Contents

# SPEED CONTROL OF DC MOTOR USING IGBT

*A PROJECT REPORT SUBMITTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS*

*FOR THE DEGREE OF*

**BACHELOR OF TECHNOLOGY IN ELECTRICAL ENGINEERING**

By

**Anant Kumar (10302006)**

**Mohammed Rehan Memon (10302008)**

**Shreerang Pradeep Munshi (10302009)**

**Sitanshu Mishra (10302015)**

**Under the guidance of**

**PROF. S. RAUTA**

**National Institute of Technology**
**Rourkela**
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# SPEED CONTROL OF DC MOTOR USING IGBT

Anant Kumar, Shreerang Pradeep Munshi., Mohammed Rehan Memon, Sitanshu Mishra
Supervision: Prof S.Routa
Technical Assistant: Mr. Ravindra Nayak

## ABSTRACT

DC Motor speed control is carried out by use of Four Quadrant Chopper drive. Insulated Gate Bipolar Transistors are used for speed control of the motor and the IGBT triggering is carried out by use of PWM converters under various loading conditions and by varying armature voltage and field voltage. The above mentioned experiment was again carried out using Thyristors and a comparative study was made.

## INTRODUCTION

DC motors are used extensively in adjustable-speed drives and position control applications. Their speeds below the base speed can be controlled by armature-voltage control. Speeds above the base speed are obtained by field-flux control. As speed control method for DC motors are simpler and less expensive than those for the AC motors, DC motors are preferred where wide-speed range control is required. DC choppers also provide variable dc output voltage from a fixed dc input voltage.

The Chopper circuit used can operate in all the four quadrants of the V-I plane. The output voltage and current can be controlled both in magnitude as well as in direction so the power flow can be in either direction. The four-quadrant chopper is widely used in reversible dc motor drives. By applying chopper it is possible to implement regeneration and dynamic braking for dc motors

### EXPERIMENTAL DETAILS

The experimental set up consists of a Four-quadrant IGBT based chopper driver model PEC-16HV3 .The ac supply in fed to the setup through an isolation transformer and it is rectified to dc for its use. The PWM converters generate pulse-modulated signal that are compared with the base signal and are fed to OPTO. Delay logic is provided to gate drivers and thus the signal obtained is the gating signal for the IGBTs in the four-quadrant chopper. Once the IGBTs are triggered they are used in pairs to control the speed of the dc motor.

DC motor Specification: 220V, 2.2A, 1420 rpm Shunt type single phase.

Another set up consisted of half bridge rectifier consisting of thyristor wherein the speed control for the same DC motor was carried out using the firing angle of the thyristor.

## CONCLUSION

1. Speed varies directly with armature voltage by keeping field voltage constant.
2. Speed varies inversely with field voltage by keeping armature voltage constant.
3. Armature voltage control gives the speed below the base speed whereas field control gives the speed control above the base speed.
4. Armature current vs. Speed at constant flux gives a drooping characteristic. Though it should have been a straight line parallel to x-axis but due to saturation effect there is slight decrease in speed and shows a drooping characteristic.
5. The IGBT based circuit gives smoother control over the entire speed range as compared with the SCR based circuit.

The above conclusions were found to be in accordance with the theoretical results.

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# LIST OF FIGURES

## CHAPTER 1



## CHAPTER 2



## CHAPTER 3

## LIST OF TABLES AND GRAPHS

### Chapter 5

# Chapter 1

# INTRODUCTION

## 1.1 INTRODUCTION

The chopper circuit shown in fig.1 can operate in all four quadrants of the Vo-Io plane. That is the output voltage and current can be controlled both in magnitude and direction. Therefore, the power flow can be in any direction.

In the first quadrant the power flows from the source to the load and is assumed to be (+ve).

In the second quadrant, the voltage is still positive but the current is negative. Therefore, the power is negative. In this case, the power flows from load to source and this can happen if the load is inductive or back emf source such as a dc motor.

In the third quadrant both the voltage and current are negative but the power is positive.

In the fourth quadrant voltage is negative but current is positive. The power is therefore negative.

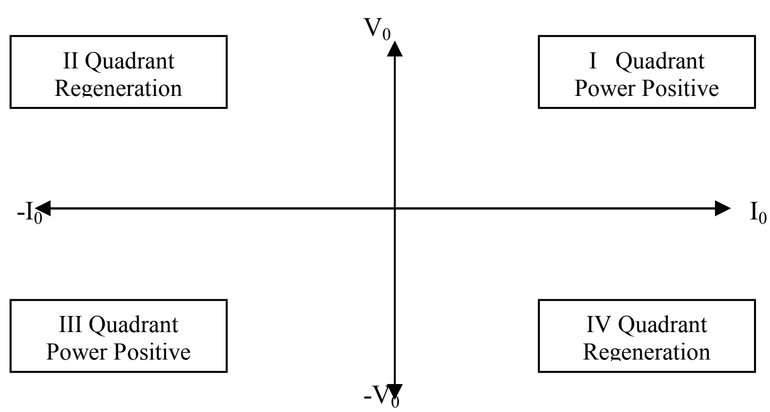



Fig 1.1: Four quadrant of $V_0$, $I_0$ plane

This chopper is widely used in reversible dc motors drives. The reversible dc motor drive requires power flow in either direction in order to achieve fast dynamics response. By employing four-quadrant chopper it is possible to implement regeneration and dynamic braking by means of which fast dynamic response is achieved.

## 1.2 CIRCUIT DESCRIPTION

The four quadrant chopper with four switching devices where diodes are connected in anti parallel with the switching devices is also referred to as full bridge converter topology. The input to the full bridge converter is fixed magnitude dc voltage $V_{dc}$. The output of the converter can be a variable dc voltage with either polarity. The circuit is therefore called as four quadrant chopper circuit or dc to dc converter. The output of the full bridge converter can also be an ac voltage with variable frequency and amplitude in which case the converter is called as dc- to-ac conversion ( inverter). In a full bridge converter when a gating signal is given to a switching device either the switching device or the diode only will conduct depending on the directions of the output load current.

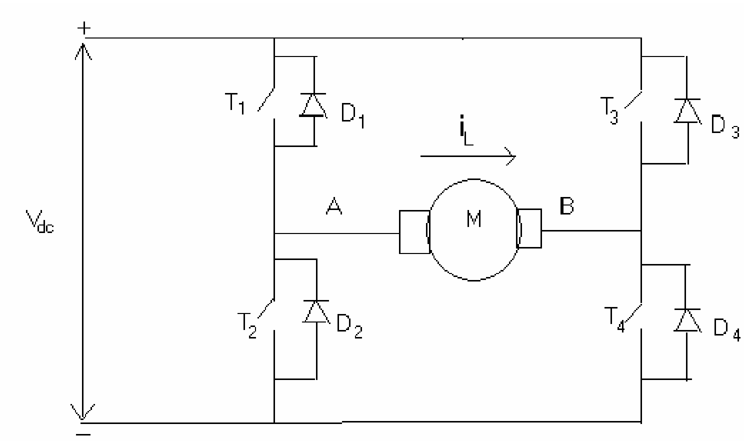



Fig 1.2: Four quadrant chopper circuit

## 1.3 SWITCHING MODES OF FOUR QUADRANT CHOPPER

The switches in the four quadrant chopper can be switched in two different modes such that:

- The output voltage swings in both direction i.e. from $+V_{dc}$ to $–V_{dc}$. This mode of switching is referred to as PWM with bipolar voltage switching.
- The output voltage swings either from –zero to $+V_{dc}$ or zero to- $V_{dc}$. This mode of switching is referred to as PWM with unipolar voltage switching.

## 1.4 OPERATION OF THE FOUR QUADRANT CHOPPER WITH BIPOLAR VOLTAGE SWITCHING

The operation of the circuit as a four quadrant chopper with bipolar voltage switching is explained, referring to the circuit diagram of Fig 1.2. When the switches T1 and T4 are turned ON by applying gating signals simultaneously, the load voltages $V_{dc}$ with terminal 'A' positive and the load current $I_L$ flows in the direction from A to B. Because of the load inductance, the current cannot change instantaneously.

The load voltage V will now be – $V_{dc}$ since the conduction of the diode D3 will connect the load terminal B to the (+) ve terminal of the source. As the load voltage is negative and the current is still positive, the power is negative. The power now flows from the load to the source. This corresponds to the operation of chopper circuit in the fourth quadrant. This operation in the fouth quadrant will continue as long as the current is positive. When T1 and T4 are off, T3 and T2 can be turned ON.

When the current passes through zero, the devices T3 and T2 can be turned on, and the load current becomes negative. The load current now passes through T3 and T2 with current direction in the load as from B to A. this brings the operation of the chopper in the third quadrant. Turning of the T3 and T2 will bring in the conduction of the diode D1 and D4 and the operation of the chopper circuit in the second quadrant.

The operation of the chopper in the first and third quadrant corresponds to power flow from the source to the load, and is considered to be forward power flow. The operation in the fourth and second quadrant corresponds to reverse power flow. The relevant waveforms showing the operation of full bridge converter in all the four quadrant is shown in the fig 1.3.

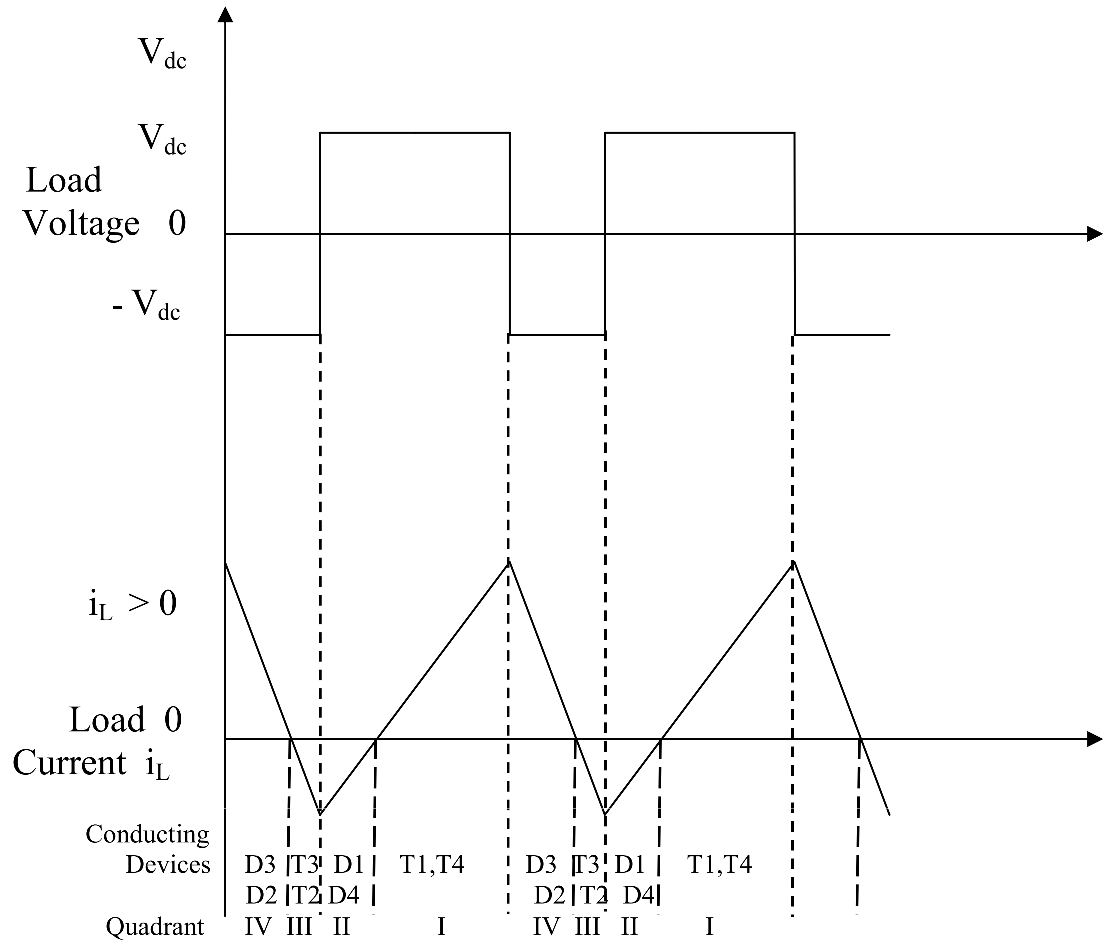



Fig.1.3 Load Voltage and Current with Inductive Load & Load Current $i_L > 0$ (positive)

### 1.5 GENERATION OF GATING SIGNALS

The gating signals for the switches in the four quadrant chopper are derived by comparing a triangular wave with a control voltage level. The generation of gating signals for a unipolar voltage switching is shown in fig 1.4.

The triangular carrier waveform is compared with the control voltage (+)v and (-)v. the pulse generated by comparing +v with triangular carrier is used to turn on T1and its compliment is used to turn on T2. The pulse generated by comparing $-v_e$ with triangular carrier is used to turn on T3 and its complement is used to turn on T4. The voltage varies from $-V_{tri}$ to $+V_{tri.}$

The fig. below shows the schematic of the generator of gating signal for the four quadrant chopper with unipolar switching. A triangular carrier wave of frequency around 2 Khz is generated . The triangular wave is compared with +Vc and –Vc in comparator 1 and comparator 2 respectively.

The output of the comparator-1 gives the gating signal to T1 and its complement gives the gating signal T2. The output of the comparator-2 gives the gating signal to T3 and its complement gives the gating signal to T4.

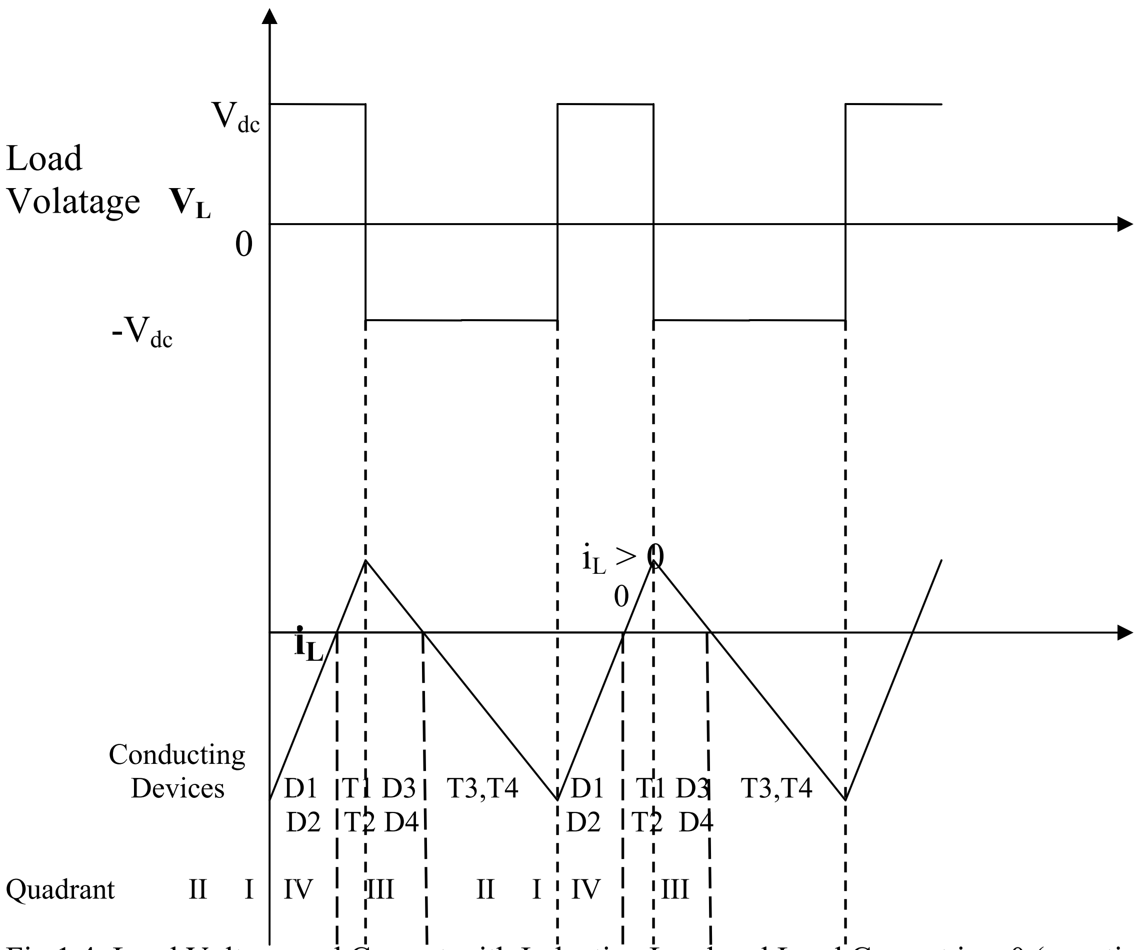



Fig 1.4. Load Voltage and Current with Inductive Load and Load Current $i_L < 0$ (negative)

## 1.6 HARDWARE DESCRIPTION

The hardware involved in the four quadrant chopper drive is screen printed on the front panel .it consists of both the power circuitry and the control circuitry.

**POWER CIRCUIT**

It consists of-

i) single phase diode bridge rectifier

ii) four quadrant chopper

iii) DC link capacitors

iv) Braking circuit

v) Field control chopper

vi) EMI filter.

The diode rectifier rectifies the input ac voltage and provides the dc voltage to the chopper. Large values of dc link capacities maintain a constant dc voltage is also used for the field circuit of the motor through a single quadrant chopper.

The chopper consists of four IGBTs rated at 900V, 60A.

# Chapter 2

# IGBT

# IGBT BASICS

## 2.1 INTRODUCTION

Recent technology advances in power electronics have arisen primarily from improvements in semiconductor power devices, with insulated gate bipolar transistors (IGBT) leading the market today for medium power applications. IGBTs feature many desirable properties including a MOS input gate, high switching speed, low conduction voltage drop, high current carrying capability, and a high degree of robustness. Devices have drawn closer to the 'ideal switch', with typical voltage ratings of 600 - 1700 volts, on-state voltage of 1.7 - 2.0 volts at currents of up to 1000 amperes, and switching speeds of 200 - 500 ns. The availability of IGBTs has lowered the cost of systems and enhanced the number of economically viable applications. The insulated gate bipolar transistor (IGBT) combines the positive attributes of BJTs and MOSFETs. BJTs have lower conduction losses in the on-state, especially in devices with larger blocking voltages, but have longer switching times, especially at turn-off while MOSFETs can be turned on and off much faster, but their on-state conduction losses are larger, especially in devices rated for higher blocking voltages. Hence, IGBTs have lower on-state voltage drop with high blocking voltage capabilities in addition to fast switching speeds.

IGBTs have a vertical structure as shown in Fig. 2.1. This structure is quite similar to that of the vertical diffused MOSFET except for the presence of the p+ layer that forms the drain of the IGBT. This layer forms a p-n junction (labeled J1 in the figure), which injects minority carriers into what would appear to be the drain drift region of the vertical MOSFET. The gate and source of the IGBT are laid out in an inter-digitated geometry similar to that used for the vertical MOSFET.

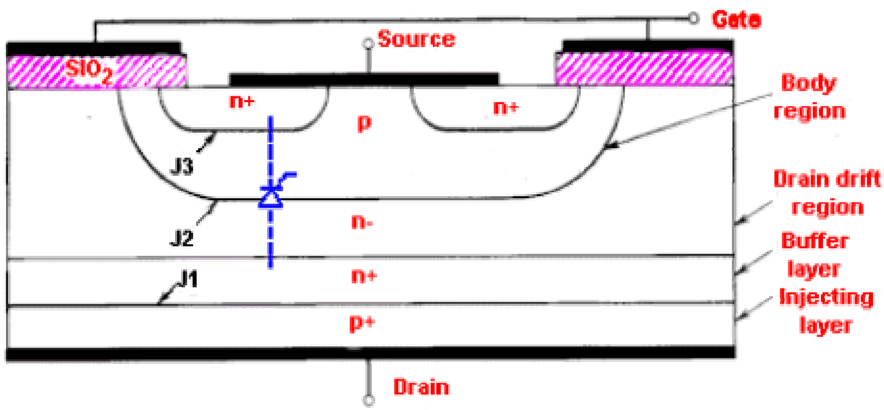



Fig. 2.1: Physical structure of an IGBT

The IGBT structure shown in Fig. 1 has a parasitic thyristor which could latch up in IGBTs if it is turned on. The n + buffer layer between the p + drain contact and the n + drift layer, with proper doping density and thickness, can significantly improve the operation of the IGBT, in two important respects. It lowers the on-state voltage drop of the device and, and shortens the turn-off time. On the other hand, the presence of this layer greatly reduces the reverse blocking capability of the IGBT. The circuit symbol for an n-channel IGBT is shown in Fig. 2.2

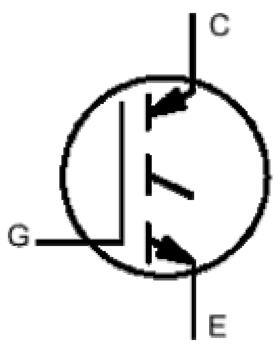



Fig. 2.2 IGBT circuit symbol

## 2.2 IGBTS SWITCHING CHARACTERISTICS

One of the main important performance features of any semiconductor switching device is its switching characteristics. Understanding the device switching characteristics greatly improves its utilization in the various applications.

The main performance switching characteristics of power semiconductor switching devices are the turn-on and turn-off switching transients in addition to the safe operating area (SOA) of the device.

Since most loads are inductive in nature, which subjects devices to higher stresses, the turn-on and turn-off transients of the IGBT are obtained with an inductive load test circuit as shown in Fig. 2.3. The load inductance is assumed to be high enough so as to hold the load current constant during switching transitions. The freewheeling clamp diode is required to maintain current flow in the inductor when the device under test (DUT) is turned off.

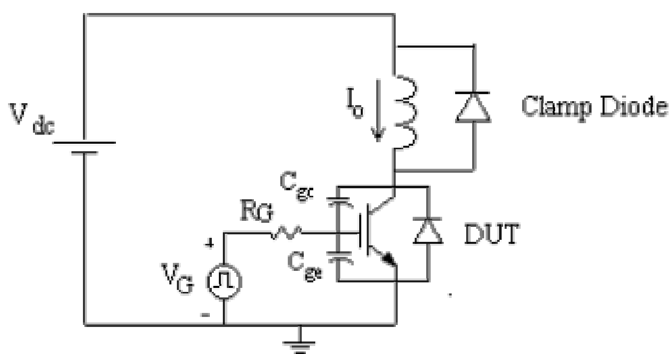



Fig. 2.3: Inductive load test circuit

## 2.3 TURN-ON TRANSIENTS

The turn-on switching transient of an IGBT with an inductive load is shown in Fig. 2.4. The turn-on switching transients of IGBTs are very similar to MOSFETs since the IGBT is essentially acting as a MOSFET during most of the turn-on interval. With gate voltage applied across the gate to emitter terminals of the IGBT, the gate to emitter voltage rises up in an exponential fashion from zero to $V_{GE(th)}$ due to the circuit gate resistance (RG) and the gate to emitter capacitance ($C_{ge}$). The Miller effect capacitance ($C_{gc}$) effect is very small due to the high voltage across the device terminals.

Beyond $V_{GE(th)}$, the gate to emitter voltage continues to rise as before and the drain current begins to increase linearly as shown above. Due to the clamp diode, the collector to emitter voltage remains at $V_{dc}$ as the IGBT current is less than Io. Once the IGBT is carrying the full load current but is still in the active region, the gate to emitter voltage becomes temporarily clamped to $V_{GE,Io}$, which is the voltage required to maintain the IGBT current at Io. At this stage, the collector to emitter voltage starts decreasing in two distinctive intervals $t_{fv1}$ and $t_{fv2}$. The first time interval corresponds to the traverse through the active region while the second time interval corresponds to the completion of the transient in the ohmic region.

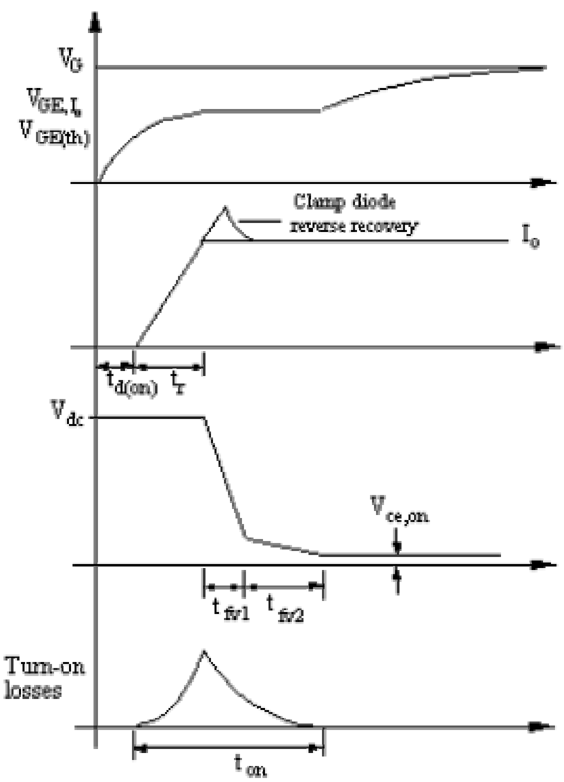



Fig. 2.4: IGBT turn-on switching transient with inductive load

During these intervals, the Miller capacitance becomes significant where it discharges to maintain the gate to source voltage constant. When the Miller capacitance is fully discharged, the gate to emitter voltage is allowed to charge up to $V_G$ and the IGBT goes into deep saturation. The resultant turn on switching losses are shown in the above figure. The on energy loss is approximately estimated via,

$$E_{on} = \frac{V_{dc} I_o}{2} t_{on}$$

The above switching waveforms are ideal in the since that the clamp diode reverse recovery effects are neglected. If these effects are included, an additional spike in the current waveform results as shown in the previous figure. As a result, additional energy losses will be incurred within the device.

## 2.4 TURN-OFF TRANSIENTS

The turn-off switching transients of an IGBT with an inductive load are shown in Fig. 2.5. When a negative gate signal is applied across the gate to emitter junction, the gate to emitter voltage starts decreasing in a linear fashion. Once the gate to emitter voltage drops below the threshold voltage ($V_{GE(th)}$), the collector to emitter voltage starts increasing linearly. The IGBT current remains constant during this mode since the clamp diode is off. When the collector to emitter voltage reaches the dc input voltage, the clamp diode starts conducting and the IGBT current falls down linearly. The rapid drop in the IGBT current occurs during the time interval $t_{fi1}$, which corresponds, to the turn-off of the MOSFET part of the IGBT (Fig. 2.5). The tailing of the collector current during the second interval $t_{fi2}$ is due to the stored charge in the n- drift region of the device. This is because the MOSFET is off and there is no reverse voltage applied to the IGBT terminals that could generate a negative drain current so as to remove the stored charge. The only way for stored charge removal is by recombination within the n- drift region. Since it is desirable that the excess carriers lifetime be large to reduce the on-state voltage drop, the duration of the tail current becomes long. This will result in additional switching losses within the device. This time increases also with temperature similar to the tailing effect in BJTs. Hence, a trade off between the on-state voltage drop and faster turn-off times must be made.

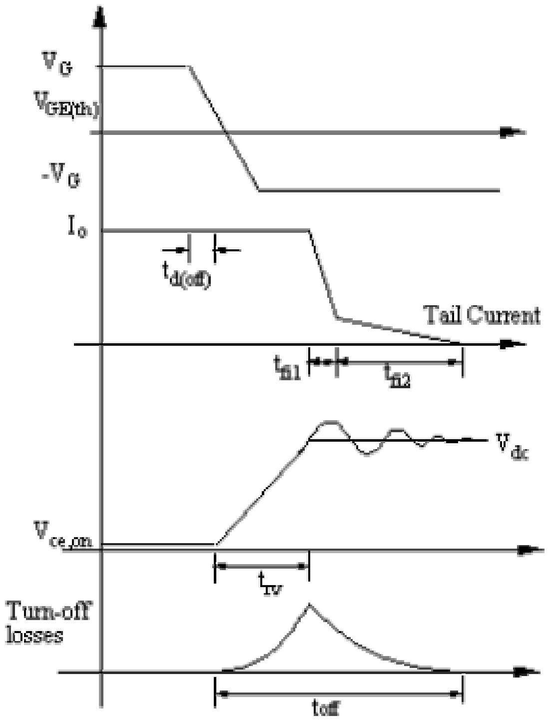



Fig. 2.5: IGBT turn-off switching transient with inductive load

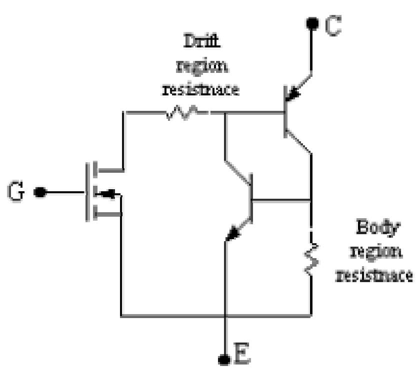



Fig. 2.6: equivalent circuit of the IGBT

The removal of stored charge can be greatly enhanced with the addition of an n+ buffer layer which acts as a sink for the excess holes and significantly shortens the tail time. This layer has a much shorter excess carrier lifetime that results in a greater recombination rate within this layer. The resultant gradient in hole density in the drift region causes a large flux of diffusing holes towards the buffer region which greatly enhances the removal rate of holes from the drift region and shortens the tail time. This device structure is referred to as Punch-Through (PT) IGBT while the structure without the n+ buffer region is referred to as Non Punch-Through (NPT) IGBT (Fig. 2.7).

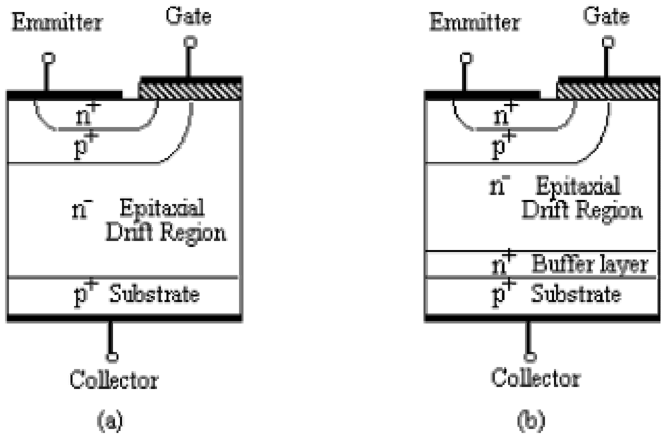



Fig. 2.7: (a) Non Punch Through (NPT) IGBT (b) Punch Through (PT) IGBT

The turn off energy loss, also shown in Fig. 3, can be evaluated in a similar fashion as the turn-on losses, namely,

$$E_{off} = \frac{V_{dc} I_o}{2} t_{off}$$

**2.5 IGBT SAFE OPERATING AREA**

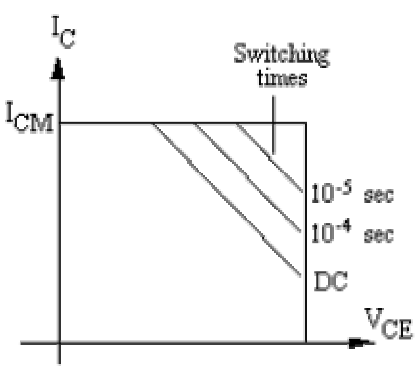
The safe operating area (SOA) of a power semiconductor device is a graphical representation of the maximum operational voltage and current limits (i-v) of the device subjected to various constraints. The forward bias safe operating area (FBSOA) and the reverse bias safe operating area (RBSOA) represent the device SOA with the gate emitter junction forward biased or reverse biased, respectively.

The IGBT has robust SOA during both turn-on and turn off. The FBSOA, shown in Fig. 6(a), is square for short switching times, similar to that of power MOSFETs. The IGBT is thermally limited for longer switching times as shown in the FBSOA figure.

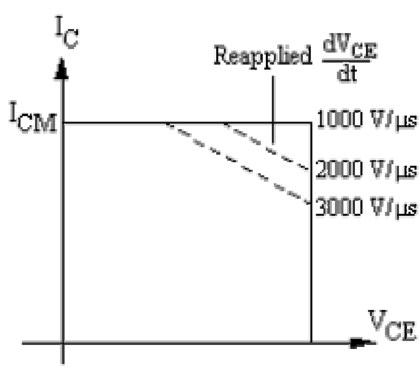
The RBSOA of IGBTs, shown in Fig. 2.8(b), is different than the FBSOA. The upper half corner of the RBSOA is progressively cut out which reduces the RBSOA as the rate of change of the collector to emitter voltage across the device, $dV_{ce}/dt$, is increased. The RBSOA is reduced as the

$dV_{ce}/dt$ is increased to avoid latch up within the device. This condition exists when higher values of $dV_{ce}/dt$ are applied may give to the rise to a pulse of forward decaying current in the body region of the device that acts as a pulse of gate current that can turn on the device. Fortunately, the $dV_{ce}/dt$ values that would cause latch up in IGBTs are much higher compared to other devices.

The maximum value of ICM is set to avoid latch up which is determined based on the dynamic latch up condition. In addition, a maximum $V_{GE}$ voltage is specified in order to limit the current during a fault condition to ICM by forcing the device out of the on-state into the active region where the current becomes constant regardless of the drain to source voltage. The IGBT must be turned off under these conditions as quickly as possible to avoid excessive dissipation. The avoidance of latch up and the continuous gate control over the collector current are very desirable features.




(a)




(b)

Fig. 2.8: (a) FBSOA (b) RBSOA of an IGBT

## 2.6 IGBT GATE DRIVE REQUIREMENTS

IGBTs are voltage controlled devices and require gate voltage to establish collector-to-emitter conduction. Recommended gate drive circuitry includes substantial ion and off biasing as shown in Fig. 2.9.

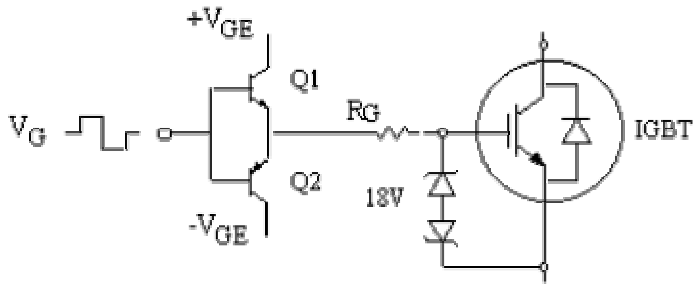



Fig. 2.9: Typical gate drive circuitry

Due to the large input gate-to-emitter capacitance of IGBTs, MOSFET drive techniques can be used. However, the off biasing needs to be stronger. A +15 V positive gate drive is normally recommended to guarantee full saturation and limit short circuit current. A negative voltage bias is used to improve the IGBT immunity to collector-to-emitter dv/dt injected noise and reduce turn-off losses as shown in Fig. 2.10.

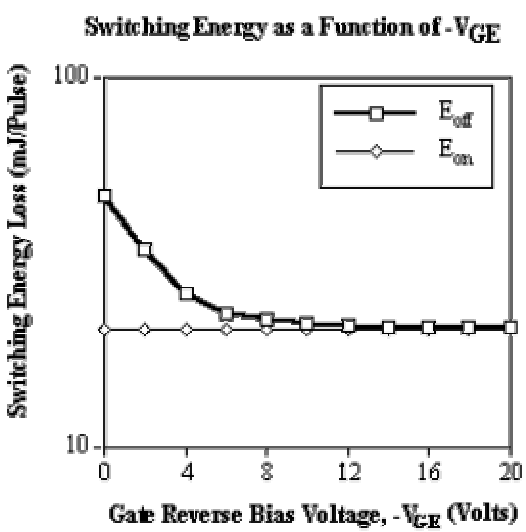



Fig. 2.10: Effect of negative bias on turn off losses

The value of the gate resistance has a significant impact on the dynamic performance of IGBTs. A smaller gate resistance charges and discharges the IGBT input capacitance faster reducing switching times and switching losses and improving immunity to dv/dt turn-on (Fig. 2.11). However, a small gate resistance can lead to oscillations between the IGBT input capacitance and the parasitic lead inductance.

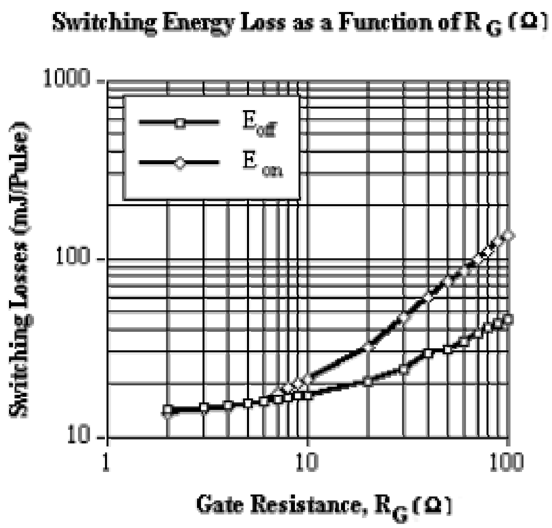



Fig. 2.11: The IGBT switching losses as a function of gate resistance, RG

The minimum peak current capability of the gate drive power supply and the average power required are given by,

$$I_{G(pk)} = \pm \frac{\Delta V_{GE}}{R_G}$$

Pavg = $V_{GE}$ . $Q_G$. fs

where,

DVGE = $V_{GE\_on} + |V_{GE\_off}|$

$Q_G$ = total gate charge (per manufacturer. spec.)

fs = switching frequency

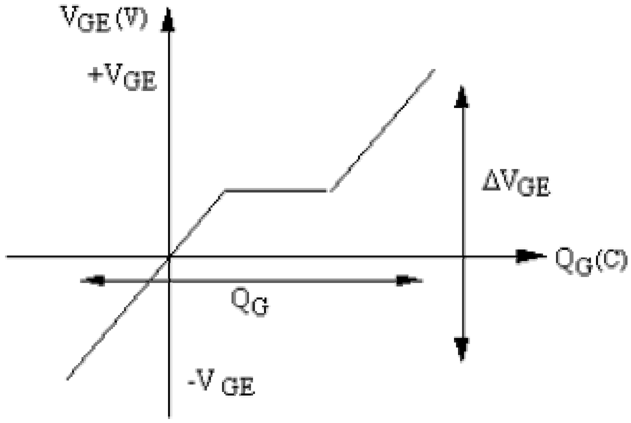



Fig. 2.12: Total IGBT Gate Charge during switching

In many applications, the gate drive circuitry needs to be isolated from the control circuit to provide the level shifting and improve noise immunity. The isolation requirements can be met by using pulse gate transformers (Fig. 2.13) or optical isolation.

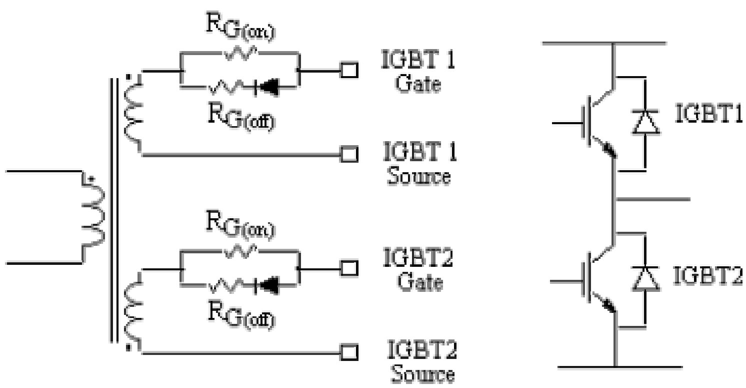



Fig. 2.13: Typical Bipolar IGBT gate drive using gate pulse transformers

In bipolar applications, separate turn-on and turn-off gate resistors are used to prevent cross conduction of an IGBT pair (Fig. 2.14). With opto-isolation, an isolated power supply is required to provide the gate power to the IGBT.

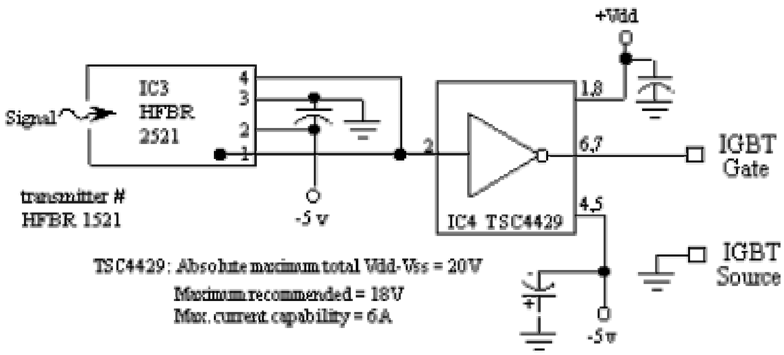



Fig. 2.14: Typical opto-isolation gate drive

**Gate drive Layout Considerations**

1. Minimize parasitic inductance between the driver output stage and the IGBT (minimizing the loop area)
2. Minimize noise coupling via proper shielding techniques
3. Utilize gate clamp protections (TVS) to minimize over voltage across gate terminals
4. Utilize twisted pairs, preferably shielded, for indirect connection between the driver and the IGBT
5. With OPTO coupling isolation, a minimum of 10,000 V/ms transient immunity must be provided (in hard switching applications)

Chapter 3

# THYRISTOR

# THYRISTORS

## 3.1 SILICON CONTROLLED RECTIFIERS

Thyristor is a four layer, three junction p-n-p-n semiconductor switching device. It has three terminals; anode, cathode and gate. The four layers of alternate p-type and n-type semiconductors forming three junctions $J_1$, $J_2$, and $J_3$. The terminal connected to outer p region is called Anode (A), the terminal connected to outer n region is called cathode (C) and that connected to inner p region is called gate (G).For large current applications, thyristors need better cooling which is achieved to a great extent by mounting them onto heat sinks

## 3.2 THYRISTOR TURN ON

The thyristor is turned on by increasing the anode current. This can be accomplished in the following ways.

***Thermals***

If the temperature of a thyristor is high, there is an increase in the number of electron-hole pairs, which increases the leakage currents. This increase in currents causes $\alpha_1$ and $\alpha_2$ to increase. Due to regenerative action ($\alpha_1 + \alpha_2$ ) may tend to unity and thyristor may be turned on.

***Light***

If light is allowed to strike the junctions of a thyristor, the electron-hole pairs increase; and the thyristor may be turned on. The light-activated thyristors are turned on by allowing light to strike silicon wafers.

***High voltage***

If the forward anode –to-cathode voltage is greater than the forward breakdown voltage $V_{BO}$ , sufficient leakage current flows to initiate regenerative turn on. This type of turn-on may be destructive and should be avoided.

***Gate current***

If a thyristor is forward biased, the injection of gate current by applying a positive gate voltage between the gate and the cathode terminals turns on the thyristor. As the gate current is increased, the forward blocking voltage is decreased.

**3.3 CIRCUIT DESCRIPTION**

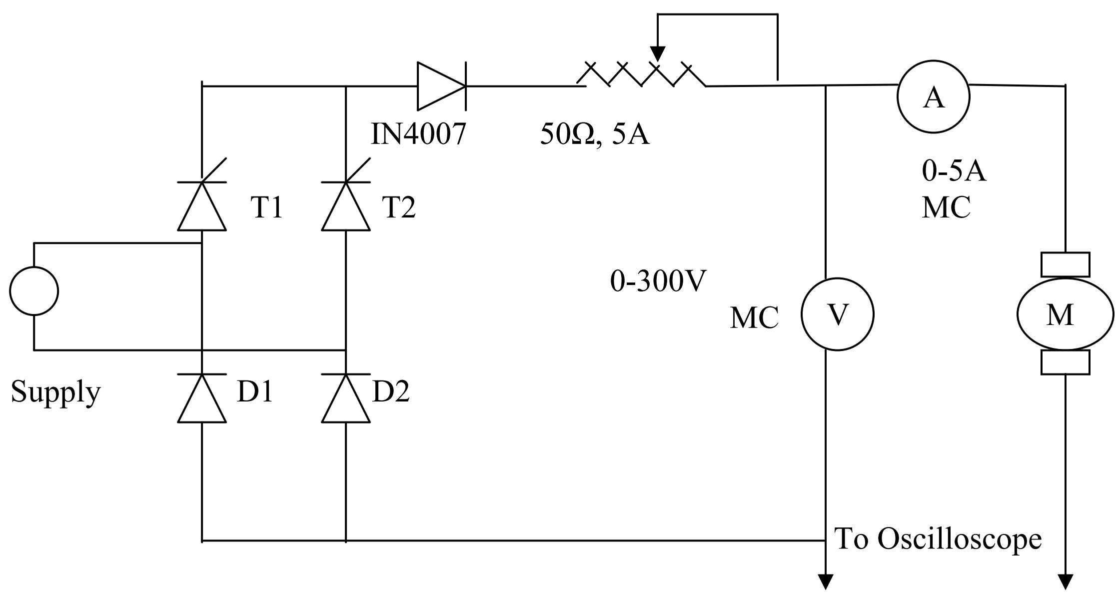



Fig 3.1: Half Controlled Bridge rectifier

The general arrangement for the speed control of the shunt motor is shown above. The firing angle control of converter regulates the armature voltage applied to the dc motor. Thus the variation of the delay angle of converter gives speed control below the base speed as we are dealing with armature circuit only. Similarly if we deal with field circuit it will give speed above the base speed only. This converter is commonly used in applications up to 15kW.

During the positive half cycle T1 is forward biased. When T1 is fires at wt= ,the load is connected to the input supply through T1 and D2.During the negative half cycle of input voltage, T2 is forward biased and the firing of T2 occurs and the load is connected to the supply through T2 and D1. In the following experiment the armature current ($I_a$) is assumed constant. The firing angle was varied in small steps and the armature voltage current and speed were measured and corresponding graphs were plotted.

Chapter 4

# DC MOTOR

# DC MOTOR

## 4.1 INTRODUCTION

DC motors are used extensively in adjustable-speed drives and position control applications. Their speeds below the base speed can be controlled by armature-voltage control. Speeds above the base speed are obtained by field-flux control. As speed control methods for DC motors are simpler and less expensive than those for the AC motors, DC motors are preferred where wide-speed range control is required. Phase controlled converters provide an adjustable dc voltage from a fixed ac input voltage. DC choppers also provide dc output voltage from a fixed dc input voltage. The use of phase controlled rectifiers and dc choppers for the speed control of dc motors have revolutionized the modern industrial controlled applications. DC drives are classified as follows:

a) single phase DC drives
b) three phase DC drives
c) chopper drives

## 4.2 APPLICATION OF DC SHUNT MOTOR

1. For a given field current in a shunt motor, the speed drop from no-load to full load is invariably less than 6% to 8%. In view of this, the shunt motor is termed as a constant speed motor. Therefore, for constant speed drives in industry DC shunt motors are employed.
2. When constant speed service at low speeds is required, DC shunt motors are preferred over synchronous motors.
3. When the driven load requires a wide range of speed control, both below and above the base speed, a DC shunt motor is employed. Eg: Lathes
4. DC shunt motor can be used as a separately excited motor, if the field winding is disconnected from armature and connected to a external voltage source.

Chapter 5

# OBSERVATIONS

**GRAPH 5.1**

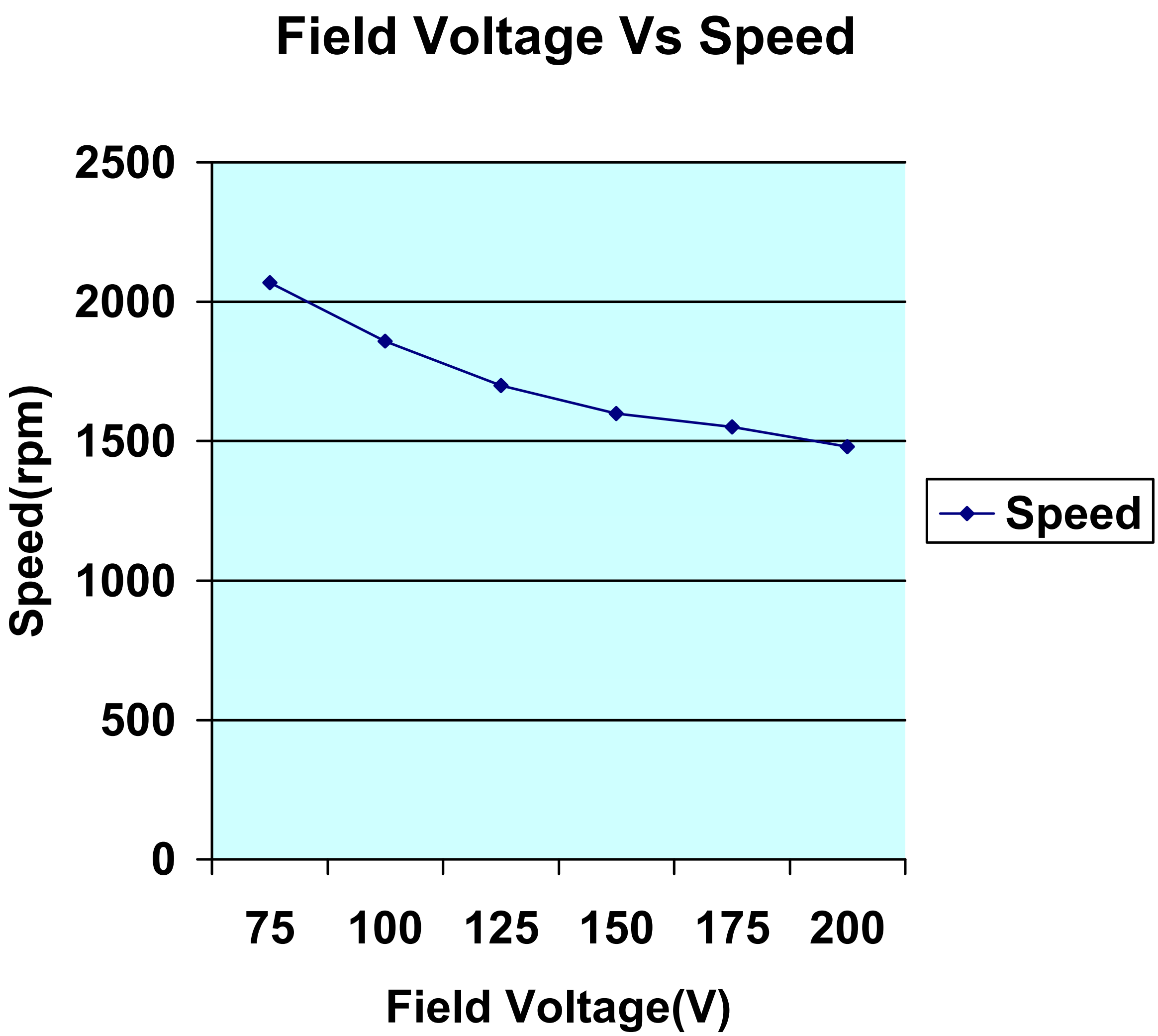



**Table 5.1 FIELD CONTROL at Va=200V**

| S.No | $V_f$ (Volts) | N (rpm) |
|---|---|---|
| 1 | 75 | 2070 |
| 2 | 100 | 1860 |
| 3 | 125 | 1700 |
| 4 | 150 | 1600 |
| 5 | 175 | 1550 |
| 6 | 200 | 1480 |

# OBSERVATION 5.1

## FIELD VOLTAGE VS SPEED

**DESCRIPTION:**

Here armature voltage $V_a$ was kept constant at 200V. The field voltage was varied and the corresponding speed was noted down.

**CONCLUSION:**

The graph was plotted and it was observed that the speed of motor is decreasing as the field voltage increases.

We also observed that using field flux method the speed greater than base speed was achieved.

# OBSERVATION 5.2

## ARMATURE VOLTAGE VS SPEED

**DESCRIPTION:**

Here field voltage $V_f$ was kept constant at 200V. The armature voltage was varied and the corresponding speed was noted down.

**CONCLUSION:**

The graph was plotted and its was observed that the speed of the motor was varies directly with increase in armature voltage.

We also observed that using armature control method the speed control was obtained for speed below the base speed.

**GRAPH 5.2**

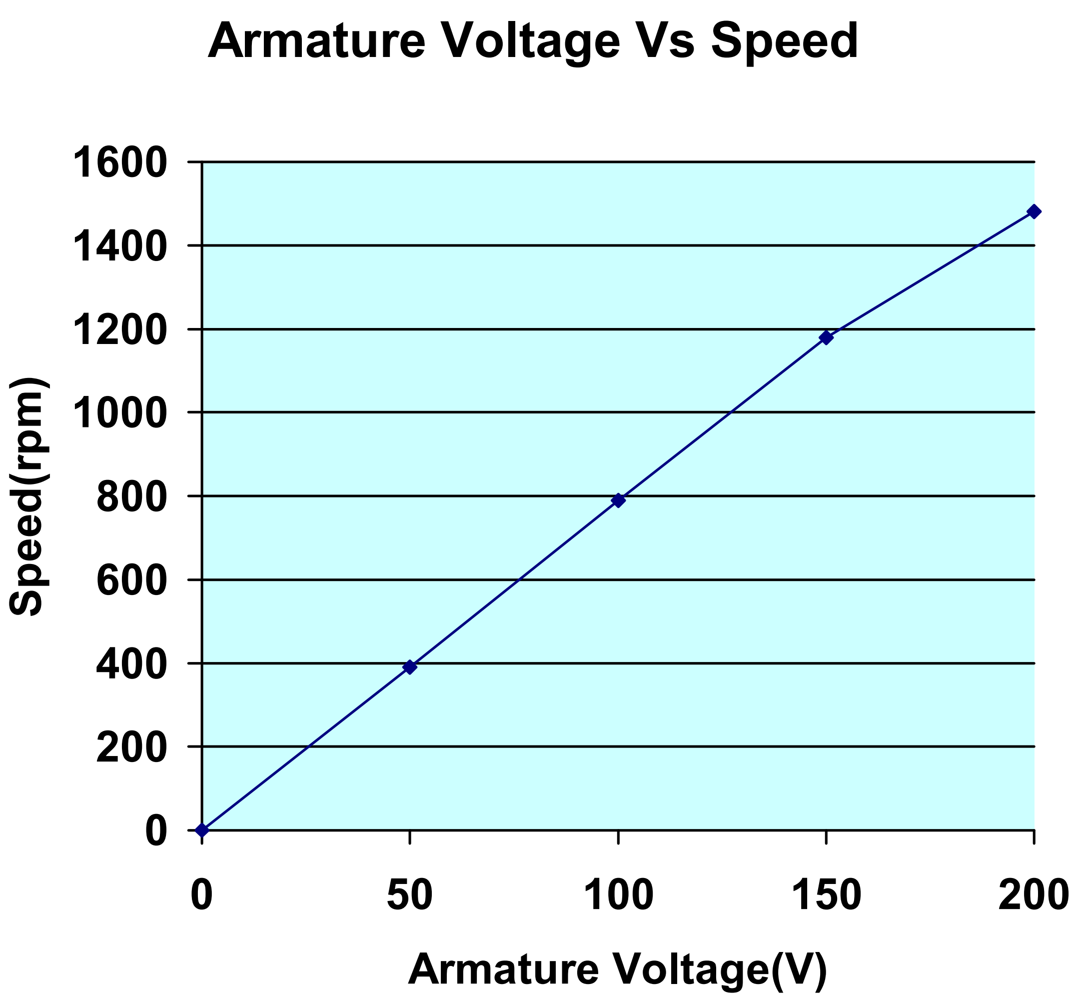



**Table 5.2 ARMATURE CONTROL $V_F$=200V**

| S.No | Armature Voltage(V) | Speed(rpm) |
|---|---|---|
| 1 | 0 | 0 |
| 2 | 50 | 390 |
| 3 | 100 | 790 |
| 4 | 150 | 1180 |
| 5 | 200 | 1480 |

**GRAPH 5.3**

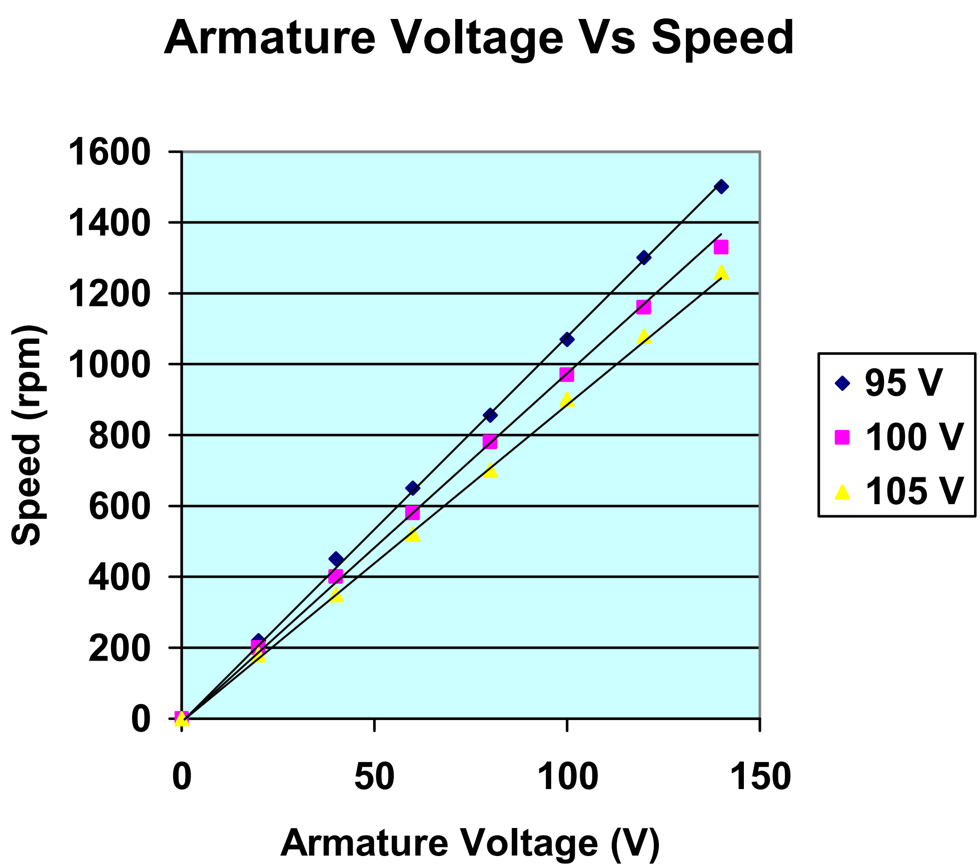



**Table 5.3** **ARMATURE VOLTAGE VS SPEED AT DIFFERENT VALUES OF $V_F$**

| S.No | Armature Voltage | Field Voltage $V_f$ (V) | | |
|---|---|---|---|---|
| | | 95V | 100V | 105V |
| 1 | 0 | 0 | 0 | 0 |
| 2 | 20 | 220 | 200 | 180 |
| 3 | 40 | 450 | 400 | 350 |
| 4 | 60 | 650 | 580 | 520 |
| 5 | 80 | 855 | 780 | 700 |
| 6 | 100 | 1070 | 970 | 900 |
| 7 | 120 | 1300 | 1160 | 1080 |
| 8 | 140 | 1500 | 1300 | 1260 |

## OBSERVATION 5.3

### ARMATURE VOLTAGE VS SPEED

### DESCRIPTION:

Here field voltage $V_f$ was kept constant at three different values and for varying armature voltage $V_a$ speed was noted down.

### CONCLUSION:

The speed increases with increase in armature voltage $V_a$. But it was observed that for lower field voltage the base speed is attained at comparatively lower armature voltage.

## OBSERVATION 5.4

### ARMATURE VOLTAGE VS SPEED( FOR VARIOUS LOADING CONDITIONS)

### DESCRIPTION:

Here field voltage $V_f$ was kept constant armature voltage $V_a$ was varied under various loading conditions i.e. capacitive, resistive and inductive and the corresponding speeds were noted down.

### CONCLUSION:

The speed increases with increase in armature voltage $V_a$. But it was observed that for inductive load the base speed is attained at comparatively lower armature voltage with respect to resistive and capacitive load. The base speed is attained at comparatively lower armature voltage for resistive load. The capacitive load gives better control on armature voltage.

**GRAPH 5.4**

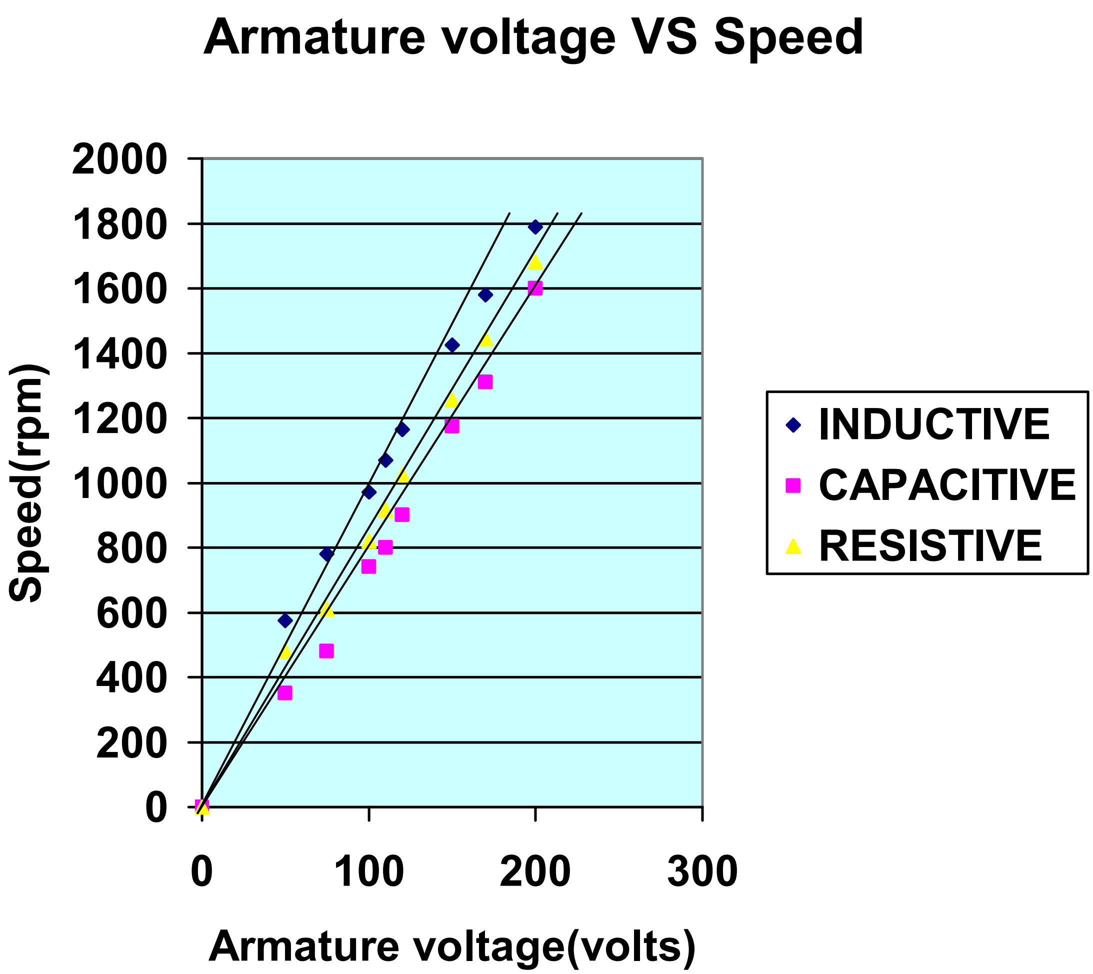



**Table 5.4** **ARMATURE VOLTAGE VS SPEED (VARIOUS LOADING CONDITIONS)**

| S.No | Voltage (V) | Speed (rpm) | | |
|---|---|---|---|---|
| | | Inductive | Resistive | Capacitive |
| 1 | 0 | 0 | 0 | 0 |
| 2 | 50 | 575 | 480 | 350 |
| 3 | 75 | 780 | 610 | 480 |
| 4 | 100 | 972 | 820 | 740 |
| 5 | 110 | 1070 | 915 | 800 |
| 6 | 120 | 1164 | 1025 | 900 |
| 7 | 150 | 1425 | 1255 | 1175 |
| 8 | 170 | 1580 | 1445 | 1310 |
| 9 | 200 | 1790 | 1680 | 1600 |

**GRAPH 5.5**

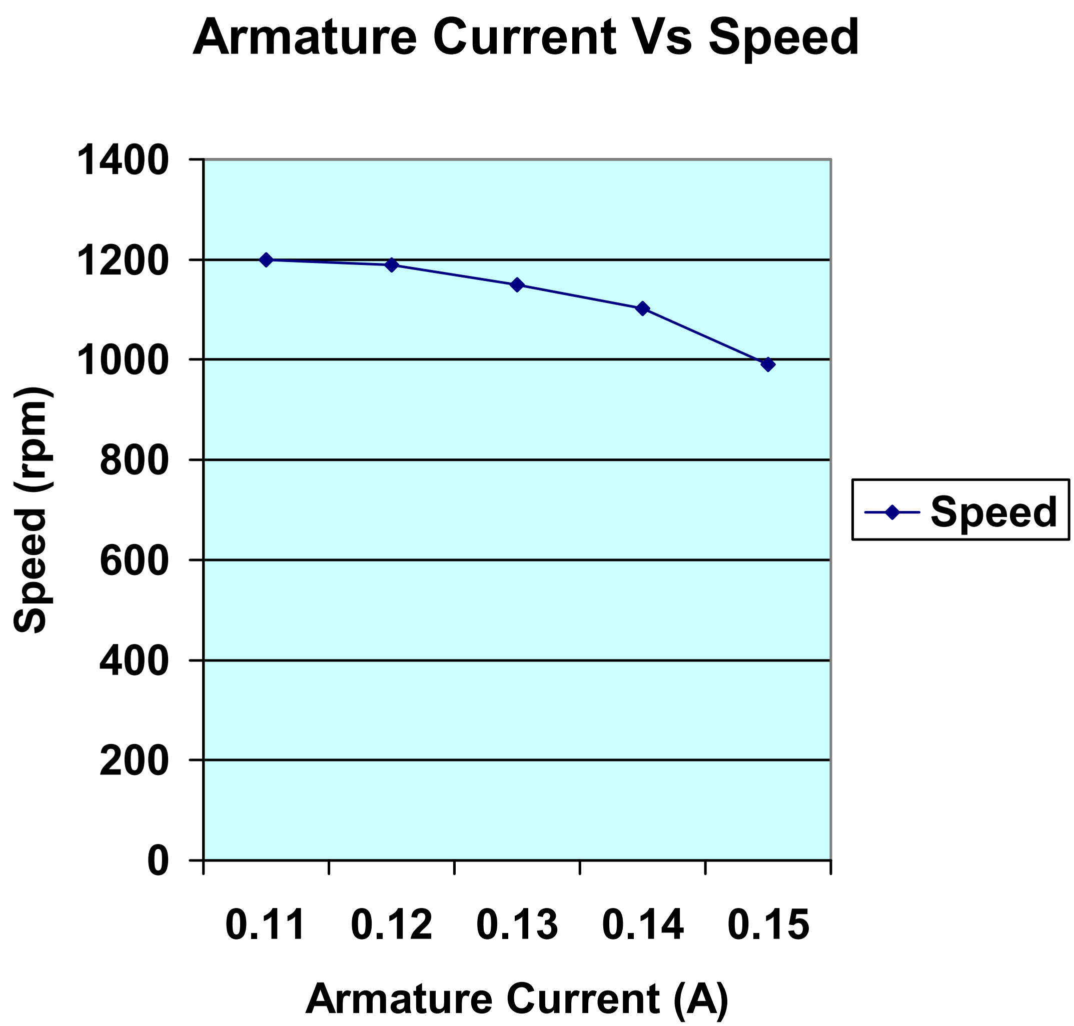



**Table 5.5 ARMATURE CURRENT VS SPEED**

| S.No | Armature Current (A) | Speed (rpm) |
|---|---|---|
| 1 | 0.11 | 1200 |
| 2 | 0.12 | 1190 |
| 3 | 0.13 | 1150 |
| 4 | 0.14 | 1103 |
| 5 | 0.15 | 990 |

## OBSERVATION 5.5

### ARMATURE CURRENT VS SPEED

### DESCRIPTION:

Here armature current $I_a$ was varied and the corresponding speeds were noted down.

### CONCLUSION:

It was found that speed decreases with increase in armature current $I_a$. It shows drooping characteristics

## OBSERVATION 5.6

### ARMATURE VOLTAGE VS SPEED( COMPARISON BETWEEN IGBT AND SCR)

### DESCRIPTION:

Here the field voltage $V_f$ was kept constant the armature voltage $V_a$ was varied and the speed was noted accordingly for IGBT and SCR.

### CONCLUSION:

It was observed that the slop of the graph for IGBT is less in comparison to that of SCR.The base speed is attained at comparatively lower armature voltage for SCR as compared to IGBT. Hence IGBT gives better control over speed.

**GRAPH 5.6**

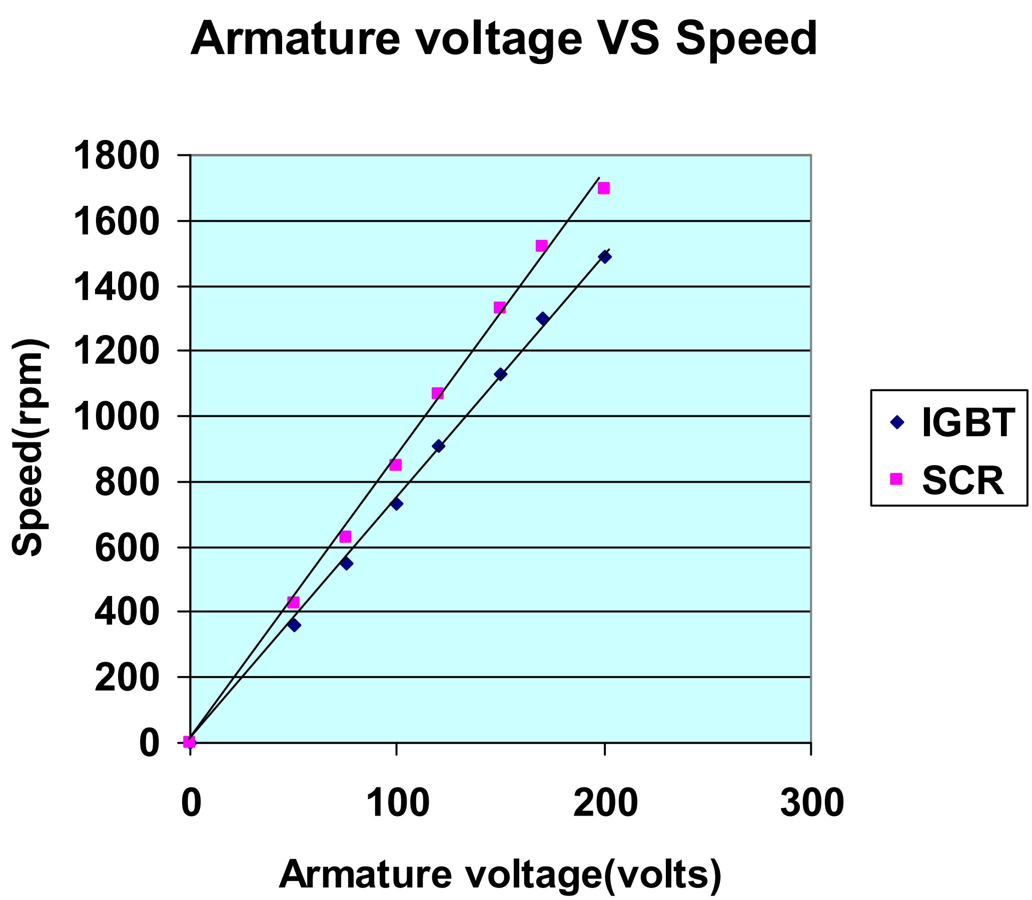



**Table 5.6 ARMATURE VOLTAGE VS SPEED**

| S.No | Armature Voltage (V) | Speed (rpm) | |
|---|---|---|---|
| | | IGBT | Thyristor |
| 1 | 0 | 0 | 0 |
| 2 | 50 | 360 | 430 |
| 3 | 75 | 550 | 630 |
| 4 | 100 | 734 | 850 |
| 5 | 120 | 910 | 1070 |
| 6 | 150 | 1130 | 1330 |
| 7 | 170 | 1300 | 1520 |
| 8 | 200 | 1490 | 1598 |

Chapter 6

# CONCLUSION

## 6.1 COMPARISON BETWEEN IGBT AND SCR

| | IGBT | SCR |
|---|---|---|
| 1.Gate- Drive Requirements | Lower | Higher |
| 2.Switching Losses | Less | More |
| 3. Snubber Circuit Requirements | Small | Large |
| 4. Efficiency | More efficient being smaller , lighter and generate fewer harmonics | Less efficient being bulky and generate more harmonics |
| 5. Switching Speed | Faster than BJT but lesser than MOSFET | Slower than IGBT & MOSFET |
| 6. Input Impedance | High | Low |
| 7. Second Breakdown Voltage | Absent | Present |
| 8. Control Parameter | Voltage Controlled device | Current Controlled device |
| 9. Cost | Costly | Cheap |

Speed control of dc motor motor using IGBT based Four Quadrant Chopper drive was carried out and following conclusions were made

1. Speed varies directly with armature voltage by keeping field voltage constant.
2. Speed varies inversely with field voltage by keeping armature voltage constant.
3. Armature voltage control gives the speed below the base speed whereas field control gives the speed control above the base speed.
4. Armature current vs Speed at constant flux gives a drooping characteristic. Though it should have been a straight line parallel to x-axis but due to saturation effect there is slight decrease in speed and shows a drooping characteristics.
5. The IGBT based circuit gives smoother control over the entire speed range as compared with the SCR based circuit.
6. IGBTs feature many desirable properties including a MOS input gate, high switching speed, low conduction voltage drop, high current carrying capability, and a high degree of robustness.
7. Devices have drawn closer to the 'ideal switch', with typical voltage ratings of 600 - 1700 volts, on-state voltage of 1.7 - 2.0 volts at currents of up to 1000 amperes, and switching speeds of 200 - 500 ns.
8. The availability of IGBTs has lowered the cost of systems and enhanced the number of economically viable applications. .
9. The insulated gate bipolar transistor (IGBT) combines the positive attributes of BJTs and MOSFETs.  BJTs have lower conduction losses in the on-state, especially in devices with larger blocking voltages, but have longer switching times, especially at turn-off while MOSFETs can be turned on and off much faster, but their on-state conduction losses are larger, especially in devices rated for higher blocking voltages. Hence, IGBTs have lower on-state voltage drop with high blocking voltage capabilities in addition to fast switching speeds.

Chapter 7

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