistor with the adjustable resistor, the dealy and fire time of SCR were determined to obtain the corresponding output waveforms for each input power.

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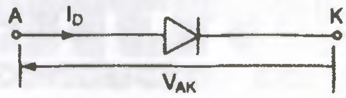
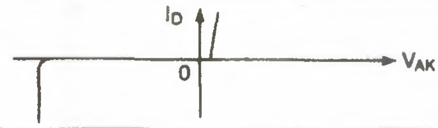
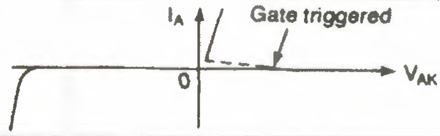
Devices	Symbols	Characteristics
Diode		
Thyristor		
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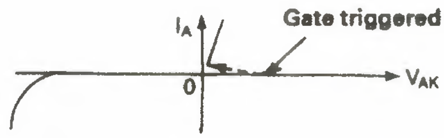
GTO		
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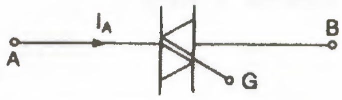
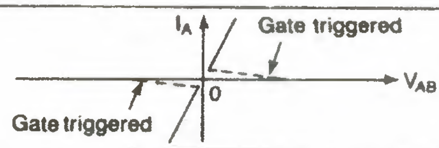
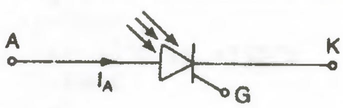
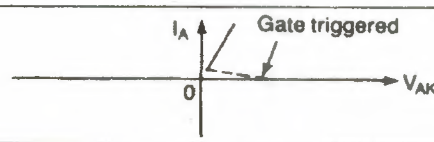
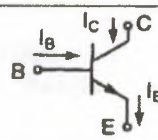
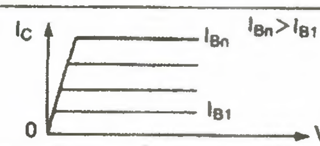
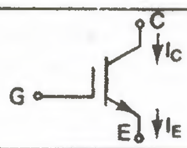
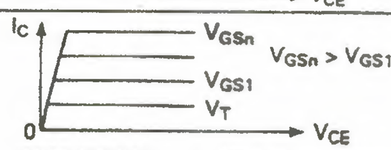
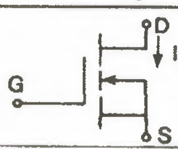
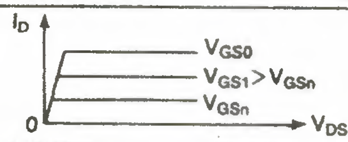
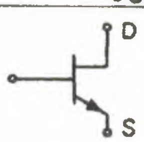
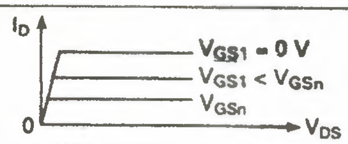
TRIAC		
LASCR		
NPN BJT		
IGBT		
N-Channel MOSFET		
SIT		

Appendix A

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Manufacturers of World Class Discrete Semiconductors

2N2646
2N2647

SILICON UNIJUNCTION TRANSISTOR

JEDEC TO-18 CASE*

DESCRIPTION

The CENTRAL SEMICONDUCTOR 2N2646, 2N2647 types are silicon PN unijunction transistors designed for general purpose industrial applications.

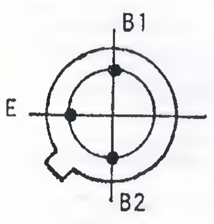
MAXIMUM RATINGS (T_A=25°C unless otherwise noted)

	SYMBOL		UNIT
RMS Power Dissipation	$P_{D(RMS)}$	300	mW
RMS Emitter Current	$I_{E(RMS)}$	50	mA
Peak Pulse Emitter Current	i_E	2.0	A
Interbase Voltage	V_{B2B1}	35	V
Emitter Reverse Voltage	V_{B2E}	30	V
Operating and Storage Junction Temperature	T_J, T_{STG}	-65 TO +150	°C

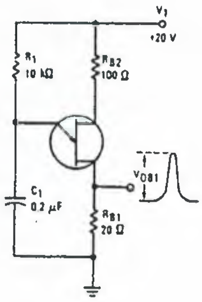
ELECTRICAL CHARACTERISTICS (T_A=25°C unless otherwise noted)

SYMBOL	TEST CONDITIONS	2N2646 MIN	2N2646 MAX	2N2647 MIN	2N2647 MAX	UNIT
η	V_{B2B1}=10V†	0.56	0.75	0.68	0.82	-
R_{BBO}	V_{B2B1}=3.0V, I_E=0	4.7	9.1	4.7	9.1	kΩ
$I_{B2(MOD)}$	V_{B2B1}=10V, I_E=50mA	15 TYP		15 TYP		mA
αR_{BBO}	V_{B2B1}=3.0V, T_A=-65°C to +150°C	0.1	0.9	0.1	0.9	%/°C
$V_{EB1(SAT)}$	V_{B2B1}=10V, I_E=50mA	3.5 TYP		3.5 TYP		V
I_{EO}	V_{B2E}=30V, I_{B1}=0		12		0.2	µA
I_P	V_{B2B1}=25V		5.0		2.0	µA
I_V	V_{B2B1}=20V, R_{B2}=100Ω	4.0	-	8.0	18	mA
V_{OB1}	See test circuit below	3.0		6.0		V

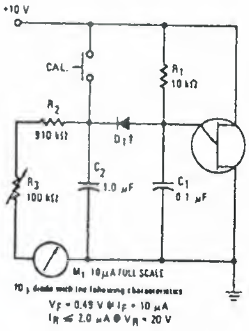
*Conforms to JEDEC TO-18 outline except for lead configuration.



V_{OB1} TEST CIRCUIT



†η TEST CIRCUIT



C106 Series

Preferred Device

Sensitive Gate Silicon Controlled Rectifiers

Reverse Blocking Thyristors

Glassivated PNPN devices designed for high volume consumer applications such as temperature, light, and speed control; process and remote control, and warning systems where reliability of operation is important.

- Glassivated Surface for Reliability and Uniformity
- Power Rated at Economical Prices
- Practical Level Triggering and Holding Characteristics
- Flat, Rugged, Thermopad Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Sensitive Gate Triggering
- Device Marking: Device Type, e.g., C106B, Date Code

MAXIMUM RATINGS (T_J = 25°C unless otherwise noted)

Rating	Symbol	Value	Unit
Peak Repetitive Off–State Voltage(1) (Sine Wave, 50–60 Hz, R_{GK} = 1 kΩ, T_C = –40° to 110°C) C106B C106D, C106D1 C106M, C106M1	V_{DRM}, V_{RRM}	 200 400 600	Volts
On-State RMS Current (180° Conduction Angles, T_C = 80°C)	$I_{T(RMS)}$	4.0	Amps
Average On–State Current (180° Conduction Angles, T_C = 80°C)	$I_{T(AV)}$	2.55	Amps
Peak Non-Repetitive Surge Current (1/2 Cycle, Sine Wave, 60 Hz, T_J = +110°C)	I_{TSM}	20	Amps
Circuit Fusing Considerations (t = 8.3 ms)	I^2t	1.65	A^2s
Forward Peak Gate Power (Pulse Width ≤ 1.0 μsec, T_C = 80°C)	P_{GM}	0.5	Watt
Forward Average Gate Power (Pulse Width ≤ 1.0 μsec, T_C = 80°C)	$P_{G(AV)}$	0.1	Watt
Forward Peak Gate Current (Pulse Width ≤ 1.0 μsec, T_C = 80°C)	I_{GM}	0.2	Amp
Operating Junction Temperature Range	T_J	–40 to +110	°C
Storage Temperature Range	T_{stg}	–40 to +150	°C
Mounting Torque(2)	—	6.0	in. lb.

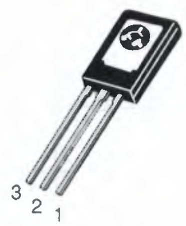
(1) V_{DRM} and V_{RRM} for all types can be applied on a continuous basis. Ratings apply for zero or negative gate voltage; however, positive gate voltage shall not be applied concurrent with negative potential on the anode. Blocking voltages shall not be tested with a constant current source such that the voltage ratings of the devices are exceeded.

(2) Torque rating applies with use of compression washer (B52200F006). Mounting torque in excess of 6 in. lb. does not appreciably lower case-to-sink thermal resistance. Anode lead and heatsink contact pad are common.

ON Semiconductor

http://onsemi.com

SCRs
4 AMPERES RMS
200 thru 600 VOLTS



TO–225AA
(formerly TO–126)
CASE 077
STYLE 2

PIN ASSIGNMENT	
1	Cathode
2	Anode
3	Gate

ORDERING INFORMATION

Device	Package	Shipping
C106B	TO225AA	500/Box
C106D	TO225AA	500/Box
C106D1	TO225AA	500/Box
C106M	TO225AA	500/Box
C106M1	TO225AA	500/Box

Preferred devices are recommended choices for future use and best overall value.

THERMAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3.0	°C/W
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	75	°C/W
Maximum Lead Temperature for Soldering Purposes 1/8″ from Case for 10 Seconds	T_L	260	°C

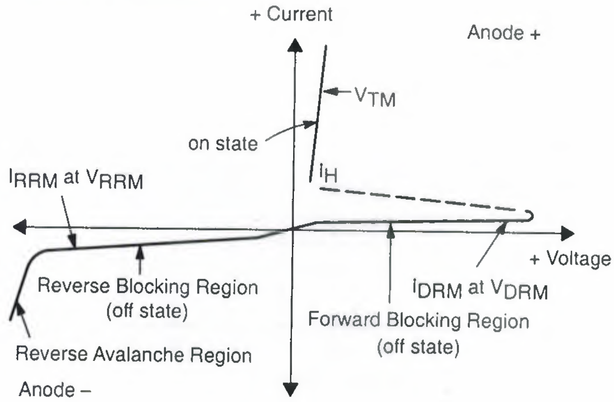
ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted.)

Characteristic		Symbol	Min	Typ	Max	Unit
OFF CHARACTERISTICS						
Peak Repetitive Forward or Reverse Blocking Current (V_{AK} = Rated V_{DRM} or V_{RRM}, R_{GK} = 1000 Ohms)	T_J = 25°C T_J = 110°C	I_{DRM}, I_{RRM}	— —	— —	10 100	μA μA
ON CHARACTERISTICS						
Peak Forward On–State Voltage(1) (I_{FM} = 1 A Peak for C106B, D, & M) (I_{FM} = 4 A Peak for C106D1, & M1)		V_{TM}	—	—	2.2	Volts
Gate Trigger Current (Continuous dc)(2) (V_{AK} = 6 Vdc, R_L = 100 Ohms)	T_J = 25°C T_J = −40°C	I_{GT}	— —	15 35	200 500	μA
Peak Reverse Gate Voltage (I_{GR} = 10 μA)		V_{GRM}	—	—	6.0	Volts
Gate Trigger Voltage (Continuous dc)(2) (V_{AK} = 6 Vdc, R_L = 100 Ohms)	T_J = 25°C T_J = −40°C	V_{GT}	0.4 0.5	.60 .75	0.8 1.0	Volts
Gate Non–Trigger Voltage (Continuous dc)(2) (V_{AK} = 12 V, R_L = 100 Ohms, T_J = 110°C)		V_{GD}	0.2	—	—	Volts
Latching Current (V_{AK} = 12 V, I_G = 20 mA)	T_J = 25°C T_J = −40°C	I_L	— —	.20 .35	5.0 7.0	mA
Holding Current (V_D = 12 Vdc) (Initiating Current = 20 mA, Gate Open)	T_J = 25°C T_J = −40°C T_J = +110°C	I_H	— — —	.19 .33 .07	3.0 6.0 2.0	mA
DYNAMIC CHARACTERISTICS						
Critical Rate–of–Rise of Off–State Voltage (V_{AK} = Rated V_{DRM}, Exponential Waveform, R_{GK} = 1000 Ohms, T_J = 110°C)		dv/dt	—	8.0	—	V/μs

(1) Pulse Test: Pulse Width ≤ 2.0 ms, Duty Cycle ≤ 2%.
(2) R_{GK} is not included in measurement.

Voltage Current Characteristic of SCR

Symbol	Parameter
V_{DRM}	Peak Repetitive Off State Forward Voltage
I_{DRM}	Peak Forward Blocking Current
V_{RRM}	Peak Repetitive Off State Reverse Voltage
I_{RRM}	Peak Reverse Blocking Current
V_{TM}	Peak On State Voltage
I_H	Holding Current



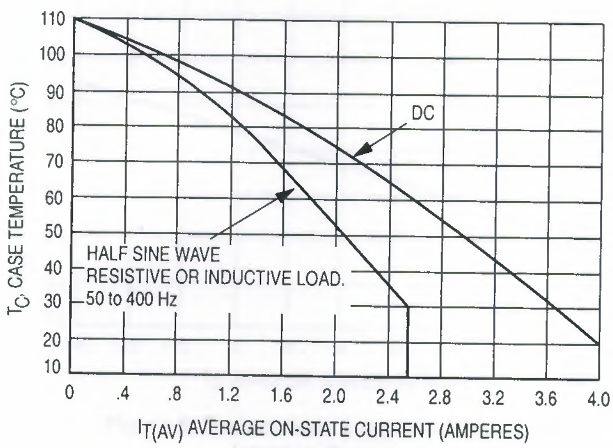


Figure 1. Average Current Derating

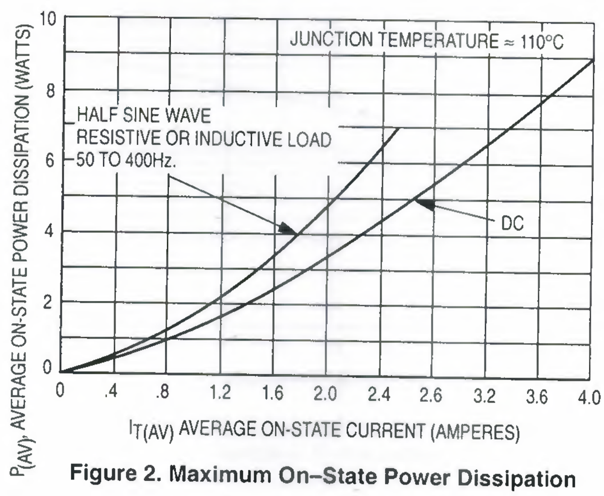


Figure 2. Maximum On–State Power Dissipation

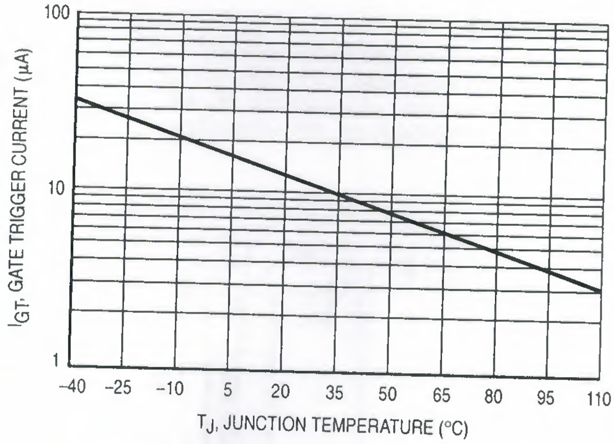


Figure 3. Typical Gate Trigger Current versus Junction Temperature

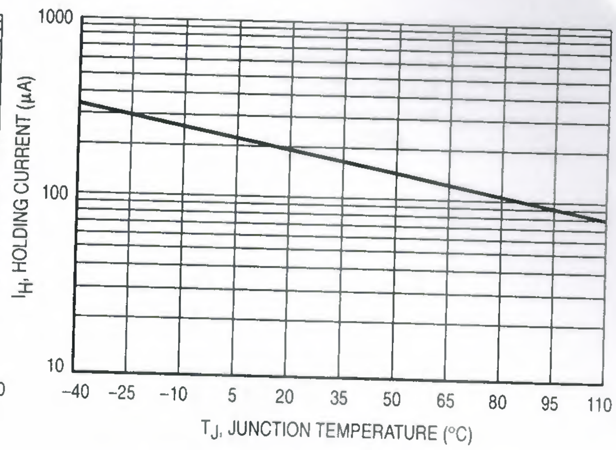


Figure 4. Typical Holding Current versus Junction Temperature

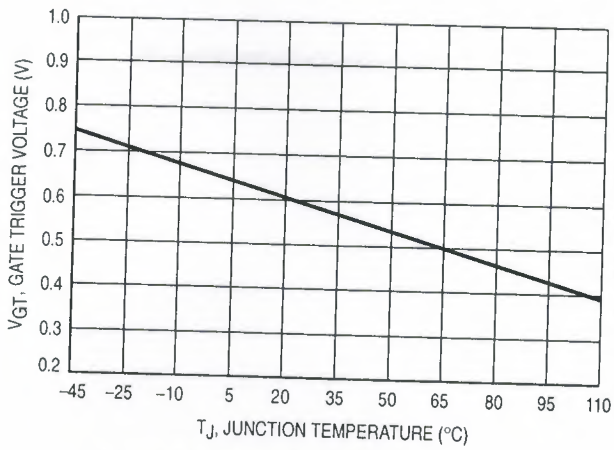


Figure 5. Typical Gate Trigger Voltage versus Junction Temperature

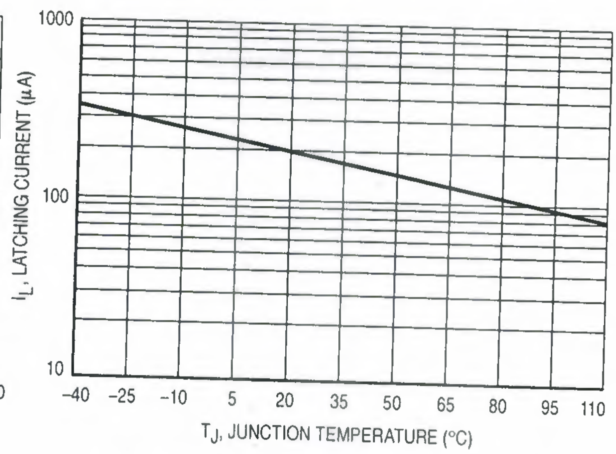
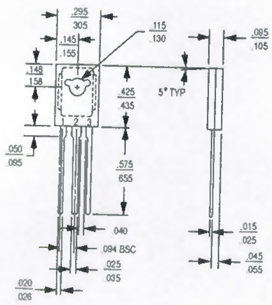


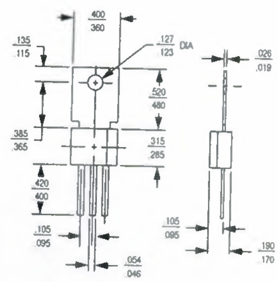
Figure 6. Typical Latching Current versus Junction Temperature

Package Interchangeability

The dimensional diagrams below compare the critical dimensions of the ON Semiconductor C-106 package with competitive devices. It has been demonstrated that the smaller dimensions of the ON Semiconductor package make it compatible in most lead-mount and chassis-mount applications. The user is advised to compare all critical dimensions for mounting compatibility.



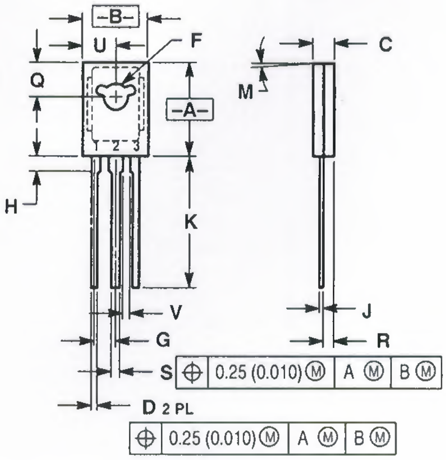
ON Semiconductor C-106 Package



Competitive C-106 Package

PACKAGE DIMENSIONS

TO–225AA
(formerly TO–126)
CASE 077–09
ISSUE W



NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.

	INCHES		MILLIMETERS	
DIM	MIN	MAX	MIN	MAX
A	0.425	0.435	10.80	11.04
B	0.295	0.305	7.50	7.74
C	0.095	0.105	2.42	2.66
D	0.020	0.026	0.51	0.66
F	0.115	0.130	2.93	3.30
G	0.094 BSC		2.39 BSC	
H	0.050	0.095	1.27	2.41
J	0.015	0.025	0.39	0.63
K	0.575	0.655	14.61	16.63
M	5° TYP		5° TYP	
Q	0.148	0.158	3.76	4.01
R	0.045	0.065	1.15	1.65
S	0.025	0.035	0.64	0.88
U	0.145	0.155	3.69	3.93
V	0.040	—	1.02	—

STYLE 2:
PIN 1. CATHODE
2. ANODE
3. GATE