NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronic Engineering

Analysis of the Converter for Controlling Synchronous Motors by PSIM

Graduation Project
EE- 400

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Dedicated To My Father
Mr. Orhan Kayserilioğlu

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ABSTRACT

Nowadays, the analysis and design of complex power electronic systems such as motor drives are usually done using a modern simulation software which can provide accurate predictions of the system's behavior in reality. Consequently, computer modeling of such systems at a desired level of accuracy becomes an essential part of the design process.

A satisfying system model usually serves as a prototype for the system behavior simulations, as well as the small signal analysis and control design.

Electronic controls are examined with some of the basic principles. In describing the various methods of control, the behavior of power circuits are described in this project. Consequently, the power electronics control and switching devices are covered here.

The work presented within this project provides an analysis of special thyristor system as a converter for controlling motors. Additionaly, real-life application using these comparison devices are presented.

There are software programs that are used in power electronics for control of motors and analysis of the systems. The software program PSIM is used as a tool in the project for the analysis and the implementation.

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Synchronous motors are available in sub-fractional self-excited sizes to high-horsepower direct-current excited industrial sizes. In the fractional horsepower range, most synchronous motors are used where precise constant speed is required. In high horsepower industrial sizes, the synchronous a motor has two important properties. First of all it is a highly efficient means of converting ac energy to work. Secondly, it can operate at leading or unity power factor and thereby provide power-factor correction.

The power semiconductor devices, such as the diode, thyristor, triac and power transistor are used in power applications as switching devices. The development of theory and application relies heavily on waveforms and transient responses, which distinguishes the subject of power electronics form many other engineering studies.

Power electronics is popular for technical as well as economical reasons. Nowadays, electronic power generation, transformation, transmission, and distribution are in AC, but almost all the terminal equipment used in industries, laboratories, locomotion, agriculture, and households require DC power. In order to satisfy these requirements, easy conversion of AC power to DC power is essential. The conversion of AC to DC power at different frequencies and DC to AC power can be effected through power electronics in a very dependable and economic manner.

In this project, the converter circuit is analysed to control a synchronous motor using PSIM software as a tool.

The project consists of 5 chapters.

In Chapter 1 and Chapter 2, a brief introduction about the power electronics and power electronic devices will be given.

In Chapter 3, power electronic converters will be explained with their specific properties.

In Chapter 4, motor drives and motor drives properties will be discussed.

In Chapter 5, the PSIM simulation and results will be presented.

In Conclusion part, the results of the presented work will be summarized.

CHAPTER ONE
INTRODUCTION TO POWER ELECTRONICS

1.1 Overview

In this chapter, an introduction to power electronics, power electronics application areas with the devices used in power electronics will be given briefly.

1.2 Introduction to Power Electronics

For many years, power electronics in the high-power area was performed with extremely slow semiconductor switches. These switches, including the thyristor and the Gate Turn-Off (GTO) thyristor, had the capacity to handle very high voltages and currents but lacked the ability to perform high frequency switching. Low-power converters, such as computer power supplies and low horsepower motor drives, have employed high-frequency switching for years and have benefited from very nice output waveforms, good control dynamic performance, and many other advantages compared to low frequency switching. Recent improvements in high-power semiconductor technology has brought switching performance similar to that of the low-power MOSFETs and IGBTs to the high-power area through the advancement of the IGBT's ratings to create the High Voltage IGBT (HVIGBT) and the development of new GTO-derived devices including the Integrated Gate Commutated Thyristor (IGCT) and the Emitter Turn-Off (ETO) thyristor. These new devices all feature high switching speed and the capability to turn off without the requirement for a turn-off snubber. With these new device technologies the high power field of power electronics can realize dramatic improvements in the performance of systems for utility applications and motor drives.

However, with these high-speed switches come new issues relating to noise, protection, performance of diodes, and thermal management in high-frequency applications.

1.3 What is Power Electronics ?

Power electronics is the control and conversion of electrical power by power semiconductor devices where in these devices operate as switches. Advent of silicon-controlled rectifiers, abbreviated as SCRs, led to the development of a new area of application called the power electronics. Prior to the introduction of SCRs, mercury-arc rectifiers were used for controlling electrical power, but such rectifier circuits were part of industrial electronics and the scope for applications of mercury-arc rectifiers was limited. Once the SCRs were available, the application area spread to many fields such as drives, power supplies, aviation electronics, high frequency inverters and power electronics originated.

Power electronics has applications that span the whole field of electrical power systems, with the power range of these applications extending from a few VA/Watts to several MVA / MW.

The main task of power electronics is to control and convert electrical power from one form to another. The four main forms of conversion are:

Rectification referring to conversion of ac voltage to dc voltage,

a) DC-to-AC conversion,

b) DC-to DC conversion and

c) AC-to-AC conversion.

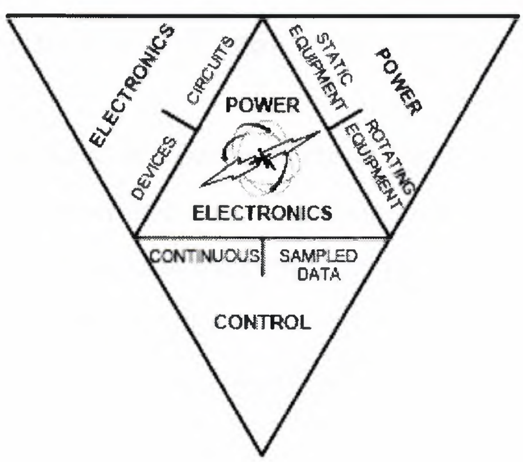


Figure 1.1 Power Electronics.

"Electronic power converter" is the term that is used to refer to a power electronic circuit that converts voltage and current from one form to another. These converters can be classified as:

a) Rectifier converting an AC voltage to a DC voltage,

b) Inverter converting a DC voltage to an AC voltage,

Chopper or a switch-mode power supply that converts a dc voltage to another dc voltage, and Cycloconverter and cycloinverter converting an ac voltage to another ac voltage.

In addition, SCRs and other power semiconductor devices are used as static switches.

1.4 Additional Insights into Power Electronics

There are several striking features of power electronics, the foremost among them being the extensive use of inductors and capacitors. In many applications of power electronics, an inductor may carry a high current at a high frequency. The implications of operating an inductor in this manner are quite a few, such as necessitating the use of litz wire in place of single-stranded or multi-stranded copper wire at frequencies above 50 kHz, using a proper core to limit the losses in the core, and shielding the inductor properly so that the fringing that occurs at the air-gaps in the magnetic path does not lead to electromagnetic interference.

Usually the capacitors used in a power electronic application are also stressed. It is typical for a capacitor to be operated at a high frequency with current surges passing through it periodically. This means that the current rating of the capacitor at the operating frequency should be checked before its use. In addition, it may be preferable if the capacitor has self-healing property. Hence an inductor or a capacitor has to be selected or designed with care, taking into account the operating conditions, before its use in a power electronic circuit.

In many power electronic circuits, diodes play a crucial role. A normal power diode is usually designed to be operated at 400 Hz or less. Many of the inverter and switch-mode power supply circuits operate at a much higher frequency and these circuits need diodes that turn ON and OFF fast. In addition, it is also desired that the turning-off process of a diode should not create undesirable electrical transients in the circuit. Since there are several types of diodes available, selection of a proper diode is very important for reliable operation of a circuit.

Analysis of power electronic circuits tends to be quite complicated, because these circuits rarely operate in steady-state. Traditionally steady-state response refers to the state of a circuit characterized by either a dc response or a sinusoidal response. Most of the power electronic circuits have a periodic response, but this response is not usually sinusoidal. Typically, the repetitive or the periodic response contains both a steady-state part due to the forcing function and a transient part due to the poles of the network. Since the responses are nonsinusoidal, harmonic analysis is often necessary. In order to obtain the time response, it may be necessary to resort to the use of a computer program.

Power electronics is a subject of interdisciplinary nature. To design and build control circuitry of a power electronic application, one needs knowledge of several areas, which are,

a) Design of analogue and digital electronic circuits, to build the control circuitry.

b) Microcontrollers and digital signal processors for use in sophisticated applications.

Many power electronic circuits have an electrical machine as their load. In ac variable speed drive, it may be a reluctance motor, an induction motor or a synchronous motor. In a dc variable speed drive, it is usually a dc shunt motor.

In a circuit such as an inverter, a transformer may be connected at its output and the transformer may have to operate with a nonsinusoidal waveform at its input.

A pulse transformer with a ferrite core is used commonly to transfer the gate signal to the power semiconductor device. A ferrite-cored transformer with a relatively higher power output is also used in an application such as a high frequency inverter.

Many power electronic systems are operated with negative feedback. A linear controller such as a PI controller is used in relatively simple applications, whereas a controller based on digital or state-variable feedback techniques is used in more sophisticated applications.

Computer simulation is often necessary to optimize the design of a power electronic system. In order to simulate, knowledge of software package such as MATLAB and the know-how to model nonlinear systems may be necessary.

The study of power electronics is an exciting and a challenging experience. The scope for applying power electronics is growing at a fast pace. New devices keep coming into the market, sustaining development work in power electronics.

1.5 Power Electronics Application Areas

The modern world is steadily becoming more and more reliant on high technology electronics and computers, which has led to a dramatic increase in the amount of power that is processed by power electronic converters. Although the majority of the power electronic converters in the world today by volume are in the low or medium power range, a significant amount of power is processed by a smaller number of very high power level converters. This trend will increase with the deregulation of the electric power industry, as the grid becomes full of interconnections between the numerous small suppliers rather than a few large utilities.

The main uses of high-power converters can be considered mainly to be either utility applications or motor drives, with only a few specialized applications, which are usually military in nature, falling outside of these two categories.

1.5.1 Utility Applications

The deregulated power system which is currently beginning to take form will require many high-power, high-performance power electronic systems. Distributed generation, with technologies including but not limited to windmills, photovoltaics, and microturbines, will require many multi-megawatt converters to interface these small localized generation sources to the grid. Many distributed generation technologies generate power in the form of DC such as photovoltaics, or variable voltage and variable frequency AC, since the wind can not be easily regulated to control the speed of a windmill. However, the electric power grid is fixed voltage, fixed frequency AC, so power conversion technology must be used in order for this technology to be effective.

In addition, increased demands on the utilities due to the sensitivity of high technology industries to power quality will require high-power converters to serve as filters and compensators on the grid. Semiconductor fabrication is particularly sensitive to power quality, where any problems with the incoming power can disrupt the process. Obviously this will become worse as finer pitch chips are made and the load on the utility increases, so the power quality is becoming a significant issue.

The concept is essentially to store energy at night when the generators are running with excess capacity available, and to release the energy during the day when the load exceeds the capacity of the generation. This allows for more efficient use of the generation equipment, so it can generate at the same capacity all the time, but requires significant efforts in both the method of storing the energy and the Power Conversion System (PCS) interfacing the storage system to the grid.

Examples of energy storage systems include pumped water, conventional batteries, flow batteries, Superconducting Magnetic Energy Storage (SMES), flywheels, capacitors, ultracapacitors, fuel cells, and many more are being developed constantly. The energy storage field is growing quickly and will require good power electronics in order to become practical.

High Voltage DC (HVDC) transmission, used to transmit large amounts of power over long distances with low losses, requires power electronic converters to interface the DC transmission line to the AC grid. Conventional HVDC DC-AC inverters have been linecommutated thyristor inverters, which have very high power ratings (up to several gigawatts) but produce very poor AC waveforms. Therefore bulky and expensive passive filters are used in order to clean the power they provide, but this technique is far from optimal. Ideally a highfrequency inverter would be used which would require almost no filtering in order to produce good quality.

1.5.2 Motor Drives

Motor drives are being built in always escalalating horsepower ratings, and with increased performance requirements. Many pieces of heavy machinery which have been conventionally Diesel based are becoming Diesel-electric, where the engine runs a generator and the motors do the direct work. This allows the use of an engine run at a constant speed where it can be well optimized. Also many ships of the future will rely on electric main drives rather than the conventional steam turbines. The use of electric drives means that there is no need to put a turbine in the bottom of the ship directly in line with the propeller shaft, which requires enormous space to be consumed by the ductwork and steam lines. If the turbine only provides electric power generation rather than direct propulsion, the turbine and reactor can be placed anywhere in the ship. However, very good reliable power electronics are necessary for the ship's motor drive.

Industrial motor drives are currently the most common high power electronics being used. These motor drives are in the range of 1000 to 20000 horsepower currently, and are facing increasing performance requirements. The output current to the motor must be as clean as possible to ensure smooth operation of the motor without torque ripples, and to minimize heating in the motor. The input to the motor drive also must be clean, in order to not contaminate the power grid. This requires an active front-end for the drive, which will draw currents that are both sinusoidal, and in phase with the input voltage, in order to minimize circulating energy in the grid and to avoid voltage distortion which comes from harmonic currents.

1.5.3 Other High Power Applications

Some other applications require very high power converters. One of these that is receiving much attention from the military is electromagnetic launchers. These applications require enormously high pulsed power, with a much lower average value. These electromagnetic launchers are being considered for launching aircraft from carriers to replace the steam catapults in use now. Electromagnetic guns are also being considered, and may be used in the future.

1.5.4 Challenges

These are some of the applications which require high power, high performance power electronic converters. Current thyristor technology allows for adequate power ratings, but the performance is only marginally acceptable now and will need to be improved in the future. The solution is the use of high-frequency, Pulse Width Modulated (PWM) converters, identical in principle to the ubiquitous small power supplies found in most electronic equipment, but with much higher power ratings. Unfortunately, this was impossible until a few years ago, due to the lack of a high-power, high-speed semiconductor switch. The conventional thyristor can be turned on by controls but cannot be forced to turn off by the gate, so it is immediately disqualified from use in PWM converters. The Gate Turn-Off (GTO) thyristor has high power capability as well and can be used for PWM converters, but the switching frequency cannot be made high enough to meet the performance requirements of the future.

The power Metal Oxide Semiconductor Field Effect Transistor (MOSFET) and the Bipolar Junction Transistor (BJT) have the required controllability and good speed, but they are not even close to having the needed power capacity.

1.5. 5 Opportunities

The prospects for high-power, high-frequency PWM converters have improved dramatically with the introduction of new power semiconductor devices. These new devices, including the Insulated Gate Bipolar Transistor (IGBT), the Integrated Gate Commutated Thyristor (IGCT), and the Emitter Turn-Off thyristor (ETO) have switching speeds similar to or better than those of a BJT, but with power ratings like the GTO. These devices also feature very good switching capability in that they can control a current level that is much higher than their normal operating current. With this new power semiconductor technology, work can begin on the high-power converters of the future.

On the other hand, power electronics is popular for technical as well as economical reasons. Nowadays, electronic power generation, transformation, transmission and distrubition are in AC, but almost all the terminal equipment used in industries, laboratories, locomotion, agricalture and households require DC power. In order to satisfy these requirements, easy conversion of AC power to DC power is essential. The conversion of AC to DC power at different frenquencies and DC to AC power can be effected through power electronics in a very dependable and economic manner.

Power electronics accupies and indispensable position in the field of battery charging uninterrupted power supply, electroplating, electronlysis, galvanisation and welding. It also plays an important role in all sorts of electric drives and lighting control. T he techniques developed during the past few years enable improved and more efficient manufacturing methods, accurate control and regulation of almost every kind of process. By means of electronic control, mechanical drives can be given almost any desired speed-torque charecteristics, the control apparatus being to all intents and purposes, inertialess and practically instantaneous in action. Feed drives of machine tools, multimotor drives in rolling mills, spinning machines, wire drawing mills, lift and many other drives may be given the required characteristics by means of electronic control. Electronically generated high-frequency energy offers possibilities in the wood-working and plastic industries for economical production of furniture, plywood and plastic articles, hardening, soldering or smelting of metals by high frequency energy increases the production of metal goods and contributes to improvement of quality of late, power electronic has assumed an extremely important role in modern main-line electric traction and power supply for urban transport systems as well as in high-voltage DC transmission.

1.6 Summary

In this chapter, an introduction to power electronics, power electronics application areas with the devices used in power electronics are given briefly.

CHAPTER TWO
POWER ELECTRONIC DEVICES

2.1 Overview

In this chapter, some commonly used devices for power electronics including their physical structure and their properties will be discussed.

2.2 Power Transistor

Transistors amplify current, for example they can be used to amplify the small output current from a logic chip so that it can operate a lamp, relay or other high current device. In many circuits a resistor is used to convert the changing current to a changing voltage, so the transistor is being used to amplify voltage.

A transistor may be used as a switch (either fully on with maximum current, or fully off with no current) and as an amplifier (always partly on).

The amount of current amplification is called the current gain, symbol hFE.

2.2.1 Types of Transistor

There are two types of standard transistors, NPN and PNP, with different circuit symbols. The letters refer to the layers of semiconductor material used to make the transistor. Most transistors used today are NPN because this is the easiest type to make from silicon. If you are new to electronics it is best to start by learning how to use NPN transistors.

The leads are labelled base (B), collector (C) and emitter (E). These terms refer to the internal operation of a transistor but they are not much help in understanding how a transistor is used, so just treat them as labels!

A Darlington pair is two transistors connected together to give a very high current gain.

In addition to standard (bipolar junction) transistors, there are field-effect transistors which are usually referred to as FETs.

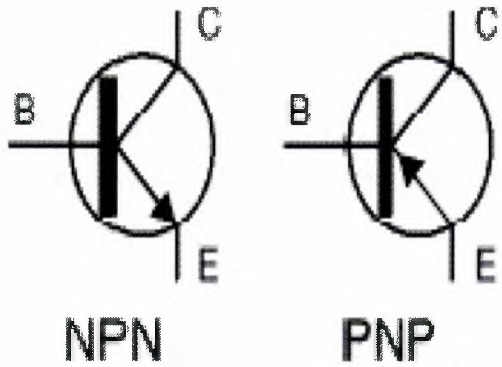


Figure 2.1 Transistors Ciccuit symbols.

2.2.2 Connecting

Transistors have three leads which must be connected the correct way round. Please take care with this because a wrongly connected transistor may be damaged instantly when you switch on.

2.2.3 Soldering

Transistors can be damaged by heat when soldering so if you are not an expert it is wise to use a heat sink clipped to the lead between the joint and the transistor body. A standard crocodile clip can be used as a heat sink.

2.2.4 Heat Sinks

Waste heat is produced in transistors due to the current flowing through them. Heat sinks are needed for power transistors because they pass large currents. If you find that a transistor is becoming too hot to touch it certainly needs a heat sink! The heat sink helps to dissipate (remove) the heat by transferring it to the surrounding air.

2.2.5 Testing a Transistor

Transistors can be damaged by heat when soldering or by misuse in a circuit. If you suspect that a transistor may be damaged there are two easy ways to test it:

2.2.5.1 Testing With a Multimeter

Use a multimeter or a simple tester (battery, resistor and LED) to check each pair of leads for conduction. Set a digital multimeter to diode test and an analogue multimeter to a low resistance range.

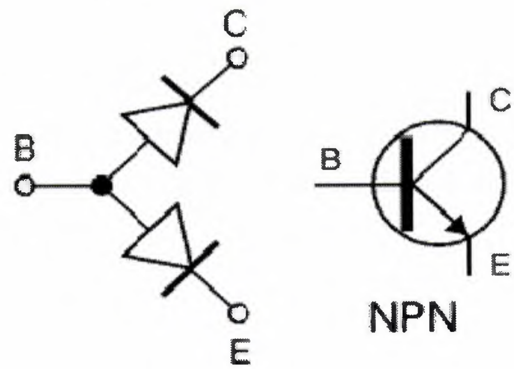


Figure 2.2 Testing an NPN Transistor.

Test each pair of leads both ways (six tests in total):
The base-emitter (BE) junction should behave like a diode and conduct one way only.
The base-collector (BC) junction should behave like a diode and conduct one way only.
The collector-emitter (CE) should not conduct either way.
The diagram shows how the junctions behave in an NPN transistor. The diodes are reversed in a PNP transistor but the same test procedure can be used.

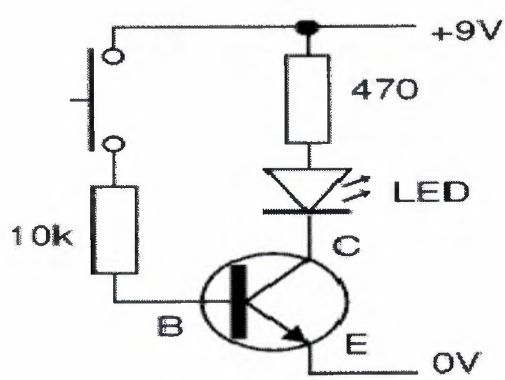


Figure 2.3 A Simple Switching Circuit to Test an NPN Tansistor.

Connect the transistor into the circuit shown which uses the transistor as a switch. The supply voltage is not critical, anything between 5 and 12V is suitable. This circuit can be quickly built on breadboard for example. Take care to include the 10kΩ resistor in the base connection or you will destroy the transistor as you test it!

If the transistor is OK the LED should light when the switch is pressed and not light when the switch is released.

To test a PNP transistor use the same circuit but reverse the LED and the supply voltage. Some multimeters have a 'transistor test' function which provides a known base current and measures the collector current so as to display the transistor's DC current gain h_{FE}.

2.2.6 Darlington Pair

This is two transistors connected together so that the amplified current from the first is amplified further by the second transistor. This gives the Darlington pair a very high current gain such as 10000. Darlington pairs are sold as complete packages containing the two transistors. They have three leads (B, C and E) which are equivalent to the leads of a standard individual transistor.

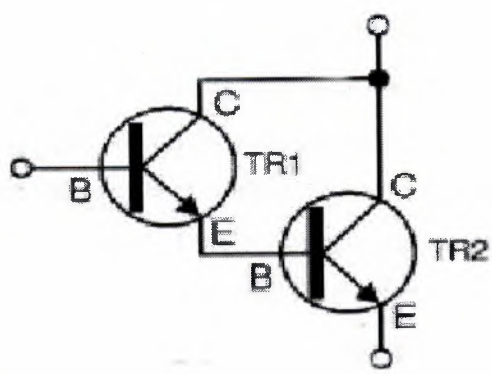


Figure 2.4 Darlington Pair of Two Transistors.

2.3 Bipolar Power Transistors

A bipolar power transistor, unlike a thyristor-like device, is two-junction, self controlled device where the collector current is under the control of the base drive current. Bipolar power transistor have recently been ousted by IGBTs(insulated gate bipolar transistors) in the higher end and by power MOSFETs in the lower end. The DC current gain (hfe) of a power transistor is low and varies widely with collector current and temperature. The gain is increased to a high value in the Darlington connection. However, the disadvantages are higher leakage current, higher conduction drop, and reduces switching frequency. The shunt resistances and diode in the base-emitter circuit help to reduce collector leakage current and establish base bias voltages. A transistor can block voltage in the forward direction only (asymmetric blocking). The feedback diode is an essential element for chopper and voltage-fed converter applications. Double or triple Darlington transistors are available in module form with matched parallel devices for higher power rating.

Power transistors have an important property known as the second breakdown effect. This is in contrast to the avalanche breakdown effect of a junction, which is also known as first breakdown effect. When the collector current is switched on by the base drive, it tends to crowd on the base-emitter junction periphery, thus constricting the collector current in a narrow area of the reverse-biased collector junction. This tends to create a hot spot and the junction fails by thermal runaway, which is known as second breakdown. The rise in junction temperature at the hot spot accentuates the current concentration owing to the negative temperature coefficient of the drop, and this regeneration effect causes collapse of the collector voltage, thus destroying the device. A similar problem arises when an inductive load is turned off. As the base-emitter junction becomes reverse-biased, the collector current tends to concentrate in a narrow area of the collector junction.

Manifacturers provide specifications in the form of safe areas (SOAs) during turn-on (FBSOA) and turn-off (RBSOA). Obviously, a well-designed polarized RC snubber is indispensable in a transistor converter.

2.4 Thyristor

The thyristor is a solid-state semiconductor device similar to a diode, with an extra terminal which is used to turn it on. Once turned on, the thyristor will remain on (conducting) as long as there is a significant current flowing through it. If the current falls to zero, the device switches off. Some resources define silicon controlled rectifiers and thyristors as synonymous1, while others define SCR's as a subset of thyristors2.

The thyristor is a four-layer semiconducting device, with each layer consisting of an alternately N or P-type material, for example N-P-N-P. The main terminals, labeled anode and cathode, are across the full four layers, and the control terminal, called the gate, is attached to one of the middle layers. The operation of a thyristor can be understood in terms of a pair of tightly coupled transistors, arranged to cause the self-latching action.

Thyristors are mainly used where high currents and voltages are involved, and are often used to control alternating currents, where the change of sign of the current causes the device to automatically switch off. This is known as synchronous operation or Zero Cross operation. This principle is used to control the desired loading by adjusting the frequency of the sinusoidal input. The range of frequencies is great because there is no limit to the number of cycles a thyristor can perform, and exhibits no "wear out" modes. This is a frequency domain method of control.

With phase angle control a thyristor is turned on at a specific and adjustable portion of the cycle of the controlling sinusoidal input. Moving the point at which the thyristor is turned on regulates power output. An example of this method of control is a dimmer switch for lights. The turn on point of a thyristor is controlled to occur at a particular point on the sine curve of the AC supply. The thyristor stays on for the remainder of that cycle and the longer the thyristor stays on, the brighter the light. Fine resolution of output is possible with this method and is suitable for slow-responding loads such as tungsten filament lamps or temperature variable resistance loads. Phase-angle control is also essential for inductive loads.

The drawback of a thyristor is that, like a diode, it only conducts in one direction. A similar self-latching 5-layer device, called a triac, is able to work in both directions. This added capability, though, also can become a shortfall. Because the triac can conduct in both directions, reactive loads can cause the triac to fail to turn off during the zero-voltage instants of the ac power cycle. Because of this, use of triacs with (for example) heavily-inductive motor loads usually requires the use of a "snubber" circuit around the triac to assure that it will turn off with each half-cycle of mains power. Inverse-parallel SCRs can also be used in place of the triac; because each SCR in the pair has an entire half-cycle of reverse polarity applied to it, the SCRs, unlike triacs, are sure to turn off.

An earlier gas filled tube device called a Thyratron provided a similar electronic switching capability, where a small control voltage could switch a large current. It is from a combination of "thyratron" and "transistor" that the term "thyristor" is derived.

Modern thyristors can switch large amounts of power (up to megawatts). In the realm of very high power applications, they are still the primary choice. However, in low and medium power (from few tens of watts to few tens of kilowatts) they have almost been replaced by other devices with superior switching characteristics like MOSFETs or IGBTs. One major problem associated with the thyristor is that it is not a fully controllable switch in the sense that triggering current direction needs to be reversed to switch it off. GTO (Gate Turn-off Thyristor) is another related device which addresses this problem. In high-frequency applications, thyristors are poor candidates due to large switching times arising out of bipolar conduction. MOSFETs, on the other hand, have much faster switching capability because of their unipolar conduction (only majority carriers carry the current).

2.4.1 Types of Thyristors

a) Silicon controlled rectifier

b) Triac, a bidirectional switching device containing two thyristor structures

c) Gate turn-off thyristor (GTO thyristor)

d) Mosfet controlled thyristor (MCT), two additional FET structures for on and off control.

e) Static induction thyristor (SITh) or Field controlled thyristor (FCTh), a gate structure can shut down anode current flow.

2.4.2 Volt-Ampere Characteristics

Figure 2.5 shows the thyristor symbol and its volt-ampere characteristics. Basically, it is a three-junction P-N-P-N device, where P-N-P and N-P-N component transistors are connected in regenerative feedback mode. The device blocks voltage in both the forward and reverse directions (symmetric blocking). When the anode is positive, the device can be triggered into conduction by a short positive gate current pulse;but once the device is conducting, the gate loses its control to turn off the device. A thyristor can also turn on by excessive anode voltage, its rate of rise (dv/dt), by a rise in junction temperature (Tj), or by light shining on the junctions.

The volt-ampere characteristics of the device indicate that at gate current Ig=0, if forward voltage applied on the device, there will be a leakage current due to blocking of the middle junction. If the voltage exceeds a critical limit(breakover voltage), the device switches into conduction. With increasing magnitude of Ig, the forward breakover voltage is reduced, and eventually at Ig3, the device behaves like a diode with the entire forward blocking region removed. The device will turn on successfully if a minimum current, called a latching current, is maintained. During conduction, if the gate current is zero and the anode current falls below a critical limit, called the holding current, the device reverts to the forward blocking state. With reverse voltage, the end P-N junctions of the device become reverse-biased and the I-V curve becomes essentially similar to that of a diode rectifier. Modern thyristors are available with very large voltage(several KV) and current (several KA) ratings.

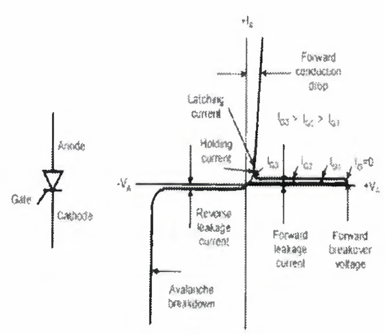


Figure 2.5 Thyristor Symbol and Volt-Ampere Characteristics.

2.4.3 Switching Characteristics

Initially , when forward voltage applied across a device, the off-state, or static dv/dt, must be limited so that it does not switch on spuriously. The dv/dt creates displacement current in the depletion layer capacitance of the middle junction, which induces emitter current in the component resistors and causes switching action. When the device turns on, the anode current di/dt can be excessive, which can destroy the device by heavy current concentration. During conduction, the inner P-N regions remain heavily saturated with minority carriers and the middle junctions remains forward-biased. To recover the forward voltage blocking capability, a reverse voltage is applied across the device to sweep out the minority carriers and the phenomena are similar to diode. However, when the recovery current goes to zero, the middle junction still remains forward-biased. This junction eventually blocks with an additional delay when the minority carriers die by the recombination process. The forward voltage can then be applied successfully, but the reapplied dv/dt will be somewhat less than the static dv/dt because of the presence of minority carriers. A suitably-designed snubber circuit can limit di/dt and dv/dt within acceptable limits. In a converter circuit, a thyristor can be turned off(or commutated) by an inductance –capacitance circuit induced transient reverse voltage.

2.4.4 Power Loss and Thermal Impedance

A thyristor has dominant conduction loss like a diode, but its switching loss is very small. The device specification sheet normally gives information on power dissipation for various duty cycles of sinusoidal and rectangular current wave. The reverse blocking loss and gate circuit loss are also included in the figure 2.6. These curves are valid up to 400 Hz supply frequency. The heat due to power loss in the vicinity of a junction flows to the case and then to the ambient through the externally mounted heat sink, causing a rise in the junction temperature T_J . The maximum T_J of a device is to be limited because of its adverse effect on device performance. For steady power dissipation P, T_J can be calculated as:

$$T_J - T_A = P(Q_{JA} - Q_{CS} - Q_{SA})$$

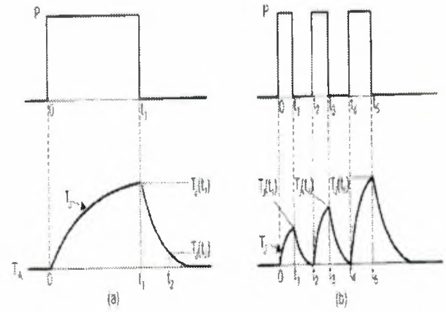
where T_A is the ambient temperature, and Q_{JC}, Q_{CS}, Q_{SA} represent the thermal resistance from junction to case, case to sink, and sink to ambient, respectively. The resistance Q_{SA} is determined by the cooling system design, and the methods of cooling may include heat sink with naturel convection cooling, forced air cooling, or forced liquid cooling. From the equation, it is evident that for a limited T_{jmax}(usually 125O ^oC), the dissipation P can be increased by reducing Q_{SA}. This means that a more efficient cooling system will increase power dissipation, that is, the power-handling capability of a device. An infinite heat sink is defined when $Q_{SA}=0$, that is, the case temperature $T_C=T_A$.

In practical operation, the power dissipation P is cyclic, and the thermal capacitance or storage effect delays the junction temperature rise, thus permitting higher loading of a device. The transient thermal equivalent circuit can be represented by a parellel RC circuit, where P is equivalent to the current source and the resulting voltage across the circuit represents the temperature T_J. Figure2.6 shows the T_J curve for the dissipation of a single power pulse. Considering the complimentary nature of heating and cooling curves, the following equations can be written:

$$T_J(t1)=T_A+PQ(t1)$$
$$T_J(t2)=T_A+P[Q(t2)-Q(t2-t1)]$$

where Q(t1) is the transient thermal impedance at time t1. The device specification sheet normally gives thermal impedance between junction and case. The additional effect due to heat sink can be added if desired. Figure2.6 shows typical junction temperature build-up for three repeated pulses. The corresponding TJ expression by the superposition principle can be given as:

$$T_J(t1)=T_A+PQ(t1)$$
$$T_J(t3)=T_A+P[Q(t3)-Q(t3-t1)+Q(t3-t2)]$$
$$T_J(t5)=T_A+P[Q(t5)-Q(t5-t1)+Q(t5-t2)-Q(t5-t3)+Q(t5-t4)]$$



a)single pulse b)multiple pulses

Figure 2.6 Junction Temperature Rise with Pulsed Power Dissipation.

2.4.5 Current Rating

Based on the criteria of limiting T_J as discussed above. Figure 2.7 shows the average current rating IT(AV) vs. permissible case temperature T_C for various duty cycles of rectangular current wave. For example, if T_C is limited to 110 °C, the thyristor can carry 12 A average current for=120 °C. If a better heat sink limits T_C to 100 °C, the current can be increased to 18 A.

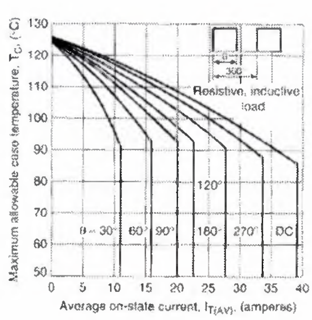


Figure 2.7 Maximum Allowable Case Temperature for Rectangular Current Wave.

2.5 The Silicon-Controlled Rectifier (SCR)

Shockley diodes are curious devices, but rather limited in application. Their usefulness may be expanded, however, by equipping them with another means of latching. In doing so, they become true amplifying devices (if only in an on/off mode), and we refer to them as silicon-controlled rectifiers, or SCRs.

The progression from Shockley diode to SCR is achieved with one small addition, actually nothing more than a third wire connection to the existing PNPN structure:

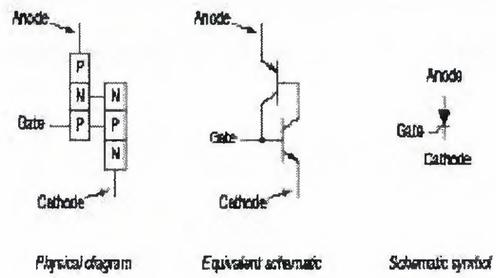


Figure 2.8 Diagrams of SCRs.

If an SCR's gate is left floating (disconnected), it behaves exactly as a Shockley diode. It may be latched by breakover voltage or by exceeding the critical rate of voltage rise between anode and cathode, just as with the Shockley diode. Dropout is accomplished by reducing current until one or both internal transistors fall into cutoff mode, also like the Shockley diode. However, because the gate terminal connects directly to the base of the lower transistor, it may be used as an alternative means to latch the SCR. By applying a small voltage between gate and cathode, the lower transistor will be forced on by the resulting base current, which will cause the upper transistor to conduct, which then supplies the lower transistor's base with current so that it no longer needs to be activated by a gate voltage. The necessary gate current to initiate latch-up, of course, will be much lower than the current through the SCR from cathode to anode, so the SCR does achieve a measure of amplification.

This method of securing SCR conduction is called triggering, and it is by far the most common way that SCRs are latched in actual practice. In fact, SCRs are usually chosen so that their breakover voltage is far beyond the greatest voltage expected to be experienced from the power source, so that it can be turned on only by an intentional voltage pulse applied to the gate.

It should be mentioned that SCRs may sometimes be turned off by directly shorting their gate and cathode terminals together, or by "reverse-triggering" the gate with a negative voltage (in reference to the cathode), so that the lower transistor is forced into cutoff. I say this is "sometimes" possible because it involves shunting all of the upper transistor's collector current past the lower transistor's base. This current may be substantial, making triggered shut-off of an SCR difficult at best. A variation of the SCR, called a Gate-Turn-Off thyristor, or GTO, makes this task easier. But even with a GTO, the gate current required to turn it off may be as much as 20% of the anode (load) current! The schematic symbol for a GTO is shown in the following illustration:

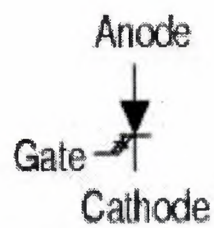


Figure 2.9 Gate Turn-Off Thyristor.

SCRs and GTOs share the same equivalent schematics (two transistors connected in a positive-feedback fashion), the only differences being details of construction designed to grant the NPN transistor a greater β than the PNP. This allows a smaller gate current (forward or reverse) to exert a greater degree of control over conduction from cathode to anode, with the PNP transistor's latched state being more dependent upon the NPN's than visa-versa. The Gate-Turn-Off thyristor is also known by the name of Gate-Controlled Switch, or GCS.

A rudimentary test of SCR function, or at least terminal identification, may be performed with an ohmmeter. Because the internal connection between gate and cathode is a single PN junction, a meter should indicate continuity between these terminals with the red test lead on the gate and the black test lead on the cathode like this:

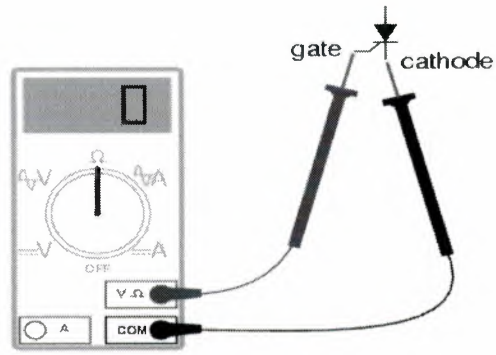


Figure 2.10 Test Diagram of GTO.

All other continuity measurements performed on an SCR will show "open" ("OL" on some digital multimeter displays). It must be understood that this test is very crude and does not constitute a comprehensive assessment of the SCR. It is possible for an SCR to give good ohmmeter indications and still be defective. Ultimately, the only way to test an SCR is to subject it to a load current.

If you are using a multimeter with a "diode check" function, the gate-to-cathode junction voltage indication you get may or may not correspond to what's expected of a silicon PN junction (approximately 0. 7 volts). In some cases, you will read a much lower junction voltage: mere hundredths of a volt. This is due to an internal resistor connected between the gate and cathode incorporated within some SCRs. This resistor is added to make the SCR less susceptible to false triggering by spurious voltage spikes, from circuit "noise" or from static electric discharge. In other words, having a resistor connected across the gate-cathode junction requires that a strong triggering signal (substantial current) be applied to latch the SCR. This feature is often found in larger SCRs, not on small SCRs. Bear in mind that an SCR with an internal resistor connected between gate and cathode will indicate continuity in both directions between those two terminals:

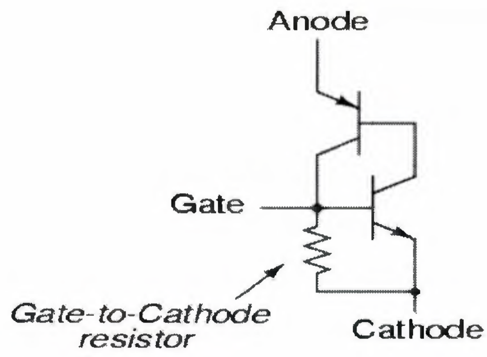


Figure 2.11 Schematic Diagram of GTO.

"Normal" SCRs, lacking this internal resistor, are sometimes referred to as sensitive gate SCRs due to their ability to be triggered by the slightest positive gate signal.

The test circuit for an SCR is both practical as a diagnostic tool for checking suspected SCRs and also an excellent aid to understanding basic SCR operation. A DC voltage source is used for powering the circuit, and two pushbutton switches are used to latch and unlatch the SCR, respectively:

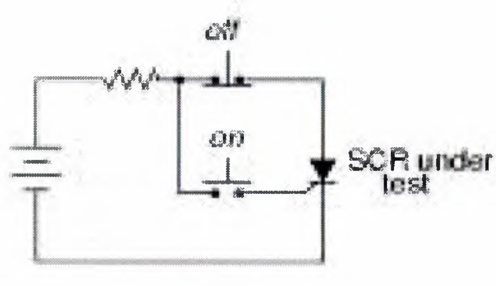


Figure 2.12 SCR Testing Circuit.

Actuating the normally-open "on" pushbutton switch connects the gate to the anode, allowing current from the negative terminal of the battery, through the cathode-gate PN junction, through the switch, through the load resistor, and back to the battery. This gate current should force the SCR to latch on, allowing current to go directly from cathode to anode without further triggering through the gate. When the "on" pushbutton is released, the load should remain energized.

Pushing the normally-closed "off" pushbutton switch breaks the circuit, forcing current through the SCR to halt, thus forcing it to turn off (low-current dropout).
If the SCR fails to latch, the problem may be with the load and not the SCR. There is a certain minimum amount of load current required to hold the SCR latched in the "on" state. This minimum current level is called the holding current. A load with too great a resistance value may not draw enough current to keep an SCR latched when gate current ceases, thus giving the false impression of a bad (unlatchable) SCR in the test circuit. Holding current values for different SCRs should be available from the manufacturers. Typical holding current values range from 1 milliamp to 50 milliamps or more for larger units.

For the test to be fully comprehensive, more than the triggering action needs to be tested. The forward breakover voltage limit of the SCR could be tested by increasing the DC voltage supply (with no pushbuttons actuated) until the SCR latches all on its own. Beware that a breakover test may require very high voltage: many power SCRs have breakover voltage ratings of 600 volts or more! Also, if a pulse voltage generator is available, the critical rate of voltage rise for the SCR could be tested in the same way: subject it to pulsing supply voltages of different V/time rates with no pushbutton switches actuated and see when it latches.

In this simple form, the SCR test circuit could suffice as a start/stop control circuit for a DC motor, lamp, or other practical load:

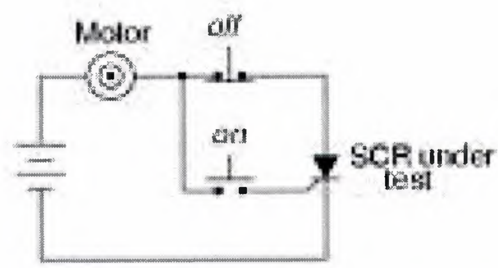


Figure 2.13 DC Motor Stop-Start Control Circuit.

Another practical use for the SCR in a DC circuit is as a crowbar device for overvoltage protection. A "crowbar" circuit consists of an SCR placed in parallel with the output of a DC power supply, for the purpose of placing a direct short-circuit on the output of that supply to prevent excessive voltage from reaching the load. Damage to the SCR and power supply is prevented by the judicious placement of a fuse or substantial series resistance ahead of the SCR to limit short-circuit current:

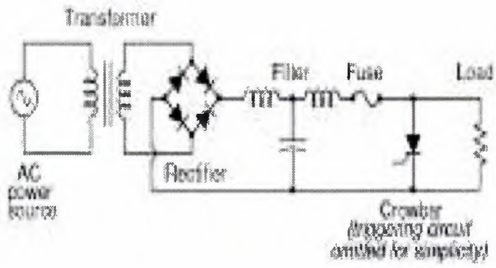


Figure 2.14 Crowbar as Used In AC-DC Power Supply.

Some device or circuit sensing the output voltage will be connected to the gate of the SCR, so that when an overvoltage condition occurs, voltage will be applied between the gate and cathode, triggering the SCR and forcing the fuse to blow. The effect will be approximately the same as dropping a solid steel crowbar directly across the output terminals of the power supply, hence the name of the circuit.

Most applications of the SCR are for AC power control, despite the fact that SCRs are inherently DC (unidirectional) devices. If bidirectional circuit current is required, multiple SCRs may be used, with one or more facing each direction to handle current through both half-cycles of the AC wave. The primary reason SCRs are used at all for AC power control applications is the unique response of a thyristor to an alternating current. As we saw in the case of the thyratron tube (the electron tube version of the SCR) and the DIAC, a hysteretic device triggered on during a portion of an AC half-cycle will latch and remain on throughout the remainder of the half-cycle until the AC current decreases to zero, as it must to begin the next half-cycle. Just prior to the zero-crossover point of the current waveform, the thyristor will turn off due to insufficient current (this behavior is also known as natural commutation) and must be fired again during the next cycle. The result is a circuit current equivalent to a "chopped up" sine wave. For review, here is the graph of a DIAC's response to an AC voltage whose peak exceeds the breakover voltage of the DIAC:

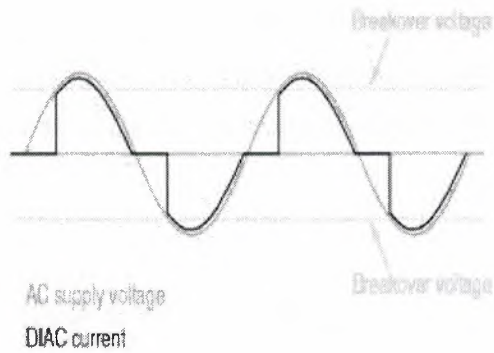


Figure 2.15 DIAC's Response to an AC Voltage

With the DIAC, that breakover voltage limit was a fixed quantity. With the SCR, we have control over exactly when the device becomes latched by triggering the gate at any point in time along the waveform. By connecting a suitable control circuit to the gate of an SCR, we can "chop" the sine wave at any point to allow for time-proportioned power control to a load.

Take the following circuit as an example. Here, an SCR is positioned in a circuit to control power to a load from an AC source:

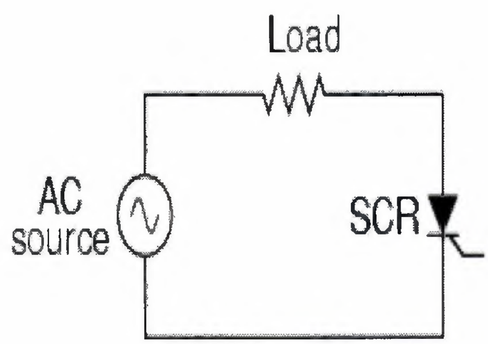


Figure 2.16 SCR is Positioned in a Circuit to Control Power.

Being a unidirectional (one-way) device, at most we can only deliver half-wave power to the load, in the half-cycle of AC where the supply voltage polarity is positive on the top and negative on the bottom. However, for demonstrating the basic concept of time-proportional control, this simple circuit is better than one controlling full-wave power (which would require two SCRs).

With no triggering to the gate, and the AC source voltage well below the SCR's breakover voltage rating, the SCR will never turn on. Connecting the SCR gate to the anode through a normal rectifying diode (to prevent reverse current through the gate in the event of the SCR containing a built-in gate-cathode resistor), will allow the SCR to be triggered almost immediately at the beginning of every positive half-cycle:

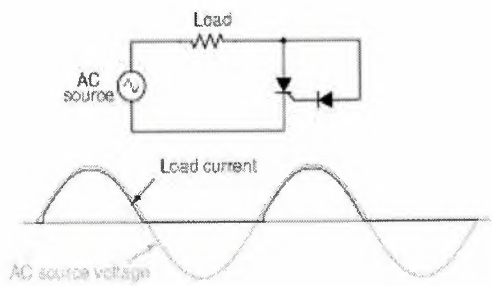


Figure 2.17 Gate Connected Directly to Anode Through a Diode.

We can delay the triggering of the SCR, however, by inserting some resistance into the gate circuit, thus increasing the amount of voltage drop required before there is enough gate current to trigger the SCR. In other words, if we make it harder for electrons to flow through the gate by adding a resistance, the AC voltage will have to reach a higher point in its cycle before there will be enough gate current to turn the SCR on. The result looks like this:

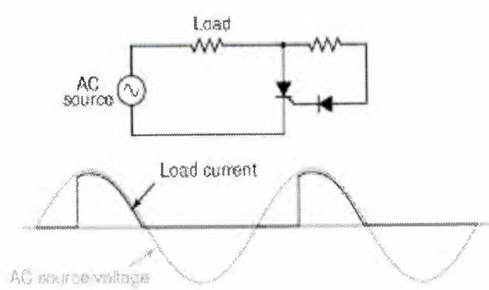


Figure 2.18 Resistance Inserted in Gate Circuit.

With the half-sine wave chopped up to a greater degree by delayed triggering of the SCR, the load receives less average power (power is delivered for less time throughout a cycle). By making the series gate resistor variable, we can make adjustments to the time-proportioned power:

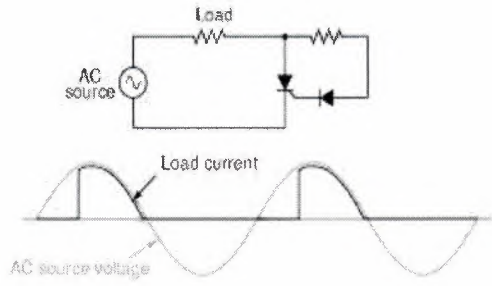


Figure 2.19 Increasing the Resistance the Threshold Level.

Unfortunately, this control scheme has a significant limitation. In using the AC source waveform for our SCR triggering signal, we limit control to the first half of the waveform's half-cycle. In other words, there is no way for us to wait until after the wave's peak to trigger the SCR. This means we can turn down the power only to the point where the SCR turns on at the very peak of the wave:

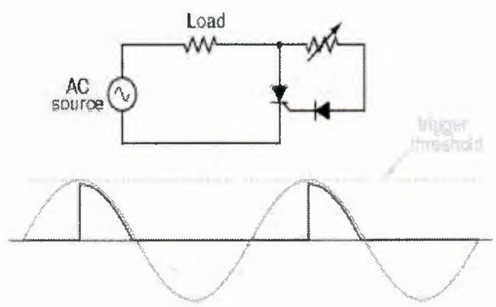


Figure 2.20 Circuit at Minimum Power Setting.

Raising the trigger threshold any more will cause the circuit to not trigger at all, since not even the peak of the AC power voltage will be enough to trigger the SCR. The result will be no power to the load.

An ingenious solution to this control dilemma is found in the addition of a phase-shifting capacitor to the circuit:

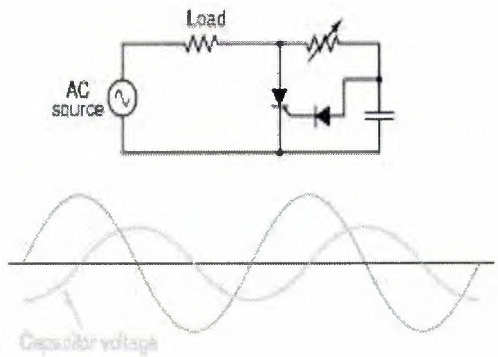


Figure 2.21 A phase Shifting Circuit.

The smaller waveform shown on the graph is voltage across the capacitor. For the sake of illustrating the phase shift, I'm assuming a condition of maximum control resistance where the SCR is not triggering at all and there is no load current, save for what little current goes through the control resistor and capacitor. This capacitor voltage will be phase-shifted anywhere from 0° to 90° lagging behind the power source AC waveform. When this phase-shifted voltage reaches a high enough level, the SCR will trigger.

Assuming there is periodically enough voltage across the capacitor to trigger the SCR, the resulting load current waveform will look something like this:

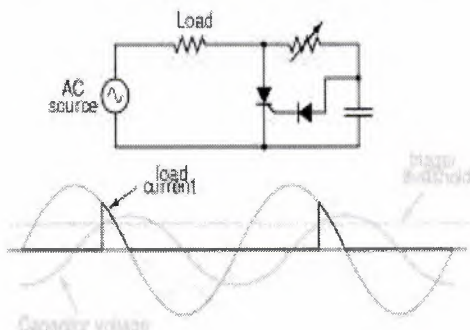


Figure 2.22 Trigger the SCR.

Because the capacitor waveform is still rising after the main AC power waveform has reached its peak, it becomes possible to trigger the SCR at a threshold level beyond that peak, thus chopping the load current wave further than it was possible with the simpler circuit. In reality, the capacitor voltage waveform is a bit more complex that what is shown here, its sinusoidal shape distorted every time the SCR latches on. However, what I'm trying to illustrate here is the delayed triggering action gained with the phase-shifting RC network, and so a simplified, undistorted waveform serves the purpose well.

SCRs may also be triggered, or "fired, " by more complex circuits. While the circuit previously shown is sufficient for a simple application like a lamp control, large industrial motor controls often rely on more sophisticated triggering methods. Sometimes, pulse transformers are used to couple a triggering circuit to the gate and cathode of an SCR to provide electrical isolation between the triggering and power circuits:

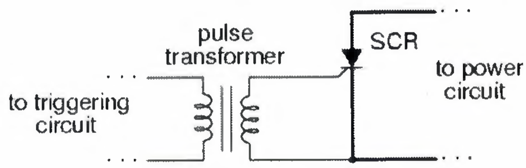


Figure 2.23 SCR to Provide Electrical Isolation.

When multiple SCRs are used to control power, their cathodes are often not electrically common, making it difficult to connect a single triggering circuit to all SCRs equally. An example of this is the controlled bridge rectifier shown here:

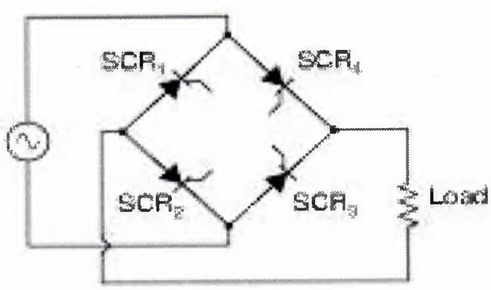


Figure 2.24 Controlled Bridge Rectifier.

In any bridge rectifier circuit, the rectifying diodes (or in this case, the rectifying SCRs) must conduct in opposite pairs. SCR1 and SCR3 must be fired simultaneously, and likewise SCR2 and SCR4 must be fired together as a pair. As you will notice, though, these pairs of SCRs do not share the same cathode connections, meaning that it would not work to simply parallel their respective gate connections and connect a single voltage source to trigger both:

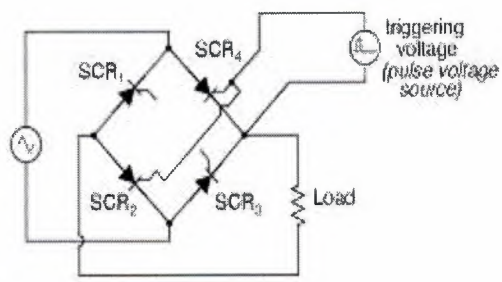


Figure 2.25 Pair of SCRs.

Although the triggering voltage source shown will trigger SCR4, it will not trigger SCR2 properly because the two thyristors do not share a common cathode connection to reference that triggering voltage. Pulse transformers connecting the two thyristor gates to a common triggering voltage source will work.

Bear in mind that this circuit only shows the gate connections for two out of the four SCRs. Pulse transformers and triggering sources for SCR1 and SCR3, as well as the details of the pulse sources themselves, have been omitted for the sake of simplicity.

Controlled bridge rectifiers are not limited to single-phase designs. In most industrial control systems, AC power is available in three-phase form for maximum efficiency, and solid-state control circuits are built to take advantage of that. A three-phase controlled rectifier circuit built with SCRs, without pulse transformers or triggering circuitry shown, would look like this:

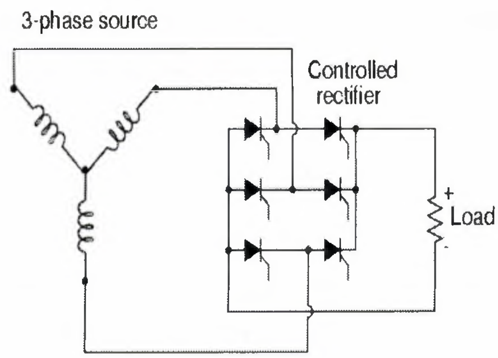


Figure 2.26 A Three Phase Controlled Rectifier Circuit Built with SCRs.

2.5.1 Circuit Operation

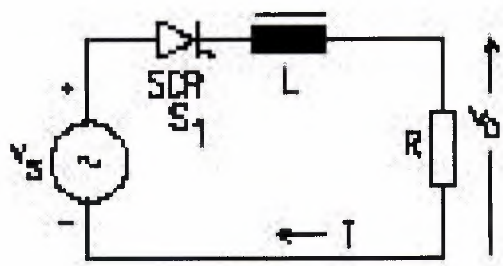


Figure 2.27 A Single SCR Circuit.

A circuit with a single SCR and an RL load is shown above. The source vs is an alternating sinusoidal source. If vs = E * sin (wt), vs is positive when $0 < wt < 180^{\circ}$, and vs is negative when $180^{\circ} < wt < 360^{\circ}$. When vs starts becoming positive, the SCR is forward-biased but remains in the blocking state till it is triggered. If the SCR is triggered at when wt = 0 then 0 is called the firing angle. When the SCR is triggered in the forward-bias state, it starts conducting and the positive source keeps the SCR in conduction till wt reaches 180 degrees. At that instant, the current through the circuit is not zero and there is some energy stored in the inductor at wt = 180 degrees. The voltage across an inductor is positive when the current through it is increasing and it becomes negative when the current through the inductor tends to fall. When the voltage across the inductor is negative, it is in such a direction as to forward-bias the SCR.

There is current through the load at the instant wt = 180 degrees and the SCR continues to conduct till the energy stored in the inductor becomes zero. After that the current tends to flow in the reverse direction and the SCR blocks conduction. The entire applied voltage now appears across the diode.

2.6 IGBT

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.2.28 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 pn 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 interdigitated geometry similar to that used for the vertical MOSFET.

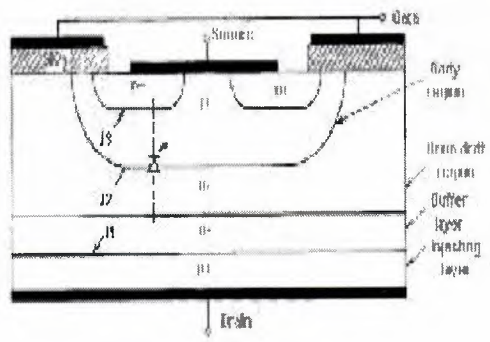


Figure 2.28 Physical Structure of an IGBT.

The IGBT structure shown in Fig. 2.28 has a parasitic thyristor which could latchup 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 lower2 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. 2.29.

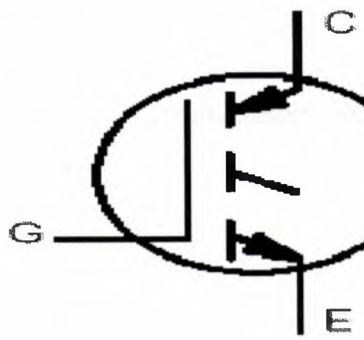


Figure 2.29 IGBT Circuit Symbol.

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.2.28. 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.

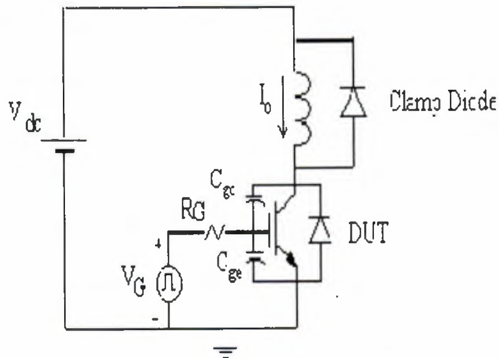


Figure 2.30 Inductive Load Test Circuit.

The turn-on switching transient of an IGBT with an inductive load are shown in Fig. 2.29. 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 VGE(th) due to the circuit gate resistance (RG) and the gate to emitter capacitance (Cge). The Miller effect capacitance (Cgc) effect is very small due to the high voltage across the device terminals.

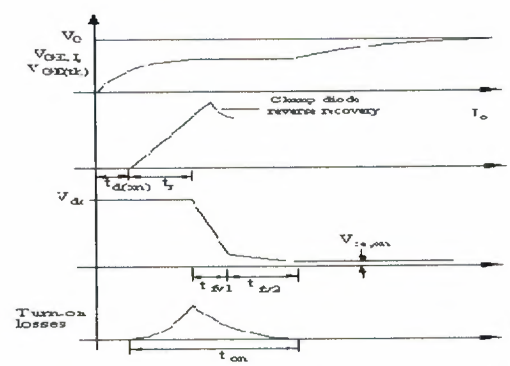


Figure 2.31 IGBT Turn-On Switching Transient with Inductive Load.

Beyond VGE(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 Vdc 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 VGE, 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 tfv1 and tfv2. 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. 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 VG 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.

The turn-off switching transient of an IGBT with an inductive load are shown in Fig.2.32.

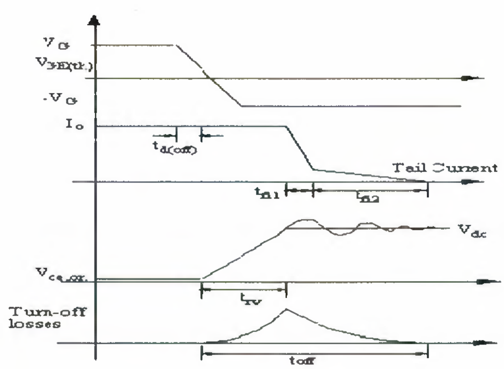


Figure 2.32 IGBT Turn-Off Switching Transient with Inductive Load.

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 (VGE(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 tfi1 which corresponds to the turn-off of the MOSFET part of the IGBT (Fig.2.33). The tailing of the collector current during the second interval tfi2 is due to the stored charge in the n- drift region of the device. This is due to the fact that 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 so as 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.

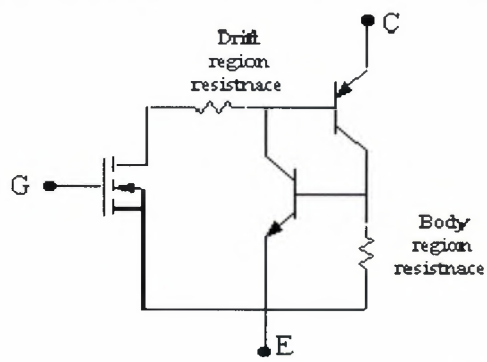


Figure 2.33 Equivalent Circuit of the IGBT.

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 life time which 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.2.34).

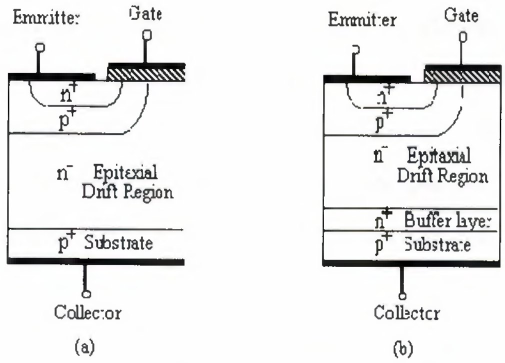


Figure 2.34 (a) Non Punch Through (NPT) IGBT (b) Punch Through (PT) IGBT.

The turn off energy loss, also shown in Fig.2.34, can be evaluated in a similar fashion as the turn-on losses, namely,

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

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.2.34 (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.2.34 (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, dVce/dt, is increased. The RBSOA is reduced as the dVce/dt is increased so as to avoid latch up within the device. This condition exists when higher values of dVce/dt are applied may give to the rise to a pulse of forward decaying current in the body region of the device which acts as a pulse of gate current that can turn on the device. Fortunately, the dVce/dt values that would cause latch up in IGBTs are much higher compared to other devices.

The maximum value of ICM is set so as to avoid latch up which is determined based on the dynamic latch up condition. In addition, a maximum VGE 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.

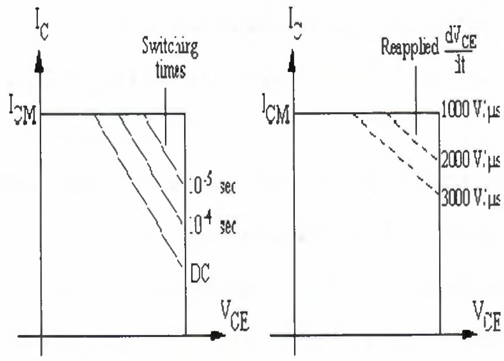


Figure 2.35 (a) FBSOA (b) RBSOA of an IGBT.

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.2.36.

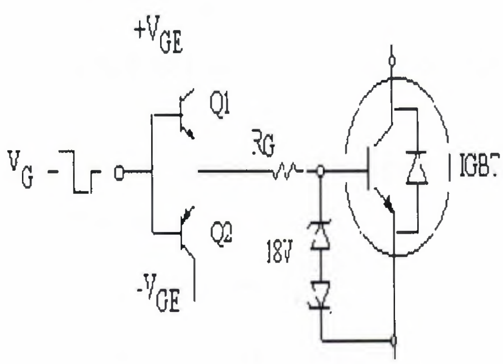


Figure 2.36 Typical Gate Drive Circuitry.

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.2.37.

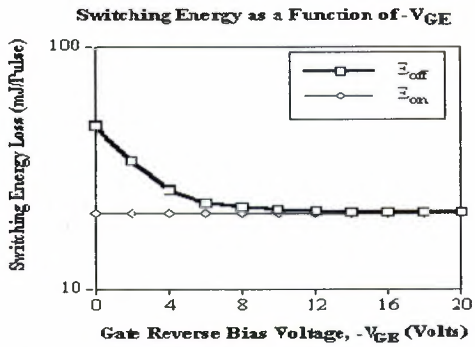


Figure 2.37 Effect of Negative Bias on Turn-Off Losses.

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.2.38). However, a small gate resistance can lead to oscillations between the IGBT input capacitance and the parasitic lead inductance.

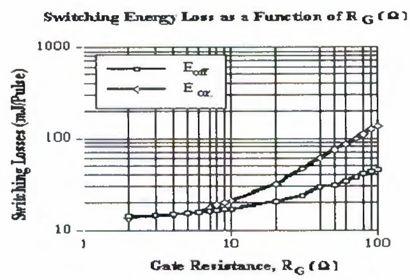


Figure 2.38 The IGBT Switching Losses as a Function of Gate Resistance, RG.

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 = VGE . QG . fs

where,

DVGE = VGE_on + |VGE_off|

QG = total gate charge (per manuf. spec.)

fs = switching frequency

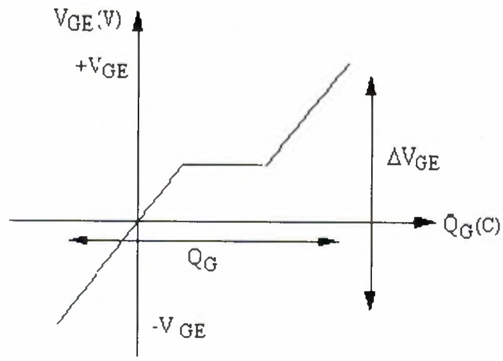


Figure 2.39 Total IGBT Gate Charge During Switching.

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.2.40) or optical isolation.

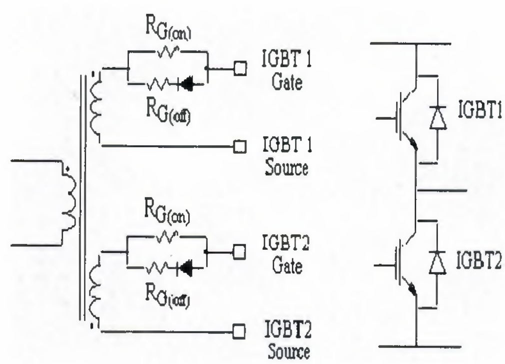


Figure 2.40 Typical Bipolar IGBT Gate Drive Using Gate Pulse Transformers.

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

2.6.1 Gate Drive Layout Considerations

Minimize parasitic inductance between the driver output stage and the IGBT (minimizing the loop area)

Minimize noise coupling via proper shielding techniques

Utilize gate clamp protections (TVS) to minimize overvoltage across gate terminals

Utilize twisted pairs, preferably shielded, for indirect connection between the driver and the IGBT

With optocoupling isolation, a minimum of 10, 000 V/ms transient immunity must be provided (in hard switching applications)

2.7 Diodes

A diode is a semiconductor device with two terminals (gate and cathode). It conducts current whenever a positive anode-cathode voltage exists. It blocks when anode-cathode voltage is negative. See diode symbol in illustration.

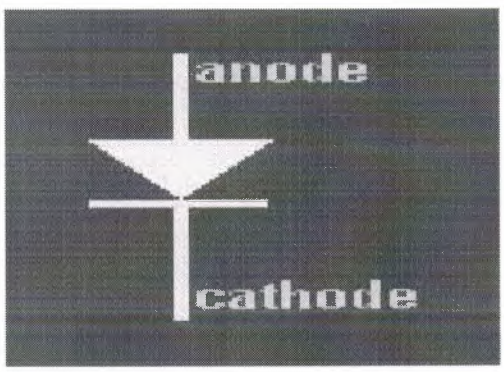


Figure 2.41 Symbol of Diode.

Most of the power electronic applications operate at a relative high voltage and in such cases, the voltage drop across the power diode tends to be small. It is quite often justifiable to use the ideal diode model. An ideal diode has zero conduction drop when it is forward-biased and has zero current when it is reverse-biased. The explanation and the analysis presented below is based on the ideal diode model.

2.7.1 Circuit Operation

A circuit with a single diode and the source vs is an alternating sinusoidal source. If vs = E * sin (wt), vs is positive when $0 < wt < 360^{o}$, and vs is negative when $180 < wt < 360^{o}$. When vs starts becoming positive, the diode starts conducting and the source keeps the diode in conduction till wt reaches 180^{o}. At that instant defined by $wt = 180^{o}$, the current through the circuit is not zero and there is some energy stored in the inductor. The voltage across an inductor is positive when the current through it is increasing and it becomes negative when the current through it tends to fall. When the voltage across the inductor is negative, it is in such a direction as to forward-bias the diode.

When vs changes from a positive to a negative value, there is current through the load at the instant $wt = 0^{o}$ and the diode continues to conduct till the energy stored in the inductor becomes zero. After that the current tends to flow in the reverse direction and the diode blocks conduction.The entire applied voltage now appears across the diode.

2.8 Summary

In this chapter, power electronic devices, their structural analysis and their specificic properties with their circuit analysis are explained briefly.

CHAPTER THREE
POWER ELECTRONIC CONVERTERS

3.1 Overview

In this chapter,power electronic converters and their classification,converters application areas,circuit diagrams and converters specification with advantages and disadvantages will be given briefly.

3.2 Introduction the Power Converters

The research into power electronic converters is concerned with new soft-switching techniques which enable efficient operation at high switching frequencies, thereby permitting the miniturisation of transformer and filter components. The work also embraces new fabrication methods for power converters and magnetic components. Power levels are typically up to 10kW with operating frequencies in excess of 1MHz at low power. The applications are mainly in low voltage DC power supplies; current projects include multi-kilowatt battery charging equipment, and at high-power-factor-rectification techniques for aircraft systems.

Devising high-performance methods of control for power electronic converters forms an integral part of this research area, for example: multi-loop control of resonant converters, control of resonant high-power-factor-rectifiers, application of high speed digital systems and adaptive control strategies.

Switching power converters are designed to convert electrical power from one form to another with high efficiency.

The high efficiency is obtained by using only switching devices, energy storage elements and transformers (all of which are ideally lossless), and relying on appropriate modulation of the switches to convert the available ac or dc voltage/current waveforms of the power source into (approximately) the ac or dc waveforms required by the load. The switches are generally semiconductor devices: diodes, thyristors, bipolar transistors of various types (operating at cutoff or saturation, not in their active region), metal–oxide–semiconductor field-effect transistors (MOSFETs) and so on.

The key to high efficiency and versatility of power electronic converters is the availability of fast, nearly ideal switches. Given the substantial frequency separation between fundamental components of waveforms involved in energy conversion (dc, or sub-kilohertz ac) and the switching frequency (tens of kilohertz or more), there exists a considerable freedom in selecting the switching waveforms that satisfy the (low frequency) energy conversion constraints.

This degree of freedom can be utilized to meet various additional functional requirements, and it has been explored in a number of directions over the last two decades. This paper reviews one such direction, namely spectral shaping achieved by adding a stochastic aperiodic) component to the switching pattern. More precisely, the paper reviews analytical concepts and practical circuit implementations of randomized switching schemes which shape the power spectra of the various switching functions to achieve EMI and acoustic noise ompliance.

An interesting feature of power electronic circuits is that, depending on the application, the same basic circuit can be modified with additional elements or used with different control methods that provide additional functionality or work better at the power levels demanded by the application. With this versatility in mind, we next review the most important power converter configurations; our review follows closely [2]–[4]. Consider first the case of a dc/dc converter—given a dc voltage of value (which can represent an input dc voltage, or an output dc voltage, or a dc difference between input and output voltages), we can easily arrange for a controlled switch to "chop" the dc waveform into a pulse waveform that alternates between the values and 0 at the switching frequency. Subsequently, the pulse waveform will be lowpass-filtered with capacitors and/or inductors that are configured to respond to its average value, i.e., its dc component. This dc component is controlled by the modulation of the switching pattern—by controlling the duty ratio of the switch, i.e., the fraction of time that the switch is closed in each cycle, we can control the fraction of time that the pulse waveform takes the value , and thereby control the dc component of this waveform. This control approach is referred to as pulsewidth modulation (PWM).

The class of switching regulators or switched-mode converters or high-frequency PWM dc/dc converters is based on this principle. Switching frequencies in the range of 20– 300 kHz are typically used today. Appropriate control of a high-frequency PWM dc/dc converter also enables conversion between waveforms that are not dc, but that are slowly varying relative to the switching frequency. If, for example, the input is a slowly varying unidirectional voltage—such as the waveform obtained by rectifying a 60-Hz sinewave—while the converter is switched at a much higher rate, say 50 kHz, then we can still arrange for the output of the converter to be essentially dc. The result would be a so-called active, or PWM rectifier. In a high-frequency PWM inverter, the situation is reversed. The essence of it is still a dc/dc converter, and the input to it is dc. However, the switching is controlled in such way that the filtered output is a slowly varying rectified sinusoid at the desired frequency. This rectified sinusoid can then be "unfolded" into the desired sinusoidal ac waveform, hrough the action of additional controllable switches arranged in a bridge configuration. In practice both the chopping and unfolding functions can be carried out by the bridge switches, and the resulting high-frequency PWM bridge inverter is the most common implementation, available in single-phase and three-phase versions. These inverters are often found in drives for ac servo-motors, such as the permanent-magnet synchronous motors (also called "brushless dc" motors) and induction motors that are popular in servo applications.

The inductive windings of the motor perform all or part of the electrical lowpass filtering in this case, while the motor inertia provides the additional mechanical filtering that practically removes the switching-frequency component from the mechanical motion.

A diode bridge is typically used to convert an ac waveform into a unidirectional or rectified waveform. If controllable switches are used instead of diodes, it is possible to partially rectify a sinusoidal ac waveform, with subsequent lowpass filtering to obtain an essentially dc waveform at a specified level. This is the operating principle of phase-controlled rectifiers, which are used as drives for dc motors or as battery charging circuits. It is also possible to interface two ac systems through a power electronic converter. An intricate use of switches—in a cycloconverter—permits the construction of an approximately sinusoidal waveform at some specified frequency by "splicing" together appropriate segments of a set of three-phase (or multi-phase) sinusoidal waveforms at a higher frequency; again, subsequent filtering improves the quality of the output sinusoid. While cycloconverters and matrix converters achieve a direct ac/ac conversion, it is more common today to construct an ac/ac converter as a cascade of a rectifier and an inverter (generally operating at different frequencies), forming a dc-link converter.

Our discussion of randomized modulation is concerned with the steady-state behavior of switching functions and, in particular, with their power spectrum. Nevertheless, the random nature of such functions results in occasional local deviations from the desired steady-state behavior, which cannot be adequately characterized in the frequency domain. While the technique of Markov chain based switching makes it possible to impose limits on the time-domain ripple, there is still a need for an accurate characterization and control of local transients.

The theory of Markov renewal processes provides the mathematical framework for a probabilistic time-domain analysis of switching functions, including both transient and steady-state aspects of their behavior; because of space limitations, we focus in the current paper mainly on steady-state characteristics. The same analytical framework can also be applied to more general schemes, such as Markov chain based switching with randomized selection of pulse shape for each state.

Our final remark addresses the system theoretic nature of the model used to generate the switching function. Markov chain based switching can be described in the time domain by a state-space model with nonlinear state dynamics and partially- linear output. Fortunately, the time evolution of the underlying probability distributions obeys linear difference/differential equations, which greatly simplifies the probabilistic analysis of randomized switching techniques.

3.3 Rectifier Circuits

Now we come to the most popular application of the diode: rectification. Simply defined, rectification is the conversion of alternating current (AC) to direct current (DC). This almost always involves the use of some device that only allows one-way flow of electrons. As we have seen, this is exactly what a semiconductor diode does. The simplest type of rectifier circuit is the half-wave rectifier, so called because it only allows one half of an AC waveform to pass through to the load:

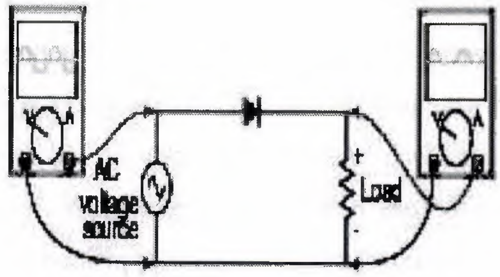


Figure 3.1 Half-Wave Rectifier Circuit.

For most power applications, half-wave rectification is insufficient for the task. The harmonic content of the rectifier's output waveform is very large and consequently difficult to filter. Furthermore, AC power source only works to supply power to the load once every half-cycle, meaning that much of its capacity is unused. Half-wave rectification is, however, a very simple way to reduce power to a resistive load. Some two-position lamp dimmer switches apply full AC power to the lamp filament for "full" brightness and then half-wave rectify it for a lesser light output:

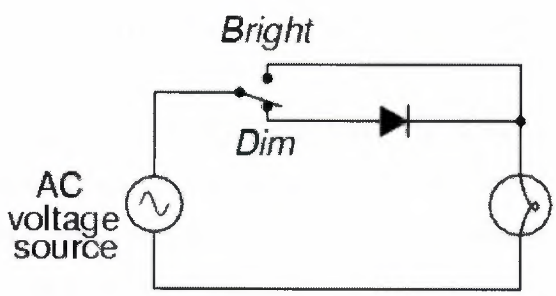


Figure 3.2 Two Position Lamp Dimmer Switch.

In the "Dim" switch position, the incandescent lamp receives approximately one-half the power it would normally receive operating on full-wave AC. Because the half-wave rectified power pulses far more rapidly than the filament has time to heat up and cool down, the lamp does not blink. Instead, its filament merely operates at a lesser temperature than normal, providing less light output. This principle of "pulsing" power rapidly to a slow-responding load device in order to control the electrical power sent to it is very common in the world of industrial electronics. Since the controlling device (the diode, in this case) is either fully conducting or fully nonconducting at any given time, it dissipates little heat energy while controlling load power, making this method of power control very energy-efficient. This circuit is perhaps the crudest possible method of pulsing power to a load, but it suffices as a proof-of-concept application.

If we need to rectify AC power so as to obtain the full use of both half-cycles of the sine wave, a different rectifier circuit configuration must be used. Such a circuit is called a full-wave rectifier. One type of full-wave rectifier, called the center-tap design, uses a transformer with a center-tapped secondary winding and two diodes, like this:

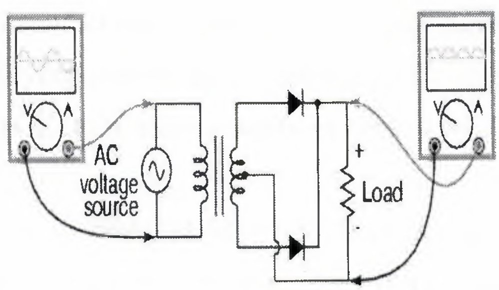


Figure 3.3 Full-Wave Rectifier Circuit (Center-Tap Design).

This circuit's operation is easily understood one half-cycle at a time. Consider the first half-cycle, when the source voltage polarity is positive (+) on top and negative (-) on bottom. At this time, only the top diode is conducting; the bottom diode is blocking current, and the load "sees" the first half of the sine wave, positive on top and negative on bottom. Only the top half of the transformer's secondary winding carries current during this half-cycle:

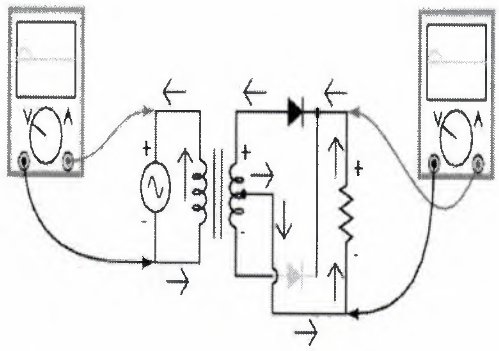


Figure 3.4 Circuit Operation with Transformer.

During the next half-cycle, the AC polarity reverses. Now, the other diode and the other half of the transformer's secondary winding carry current while the portions of the circuit formerly carrying current during the last half-cycle sit idle. The load still "sees" half of a sine wave, of the same polarity as before: positive on top and negative on bottom:

One disadvantage of this full-wave rectifier design is the necessity of a transformer with a center-tapped secondary winding. If the circuit in question is one of high power, the size and expense of a suitable transformer is significant. Consequently, the center-tap rectifier design is seen only in low-power applications.

Another, more popular full-wave rectifier design exists, and it is built around a four-diode bridge configuration. For obvious reasons, this design is called a full-wave bridge:

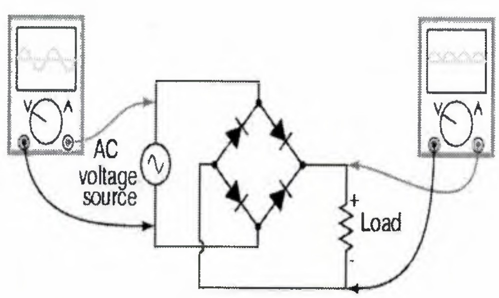


Figure 3.5 Full-Wave Rectifier Circuit (Bridge Design).

Current directions in the full-wave bridge rectifier circuit are as follows for each half-cycle of the AC waveform:

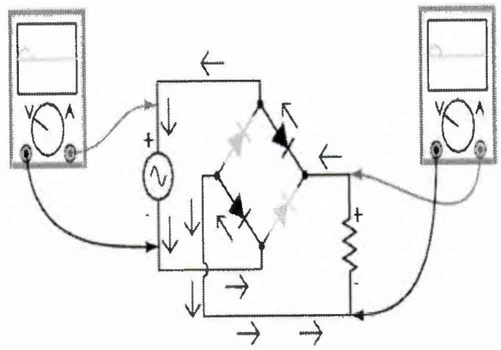


Figure 3.6 Circuit Operation of Full-Wave Rectifier (Bridge Design).

Remembering the proper layout of diodes in a full-wave bridge rectifier circuit can often be frustrating to the new student of electronics. I've found that an alternative representation of this circuit is easier both to remember and to comprehend. It's the exact same circuit, except all diodes are drawn in a horizontal attitude, all "pointing" the same direction:

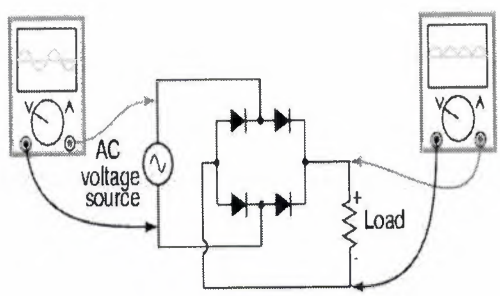


Figure 3.7 Full-Wave Bridge Rectifier Circuit (Alternative Layout).

One advantage of remembering this layout for a bridge rectifier circuit is that it expands easily into a polyphase version:

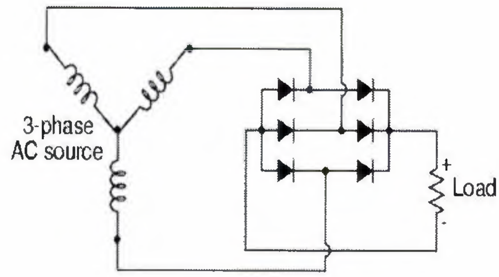


Figure 3.8 Three-Phase Full-Wave Bridge Rectifier Circuit.

Each three-phase line connects between a pair of diodes: one to route power to the positive (+) side of the load, and the other to route power to the negative (-) side of the load. Polyphase systems with more than three phases are easily accommodated into a bridge rectifier scheme. Take for instance this six-phase bridge rectifier circuit:

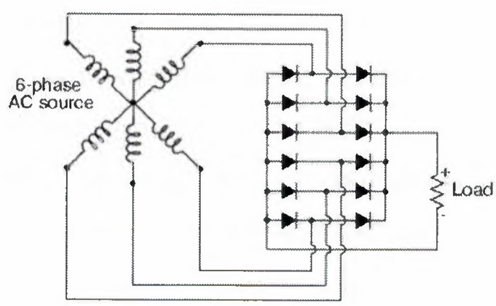


Figure 3.9 Six-phase Full-Wave Bridge Rectifier Circuit.

When polyphase AC is rectified, the phase-shifted pulses overlap each other to produce a DC output that is much "smoother" (has less AC content) than that produced by the rectification of single-phase AC. This is a decided advantage in high-power rectifier circuits, where the sheer physical size of filtering components would be prohibitive but low-noise DC power must be obtained. The following diagram shows the full-wave rectification of three-phase AC:

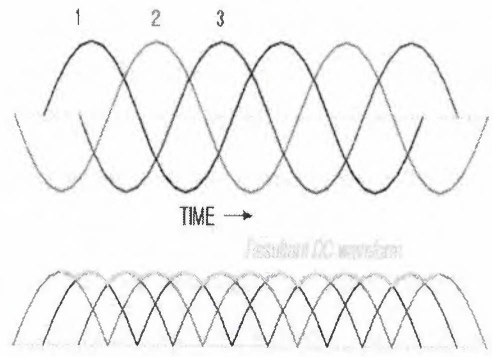


Figure 3.10 Full-Wave Rectification of Three-Phase AC Wave.

In any case of rectification (single-phase or polyphase) the amount of AC voltage mixed with the rectifier's DC output is called ripple voltage. In most cases, since "pure" DC is the desired goal, ripple voltage is undesirable. If the power levels are not too great, filtering networks may be employed to reduce the amount of ripple in the output voltage.

Sometimes, the method of rectification is referred to by counting the number of DC "pulses" output for every 360° of electrical "rotation." A single-phase, half-wave rectifier circuit, then, would be called a 1-pulse rectifier, because it produces a single pulse during the time of one complete cycle (360°) of the AC waveform. A single-phase, full-wave rectifier (regardless of design, center-tap or bridge) would be called a 2-pulse rectifier, because it outputs two pulses of DC during one AC cycle's worth of time. A three-phase full-wave rectifier would be called a 6-pulse unit.

Modern electrical engineering convention further describes the function of a rectifier circuit by using a three-field notation of phases, ways, and number of pulses. A single-phase, half-wave rectifier circuit is given the somewhat cryptic designation of 1Ph1W1P (1 phase, 1 way, 1 pulse), meaning that the AC supply voltage is single-phase, that current on each phase of the AC supply lines moves in one direction (way) only, and that there is a single pulse of DC produced for every 360° of electrical rotation. A single-phase, full-wave, center-tap rectifier circuit would be designated as 1Ph1W2P in this notational system: 1 phase, 1 way or direction of current in each winding half, and 2 pulses or output voltage per cycle. A single-phase, full-wave, bridge rectifier would be designated as 1Ph2W2P: the same as for the center-tap design, except current can go both ways through the AC lines instead of just one way. The three-phase bridge rectifier circuit shown earlier would be called a 3Ph2W6P rectifier.

Is it possible to obtain more pulses than twice the number of phases in a rectifier circuit? The answer to this question is yes: especially in polyphase circuits. Through the creative use of transformers, sets of full-wave rectifiers may be paralleled in such a way that more than six pulses of DC are produced for three phases of AC. A 30° phase shift is introduced from primary to secondary of a three-phase transformer when the winding configurations are not of the same type. In other words, a transformer connected either Y-Δ or Δ-Y will exhibit this 30° phase shift, while a transformer connected Y-Y or Δ-Δ will not. This phenomenon may be exploited by having one transformer connected Y-Y feed a bridge rectifier, and have another transformer connected Y-Δ feed a second bridge rectifier, then parallel the DC outputs of both rectifiers. Since the ripple voltage waveforms of the two rectifiers' outputs are phase-shifted 30° from one another, their superposition results in less ripple than either rectifier output considered separately: 12 pulses per 360° instead of just six:

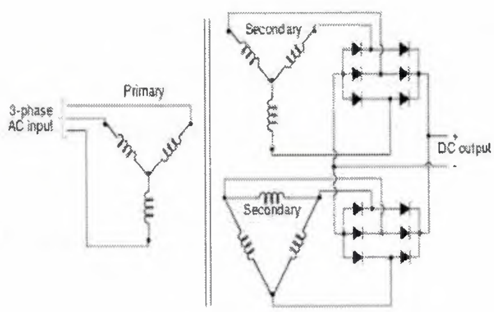


Figure 3.11 3Ph2W12P Rectifier Circuit.

3.3.1 Rectification Efficiency

Rectification efficiency measures how efficiently a rectifier converts a.c. to d.c. It is defined as the ratio of the d.c. output power to a.c. input power, where d.c. output power is a product of the average current and voltage.

3.4 DC Choppers

In many industrial application, it is required to convert a fixed-voltage dc source into a variable-voltage dc source.A dc chopper converts directly from dc to dc and is also known as a dc-to-dc converter.A chopper can be considered as a dc equivalent to an ac transformer with a continuously variable turns ratio.Like a transformer,it can be used to step-down or step-up a dc voltage source.

Choppers are widely used for traction motor control in electric automobiles,trolley cars,marine hoists,and forklift trucks.They provide smooth acceleration control,high efficiency,and fast dynamic response.Choppers can be used in regenerative braking of dc motors to return energy back into the supply,and this feature results in energy savings for transportation systems with frequent stops.Choppers are used in dc voltage regulators,and also used,in conjunction with an inductor,to generate a dc current source,especially for the current source inverter.

3.4.1 Chopper Classification

Depending on the directions of current and voltage flows,chopper can be classified into five types:

3.4.1.1 Class A Chopper: The load current flows into the load.Both the load voltage and the load current are positive.

3.4.1.2 Class B Chopper: The load current flows out of the load.The load voltage is positive ,but the load current is negative.

3.4.1.3 Class C Chopper: The load current is either positive or negative.The load voltage always positive.The class A and class B choppers bcan be combined to form a class C chopper.

3.4.1.4 Class D Chopper: The load current is always positive.The load voltage is either positive or negative.A class D chopper can also operate either as a rectifier or as an inverter.

3.4.1.5 Class E Chopper: The load current is either positive or negative.The load voltage is also either positive or negative.Two class C choppers can be combined to form a class E chopper.

3.5 DC-DC Converters

A dc-dc converter is also known as a chopper or switching regulator. A transistor chopper is shown in figure 3.12. The average output voltage is obtained by controlling the conduction time t of transistor Q1. If T is the chopping period, then t1 = aT. a is called the duty cycle of the chopper.

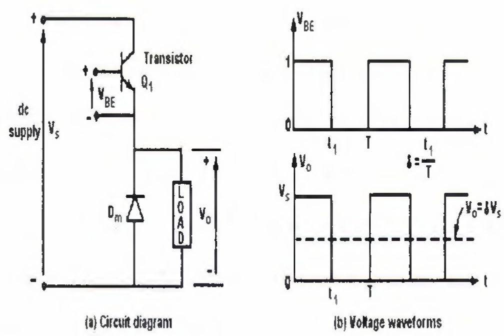


Figure 3.12 Schematic Diagrams of DC-DC Converter.

3.6 Inverter

Recent advances in modern power semiconductor device technologies have led to high utilization of power converters in a large number of applications and have opened up a host of new converter topologies for many new applications. The most visible gain in industrial and commercial products is occurring in the area of power inverters, which convert a dc voltage into a single or polyphase ac voltage at a desired amplitude and frequency. Technology advances in these areas 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.

Today, the power converter topology of choice for ac output applications is the 'hard switching' dc/ac voltage source inverter shown in Fig 3.13. The ac output voltage is synthesized using a pulse width modulated (PWM) switching waveform, which has a controlled amplitude low-frequency 'fundamental' component, and high-frequency modulation components. The modulation components of the current are filtered by the low pass characteristic of typical inductive power electronic loads.

A key factor in reducing the size of reactive components used for filtering and energy storage, improving transient performance and meeting stringent harmonic specifications is the switching frequency of the inverter. Although IGBTs switch rapidly, switching losses occur during device turn-on and turn-off due to the transient existence of both voltage across, and current in the device. These stresses require significant device derating for switching frequencies in excess of 5 - 6 kHz, thus increasing system cost. In addition, IGBT losses are further increased at turn-on by the charge stored in the complementary switch's anti-parallel diode and at turn off by the energy trapped in the parasitic inductance of the IGBT package and device interconnections.

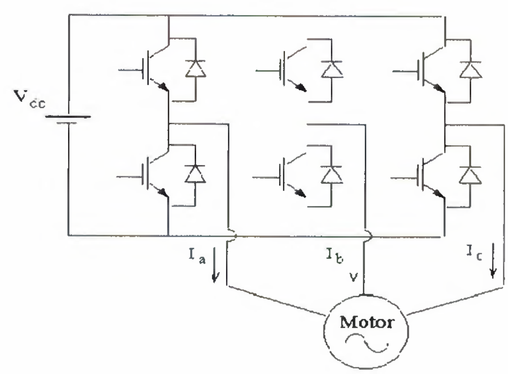


Figure 3.13 The Voltage Source Inverter (VSI).

Another issue is the transient on the inverter output voltage caused by IGBT switching resulting in dv/dt's in excess of 5-10,000 volts/μs. Impressing such high dv/dt across motor loads can cause severe problems and results in transient voltages of twice the nominal value across motor windings, which can cause winding insulation breakdown. Also associated with the high switching speed is the broad band electro-magnetic interference (EMI) that is generated on the inverter output. This EMI has frequency content spanning from 10 kHz to 30 MHz and is difficult to suppress.

In order to obtain additional system improvements, a fundamentally different approach is needed. One technique that has demonstrated promise is soft switching. This paper presents a brief discussion and comparison of various soft switching inverters. In addition, substantive data verifying the performance and attributes of the actively clamped resonant dc link inverter will also be presented.

3.6.1 Resonant Pole Inverters

One of the earliest soft switching topologies proposed for dc-to-ac inverters is the Resonant Pole Inverter (RPI) [1]. A phase leg of the RPI is shown in Fig 3.14. In this figure, an inverter pole consisting of switches S1 and S2 is configured. In order to achieve ZVS, a resonant inductor is placed in series with a filter capacitor across which the load is placed. The inverter phase voltage, Vf, can be modulated to generate the desired low frequency voltage waveform.

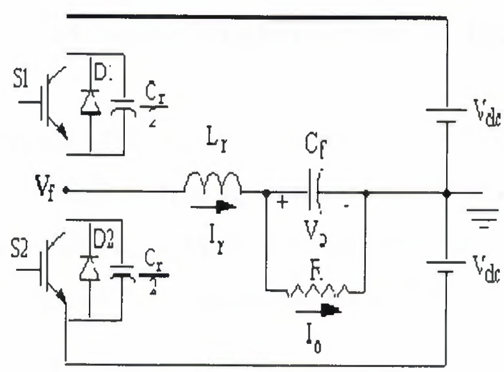


Figure 3.14 The Resonant Pole Inverter (RPI) Phase Leg.

The inverter operation is explained as follows. Assuming switch S1 conducting, the state equations characterizing the systems are given by,

$$\frac{d}{dt}I_r = \frac{V_{dc} - V_o}{L_r}$$

$$\frac{d}{dt}V_o = \frac{I_r - V_o/R}{C_f}$$

Normally, if the switching frequency is much higher than the fundamental frequency of Vo, the output voltage can be assumed to be constant over a switching cycle. As a result, Ir will ramp up in a linear fashion as shown in Fig. 3.15 (M1). When switch S1 is turned off, a resonant transition cycle is initiated where the resonant inductor resonates with the output capacitances of switches S1 and S2, namely, Cr. This charges the output capacitance of S1 and discharges the output capacitance of S2 causing D2 to conduct (Fig.3.15 (M2)). If the inductor current is assumed to be nearly constant, the switches' output capacitances will charge/discharge in a linear fashion. When D2 conducts, S2 can be turned on under ZVS conditions.

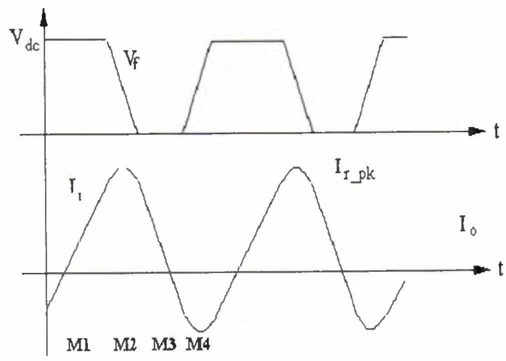


Figure 3.15 Typical Waveforms of the RPI Phase Leg.

In order to guarantee ZVS for the inverter switches, two conditions need to be met. First, the current, Ir, must be flowing in the switch to be turned off. Second, there needs to be a minimum current, Ir_min, to ensure that the pole voltage will reverse. Assuming that Cf >>Cr, the ZVS condition can be stated as,

$$I_{r\_min} = \frac{2}{Z_o}\sqrt{V_{dc}V_o}$$

where $$Z_o = \sqrt{L_r/C_r}\ .$$

When the antiparallel diode of S2 conducts, the switch S2 can be turned on under ZVS. An example showing the synthesis of an ac inductor current with a RPI is shown in Fig.3.16

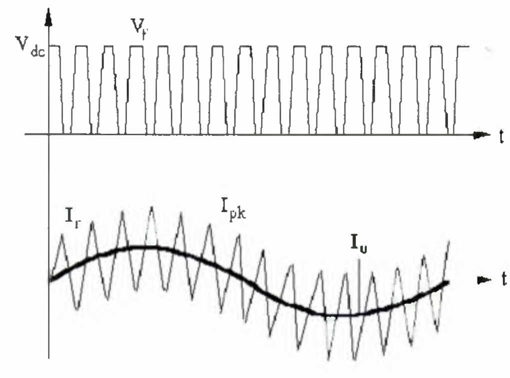


Figure 3.16 Synthesis of an AC Inductor Current in the RPI.

The operation of the RPI shows that in order to achieve ZVS, the current has to commutate from the main diodes to the main switches. This commutation can be maintained as long as the amplitude of the output voltage, Vo, is lower than Vdc. The switching frequency can be approximately derived to be,

$$f_s \cong \frac{V_{dc}^2 - V_o^2}{4V_{dc}L_r I_o}$$

By investigating equation, it is clear that as the output voltage varies from 0 to 0.8Vdc, the switching frequency varies from a maximum, fmax to 0.6fmax. When Vo=Vdc, the switching frequency is zero. This sets a limit on the maximum output voltage attainable at the lowest switching frequency.

Another aspect of the RPI is the requirement that the peak device current is 2.2 to 2.5 p.u. This is due to the fact that in order to achieve ZVS, the resonant current must reverse polarity at the peak of the output current to insure that the antiparallel diodes conduct prior to device turn on. Normally, device cost, especially IGBTs, is proportional to the current turn off capability. As a result, the device cost of a RPI is 2.0 p.u. compared with a hard switched VSI.

3.6.2 The Resonant DC Link Inverter

One of the earliest and most mature of a large number of soft switching inverter topologies that have been proposed is the resonant dc link (RDCL) inverter. The basic version of the resonant dc link inverter is shown in Fig.3.17.

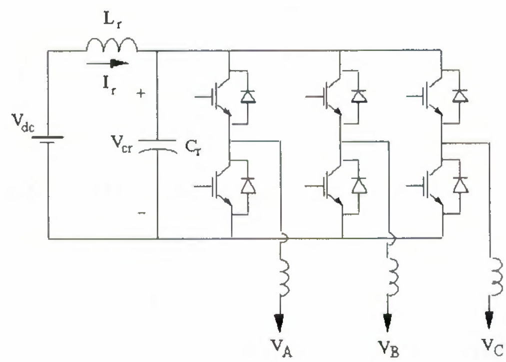


Figure 3.17 The Resonant DC Link Inverter (RDCLI).

In the RDCLI, the voltage across the resonant capacitor is also impressed across the six power devices. This voltage has an average or dc value which is equal to the dc bus, Vdc, and an oscillating or resonant component. The combined voltage is referred to as the resonant link. The resonant link is 'excited' and maintains resonance through appropriate control of the inverter switches such that the resonating dc bus voltage periodically reaches zero volts. Switching of the devices is synchronized to the link zero crossings to obtain the desired low switching loss.

The main inverter devices are only allowed to change state at the link voltage zero crossings. This forces the inverter output to consist of an integral number of resonant link voltage pulses, a significantly different strategy from the pulse width modulation used in conventional hard switching inverters. The desired low frequency output voltage now has to be synthesized using discrete resonant pulses, using a discrete pulse modulation (DPM) strategy. Typical low voltage synthesis is shown in Fig.3.18.

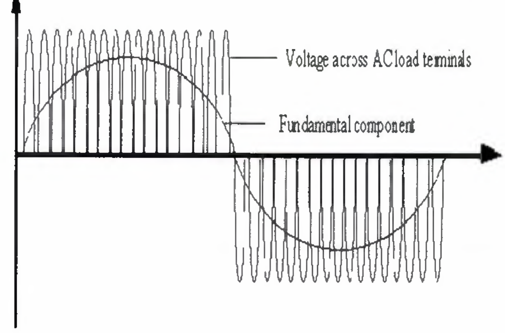


Figure 3.18 Typical Line-to-Line Voltage Synthesis Using DPM.

In order to simplify the analysis of a RDCLI, and since the resonant frequency is much higher than the fundamental frequency of the synthesized waveform, an equivalent circuit of the system during each resonant pulse is shown in Fig.3.19. Note here that the load current Ix, is assumed to be constant during a resonant pulse duration.

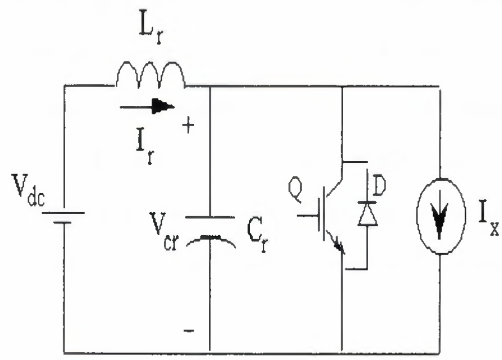


Figure 3.19 Equivalent Circuit of the RDCLI During Each Resonant Pulse.

If the switch Q is turned off, applying Vdc to the circuit results in a resonant cycle and the capacitor voltage Vcr is given by,

$$V_{cr}(t) = V_{dc}(1 - \cos \omega t)$$

where w is the resonant frequency of the LC circuit. When wt=2p, the capacitor voltage goes back to zero setting up a zero voltage switching condition for the switch Q. When the switch is turned on, the inductor current will ramp up in a linear fashion. Sufficient energy has to be stored in Lr before the switch is turned off to ensure that the capacitor voltage will return to zero. Typical waveforms with zero load current are shown in Fig.3.20.

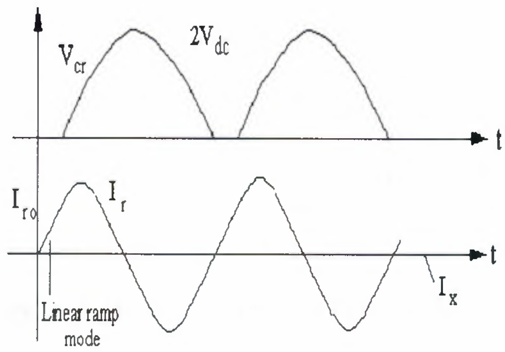


Figure 3.20 Typical Waveforms of the RDCLI With Ix=0.

The value of the current Ix depends on the individual phase currents and the switching functions of the six inverter devices. Note here that Ix can change significantly from a switching cycle to the next depending on the switching strategy. However, during the resonant cycle itself, Ix remains fairly constant since the inverter states are preserved. In order to ensure ZVS, the inverter actively controls the current (Ir-Ix) to ensure that the resonant cycle starts with the same initial conditions. Hence, the resonant cycle is controlled in a dead beat fashion and is independent of the value of Ix.

As devices are switched, the L-C resonant circuit excitation initial conditions are changed. This can result in high peak voltage stresses. Typically, the main devices voltage stress in a RDCLI is 2Vdc. Consequently, the resonant dc link circuit is always used with a means to limit the peak voltage stress across the device. Two possible variations include the passively clamped and the actively clamped resonant dc link inverters shown in Figs.3.21 and 3.22, respectively.

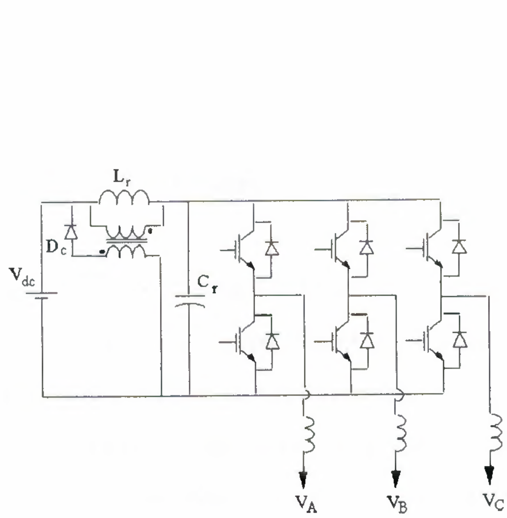


Figure 3.21 The Passively Clamped RDCLI.

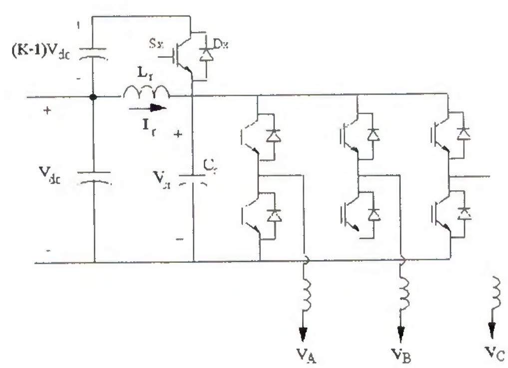


Figure 3.22 The Actively Clamped RDCLI (ACRDCLI).

The value of the clamping ratio K can be related to the ratio of the resonant frequency, fo, to the switching frequency, fs, and is given by,

$$\frac{\pi \cdot f_o}{f_s} = \cos^{-1}(1 - K) + \frac{\sqrt{K(2 - K)}}{K - 1}$$

For clamping voltages less than 2Vdc, the value of the clamp ratio, K, is inversely proportional to the ratio of the resonant tank frequency to the switching frequency. The link frequency decreases with decreasing K, approaching zero for K=1. This suggests that with a given set of components, there is a limit below which clamping is not practical. However, component design can be modified to obtain the frequencies of interest.

3.6.3 Quasi Resonant DC Link Inverters

Another approach which maintains PWM control and achieves soft switching is the quasi resonant DC link (QRDCLI) type inverters. Unlike the RDCLI, where the resonant bus is continuously oscillating, the resonant bus is clamped to a near constant value during the power delivery mode and goes into a state of resonance when the pulse is terminated or during the rise of the bus voltage to the clamp level. The resultant near-square wave pulses are similar to the conventional hard switched inverters.

One of the potential QRDCLI is the voltage clamped parallel resonant converter (VCPRC) shown in Fig.3.23. In this case, the conduction through Dz and Sx extends the pulse width of the resonant bus voltage, Vcr.

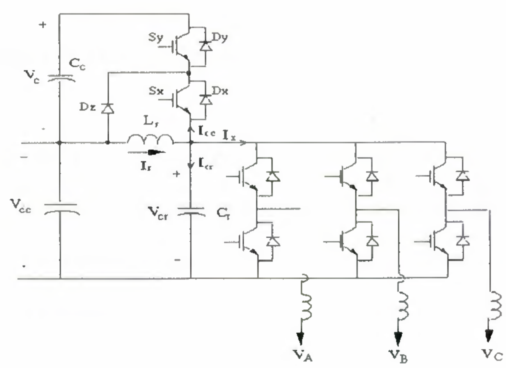


Figure 3.23 Voltage Clamped Parallel Resonant Converter.

This converter is a variation of the actively clamped resonant dc link, shown earlier in Fig.3.22, with the addition of Sy, Dy and Dz which are primarily used to extend and control the pulse width of the resonant bus voltage, Vcr. These additional devices enable the inverter to realize PWM control schemes. Typical waveforms of the inverter are shown in Fig.3.24.

The operation of the inverter is explained next. Starting with Vcr = 0 in mode M1, all of the inverter switches are shorted to ramp up current in the resonant inductor. This is done to overcome the resonant tank losses and ensure that Vcr returns to zero when the pulse is terminated. At the start of M2, the main devices are gated into the desired switch state, and a resonance cycle between Lr and Cr is initiated. Once Vcr reaches the clamp voltage level, Dx and Dy clamp the resonant bus voltage at Vdc+Vc

(M3). Once Dx and Dy conduct, Sx and Sy are turned on under a zero voltage condition. This causes the resonant current to ramp down and reverse as shown in Fig.3.24.

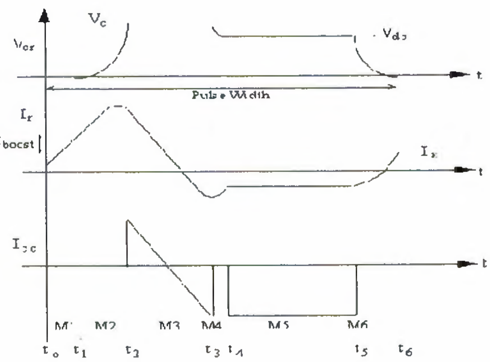


Figure 3.24 Typical Waveforms of the VCPRC.

As the resonant current reverses, the current is diverted into the auxiliary switches Sx and Sy. Note here that enough energy has to be stored in Lr to ensure that Vcr resonates back to zero when the pulse is terminated. The minimum current required to guarantee the reset of the resonant capacitor voltage is given by,

$$I_{min} = I_x - I_r = \frac{\sqrt{V_{dc}^2 - V_c^2}}{Z_o}$$

At the end of M3, both Sx and Sy may be turned off thus allowing Vcr to resonate back to zero and producing the minimum pulse width possible. This is similar to the ACRDCLI discussed earlier. However, if only Sy is turned off (M4), Vcr resonates down to Vdc and is clamped at this level by Dz. The pulse width of Vcr can now be extended by keeping Sx on and allowing Ir to free wheel through Dz (M5). During this mode, both Ir and Ix flow through Dz and Sx. To terminate the pulse, Sx is gated off at which point Vcr resonates back to zero.

The objective of quasi resonant link invertes (QRDCLI) is to control the pulse width of the resonant bus voltage to obtain the desired spectral purity. These converters have less degrees of freedom compared with their resonant pole counterparts and are thus more constrained in obtaining a specified spectrum. Since only one pulse width is variable, it is difficult to control a three phase three wire output completely. If two pulse widths are independently controllable, it is possible, at least in principle, to control the output space vector at any desired value. However, if the frequency is to be maintained constant, then three pulse widths need to be controlled independently.

One advantage of quasi resonant link inverters is that they can be controlled by conventional PWM techniques. The benefits of this include both the high efficiency of zero-voltage switching and a familiar spectral performance of PWM schemes. However, full bus utilization and superior spectral performance (high switching frequencies) are not attainable simultaneously. This is due to the fact that there is a minimum pulse width that can be realized by these inverters. This point is illustrated using the VCPRC as an example.

A space vector PWM scheme is used, in which one switching period is composed of two active states and one zero state. In the VCPRC, a resonant frequency of 65kHz is assumed thus yielding a minimum pulse width of 15 msec. Using a switching frequency of 16kHz, the voltage vectors that can be achieved are shown in Fig.3.25. Notice here that full bus utilization is not attainable because of the minimum zero state vector (as with the resonant pole inverter). In addition, as the voltage vector transitions from one state to another, there is a region of unattainable vectors. Thus, the spectral performance of this modulation scheme will be degraded from a pure PWM spectrum.

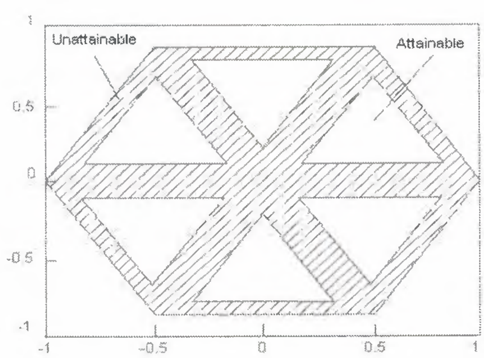


Figure 3.25 Voltage Vectors with a Space Vector Modulator. Tmin = 15msec, fs = 16kHz.

Figure 3.26 shows the voltage vectors achievable with a lower switching frequency of 4kHz. While more voltage vectors are attainable, it is not necessary to invoke soft switching for the devices at such low switching frequencies. Furthermore, with the same devices and resonant components, it would be possible to switch a resonant link inverter at approximately 65 kHz and yet achieve a spectral performance which is much better than that of the 'PWM inverter' switching at 4 kHz. This simplified analysis demonstrates that quasi-resonant dc link inverters typically cannot match the performance of the basic resonant dc link inverters. Further, the simplicity and lower parts count of the basic resonant dc link circuit makes this approach a more cost effective solution.

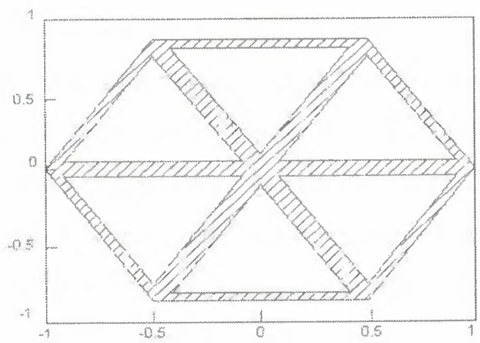


Figure 3.26 Voltage Vectors with a Space Vector Modulator. Tmin = 15msec, fs = 4kHz.

3.6.4 Resonant Snubber Inverters

Another soft switching technique employs resonant snubbers or lossless active snubbers to achieve ZVS. Capacitive snubbers are normally utilized to achieve ZVS for the main devices. However, unlike dissipative snubbers, the snubber energy is recovered in a lossless manner.

One example of a resonant snubber inverter (RSI) is shown in Fig.3.27. In this topology, soft switching snubber circuits are added to each leg of the inverter. Each snubber circuit consists of a resonant inductor along with an auxiliary switch and an antiparallel diode

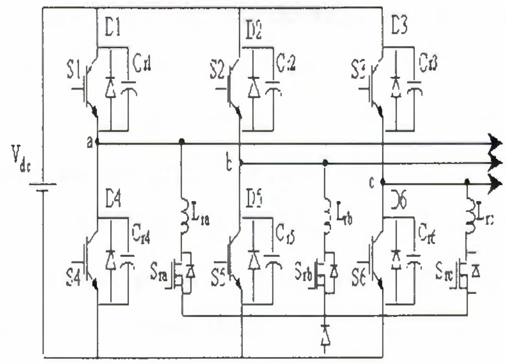


Figure 3.27 Resonant Snubber PWM Inverter (RSI).

Starting with D2, D3, and D4 conducting, in order to turn off D4 and turn on S1 under ZVS, the auxiliary switches Srb and Src are turned on causing a charging current through Lra, Lrb and Lrc. When the current in Lra is higher the load current, the current commutates from D4 to S4. Now, S4 can be turned off thus initiating a resonant cycle between Lra and the snubber capacitors across the main switches. Eventually, the voltage across S1 is clamped to zero and its antiparallel diode conducts so that S1 can be turned on under ZVS. At this point, the auxiliary currents gradually ramp down to zero. Note here that the auxiliary devices are switched under current condition. Figure 16 shows typical waveforms of a ZVS commutation.

In order to insure ZVS commutation, a minimum transition time is required. The total commutation time includes a charging phase, a resonant phase and a discharging phase and is given by,

$$T_r = \frac{3\sqrt{3} \cdot I_m L_r}{V_{dc}} + \pi\sqrt{2L_r C_r}$$

where Im is the maximum peak phase current.

One of the limitations of this topology is the maximum attainable switching frequency and low DC bus utilization. This is due to the minimum dwell time required to ensure ZVS transition. Another drawback of this topology is due to the fact that in a zero voltage state, where the top or bottom switches are conducting, the auxiliary circuit can not be triggered unless one of the active switches is turned off. This may result in additional switching actions and/or hard switching instants especially at light loads.

Another resonant snubber soft switched inverter topology is the zero voltage transition PWM inverter (ZVTI) shown in Fig.3.28. A soft switching commutation circuit which consists of a three phase rectifier operating in discontinuous conduction with an additional auxiliary switch and diode.

During most of the switching cycle, the auxiliary switch Sx is off and the diode Dx is blocking. When a diode in the main bridge needs to be turned off, the auxiliary switch Sx is turned on and the auxiliary inductors' current start increasing from zero. This causes the current in the main diodes to gradually decrease to zero.

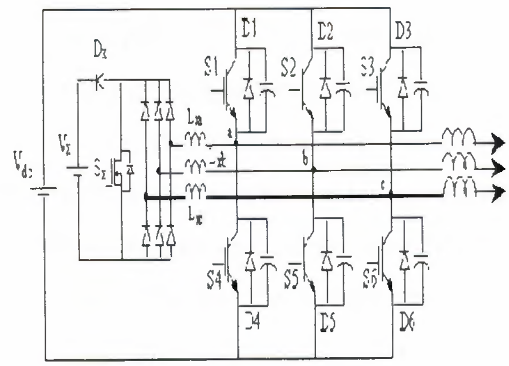


Figure 3.28 The Zero Voltage Transition PWM Inverter (ZVTI).

When the current is transferred to the antiparallel conducting switch, the switch is turned off thus initiating a resonant cycle where the energy stored in the auxiliary inductors charge and discharge the snubber capacitors across the main switches swinging the node voltages Van, Vbn and Vcn to the opposite rail. This causes the antiparallel diodes of the incoming switches to turn on providing a ZVS condition for the incoming switches. The commutation circuit is then deactivated by turning off the auxiliary switch. The remaining energy stored in the auxiliary inductors is returned to the auxiliary dc source. Figure 18 shows the inverter waveforms for soft commutation from diodes D2 and D3 to S5 and S6, respectively. Here, ia is assumed to be negative while ib and ic are assumed to be positive. Initially, all top switches are conducting and the commutation is started by turning off switch S1 while turning on the auxiliary switch Sx.

The total commutation time consists of three distinct phases: a charging phase Tc, a resonant phase Tr and a discharging phase Td. The total commutation time Tx is given by,

$$T_x = \frac{3\sqrt{3}\ I_m L_x}{V_{dc}} + \pi\sqrt{L_x C}$$

where Im is the maximum peak phase current and C is the parasitic capacitance of nodes a, b, or c. It is clear from (9) that un upper limit exists on the maximum frequency attainable and the maximum DC bus utilization.

One drawback of the ZVTI is the fact that soft switching will be lost at light loads, i.e. near the zero crossings of phase currents. This is clear from the previous example where S1 has to be turned off first to start the commutation cycle. At light loads, the energy will not be enough to guarantee ZVS for S1 and S4.

Another drawback of this inverter topology is that more switching action of the main switches is required. With a zero voltage vector, no voltage is available to charge the resonant inductors which requires one of the switches to be turned off first. In addition, the switching timing in the ZVT commutation is critical which adds to the control complexity. Finally, the commutation circuit requires ultra fast recovery diodes to block reverse recovery currents and voltage clamping devices to avoid over-voltage across the auxiliary switch.

At higher power levels (> 1 MW), other topologies such as the Auxiliary Resonant Commutated Pole inverter, shown in Fig.3.29, may offer a more reasonable solution than the basic resonant dc link inverter, the ZVTI and RPI.

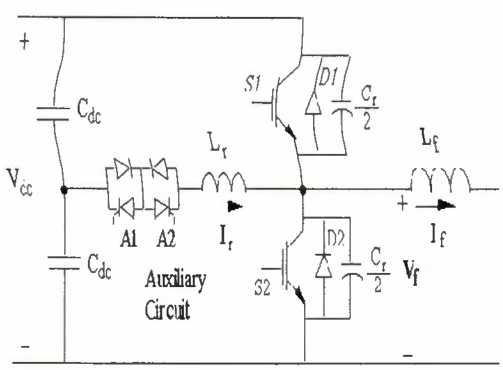


Figure 3.29 The Auxiliary Resonant Commutated Pole Inverter (ARCP).

An LC snubber circuit triggered by the auxiliary switches A1 and A2 are utilized to achieve ZVS for the main devices. A resonant current is initiated in the L-C circuit by gating an auxiliary device to ensure that the main devices are carrying some current, which can then be turned off under a zero voltage switching condition. The auxiliary devices turn-off when the current in the L-C circuit naturally reaches zero.

The basic operation of the ARCP is shown in Fig.3.30. The inverter voltage and the resonant current waveforms are shown for commutation from a main diode to a main switch and vise versa.

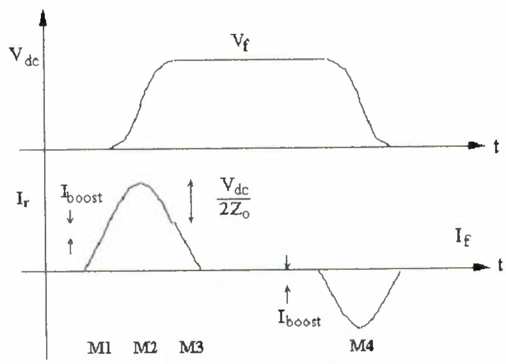


Figure 3.30 Typical Switching Waveforms of the ARCP Inverter.

Initially, the load current, If, is assumed to be positive and flowing out of D2 (Fig.3.31(a)). In order to start the commutation process, the switch A2 is gated on. This will impress a voltage of Vdc/2 across the inductor and hence the inductor current will ramp up with a slope of Vdc/2Lr (Fig.3.31(b)). During this time, the main switch S2 remains gated on.

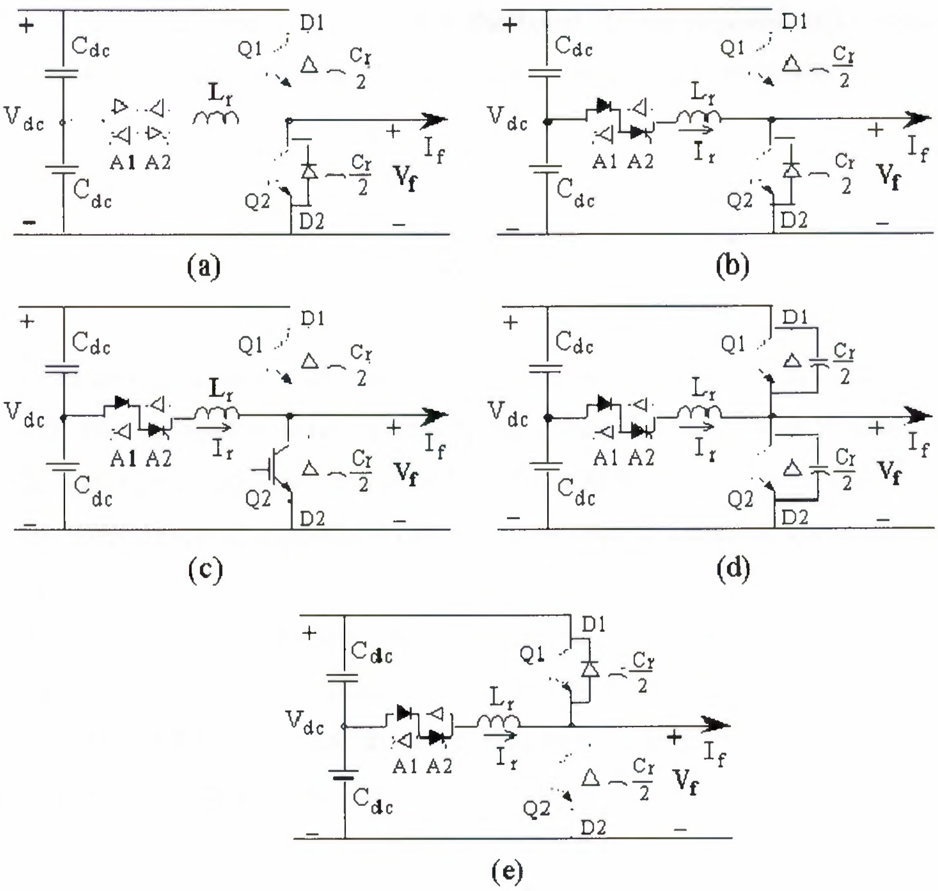


Figure 3.31 Circuit Modes of the ARCP During Commutation from D2 to S1.

When Ir exceeds If, the diode turns off and the boost phase will start where the current will flow in the main device (Fig.3.31(c)). When Ir reaches the required boost level, the device, S2, can be turned off. The resonant cycle will start and the pole voltage will swing to the opposite rail (Fig.3.31(d)). When Vf reaches Vdc, the diode D1 turns on and clamps the voltage to Vdc. The main switch S1 can now be turned on under ZVS (Fig.3.31(e)). Once D1 conducts, the inductor current will ramp down to zero with a slope of –Vdc/2Lr where the auxiliary device A2 can now be turned off under ZCS. The peak inductor current is approximately given by,

$$I_{r\_pk} = I_f + I_{boost} + \frac{V_{dc}}{2Z_o}$$

where $$Z_o = \sqrt{L_r / C_r} \ ,$$

If is the load current, and Iboost is the boost current required to overcome the resonant tank losses and ensure a ZVS transition. The total commutation time is given by,

$$T_c = 2L_r \frac{I_f + I_{boost}}{V_{dc}} + \pi\sqrt{L_r C_r}$$

In order to commutate from an active device to a diode, the switch can simply be turned off, and the load current will swing the voltage from one rail to another. But normally, a minimum load current should be flowing in the device in order to turn it off. So, at low load current, the auxiliary circuit can be used to assist the commutation of a main device.

In order to decide whether the auxiliary circuit is needed or not, a threshold current level Ifmin can be set. If the load current If is less than Ifmin , the auxiliary circuit will be used to assist the device commutation, otherwise, it is not used. The threshold current can be found from,

$$I_{f\_min} = \frac{C_r V_{dc}}{T_{max}}$$

where Tmax is the maximum commutating time (i.e. the time required to commutate off full load current from a conducting diode). Equation can be used to evaluate Tmax by setting If to be Ifmax.

Commutating a light load current from S1 to D2, the process starts by gating on A1, where the inductor current ramps up with a slope of Vdc/2Lr. Note that the inductor current now is in the other direction since A1 is on. When Ir reaches Iboost the switch S1 is turned off under ZVS. When S1 turns off, the resonant cycle starts, and the pole voltage will swing to the negative rail. When Vf reaches zero, the lower diode D2 will conduct and clamp Vf to zero. The inductor current will ramp down forced by –Vdc/2 to zero where the auxiliary device A1 can be turned off under ZCS. The peak inductor current will be approximately,

$$I_{r\_pk} = I_{boost} + \frac{V_{dc}}{2Z_c}$$

On the other hand, commutating a heavy load current from S1 to D2, the process starts by turning off S1 where the load current will be enough to swing the inverter voltage to the opposite rail.

ARCP inverters offer pulse width modulation capability on the inverter output at lower switching frequencies, and may be realized using IGBTs or gate turn-off thyristors, GTOs.

One of the drawbacks of the ARCP inverter is the minimum dwell time imposed by the reset cycle and slope variability on the output which limits the precision with which the inverter output spectral content can be specified. Fig.3.32 shows the effect of the minimum dwell time on the dc bus utilization for various switching frequencies. In this case, it is assumed that spectral purity is important and thus pulse dropping is not allowed. For a resonant pulse

duration of 4ms (Lr = 5mH and Cr = 0.33mF), Fig.3.32 shows that a duty cycle greater than 0.85 or lower than 0.15 cannot be obtained at 20kHz. This represents a significant constraint on the dc bus utilization, if spectral purity is important.

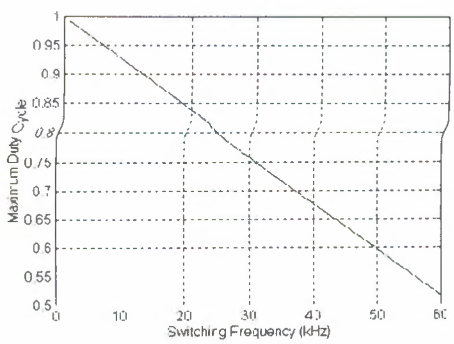


Figure 3.32 Maximum Duty Cycle Attainable v.s. Switching Frequency for an ARCP Inverter with Lr=5uH, Cr= 0.33Uf.

Another drawback of the ARCP inverter is the reverse recovery associated with the auxiliary switches since these swicthes are realized using thyristor based devices like MCTs. If the resonant inductor value, Lr, is low to reduce dwell time, the di/dt of the discharge current will be rather high and the reverse recovery current would be quite significant. Additional snubber circuitry would be needed to absorb the energy stored in the resonant inductor.

3.7 Summary

In this chapter,power electronic converters,rectifiers,choppers,inverters and their structural and electronical analysis are given by their diagrams and characteristics.

CHAPTER FOUR
MOTOR DRIVES

4.1 Overwiev

Electronic motor drives fall into one of two categories: AC and DC. AC motor drives control AC induction motors, DC motor drives typically control shunt-wound DC motors (which have separate armature and field circuits), and they both control the speed, torque, direction, and resulting horsepower of a motor.

In this chapter, motor drives with their properties and their protection rules will be examined.

4.2 DC Universal Motor Drive

A thyristor supplies the motor during the positive mains half cycle.Both the thyristor and its control are connected in such a way that teh motor back.EMF compansates load variations to adjust the speed.This low cost circuit is popular for low-power and intermittent-use equipment.

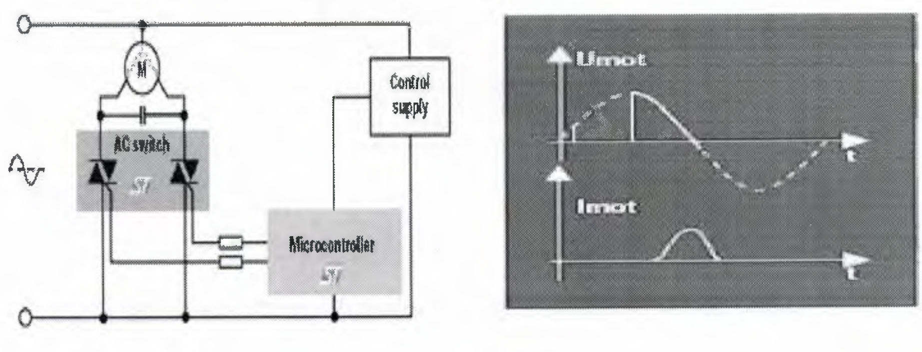
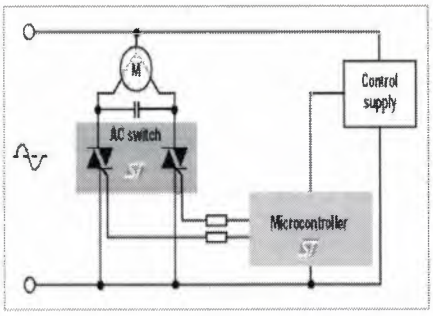


Figure 4.1 DC Universal Motor Drives.

4.3 Bi-Directional Induction Motor Drive

When a motor with a phase-shift capacitor is used, the direction of rotation can be reversed by means of two AC switches which connect the phase-shift capacitor in series with either of the two stator windigs.



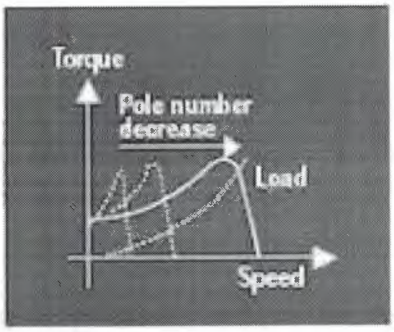


Figure 4.2 Direct-Reverse Speed Induction Motor Drive.

4.4 Multi-Winding On/Off Induction Motor Drive

Here the stator coil is divided into 3 or 4 pairs of windings. The speed is adjusted stepwise by connecting different combinations of these windings to the mains through AC switches in order to change the number of active stator poles and the base speed.

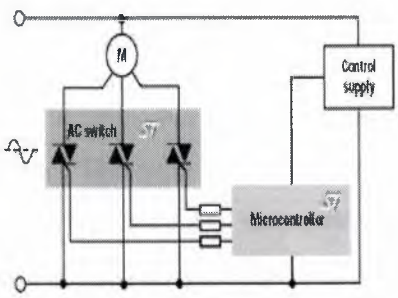


Figure 4.3 Multi-Winding On/Off Induction Motor Drive.

4.5 Three Main Components of an Electric Drive

An electric drive has three main components:

a) The electric motor

b) The power electronic converter

c) The drive controller

4.6 Electric Drive Basic Topology

The following figure shows the basic topology of an electric drive. Beside the three main components, the figure shows an electric power source, a mechanical load, electric and motion sensors, and a user interface.

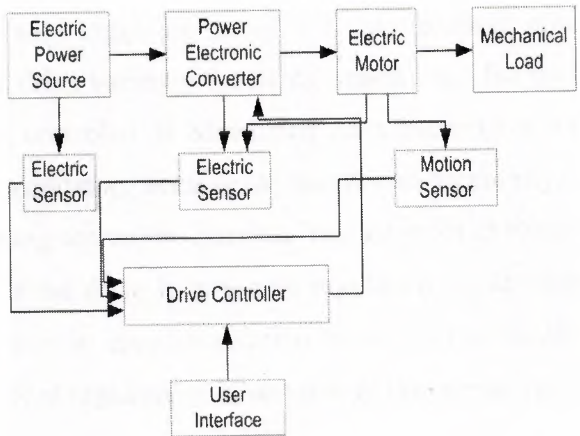


Figure 4.4 Schematic Diagram of an Electric Drive.

The motor used in an electric drive is either a direct current (DC) motor or an alternating current (AC) motor. The type of motor used defines the electric drive's classification into DC motor drives and AC motor drives. The ease of producing a variable DC voltage source for a wide range of speed control made the DC motor drive the favorite electric drive up to the 1960s. Then the advances of power electronics combined with the remarkable evolution of microprocessor-based controls paved the way to the AC motor drive's expansion. In the 1990s, the AC motor drives took over the high-performance variable-speed applications.

The power electronic converter produces variable AC voltage and frequency from the electric power source. There are many types of converters depending on the type of electric drive. The DC motor drives are based on phase-controlled rectifiers (AC-DC converters) or on choppers (DC-DC converters), while the AC motor drives use inverters (DC-AC converters) or cyclo converters (AC-AC converters). The basic component of all the power electronic converters is the electronic switch, which is either semicontrolled (controllable on-state), as in the case of the thyristor, or fully controllable (controllable on-state and off-state), as in the cases of the IGBT (insulated gate bipolar transistor) and the GTO (gate turn off thyristor) blocks. The controllable feature of the electronic switch is what allows the converter to produce the variable AC voltage and frequency.

The purpose of the drive controller is essentially to convert the desired drive torque/speed profile into triggering pulses for the electronic power converter, taking into account various drive variables (currents, speed, etc.) fed back by the sensors. To accomplish this, the controller is based first on a current (or torque) regulator. The current regulator is mandatory because, as mentioned previously, it protects the motor by precisely controlling the motor currents. The set point (SP) of this regulator can be supplied externally if the drive is in torque regulation mode, or internally by a speed regulator if the drive is in speed regulation mode. In the SimPowerSystems Electric Drives library, the speed regulator is in series with the current regulator and is based on a PI controller that has three important features. First, the SP rate of change is limited so that the desired speed ramps gradually to the SP, in order to avoid sudden step changes. Second, the speed regulator output that is the SP for the current regulator is limited by maximum and minimum ceilings. Finally, the integral term is also limited in order to avoid wind-up. The following figure shows a block diagram of a PI controller-based speed controller.

4.6.1 Commutation Failure

Occurs in the half-bridge configuration when a leg is misfired and the conduction period of one device overlaps the conduction period of the other device. This results in a direct short circuit of the DC supply. The prevention is to introduce a dead-time to the switching signals.

4.6.2 Over-Voltage Spike

Occurs when the switching current is too high and induces a voltage spike by the parasitic inductance. This voltage spike could cause the voltage breakdown of the power devices.

4.6.3 Unwanted Shut Down

The protection circuit shuts down the system when it detects an overcurrent. However, when it works in a noisy environment, the harmonic currents cause an unnecessary shut-down.

4.6.4 Thermal Over-Run

Occurs when the drive systems operate in high temperature environment or the cooling systems break down.

In practice, the protection is carried out by internal electronic monitoring with several levels of current as a percentage of the drive's full load current. To prevent overcurrent trips as a result of machine load transients a peak limit system is normally included. For industrial drives, this is set at about 185% of the full load current. This causes the control electronics to turn off the power devices for a reset time.

The overcurrent trip is normally set at 215% of the full load current. If an overcurrent is detected, the system will be shut down power devices until it is reset by the processor. An intelligent gate drive is usually used to detect the on-state voltage of the power devices. If it exceeds a specific level, the gate drive signal will be off for hundreds of μs and reapplied.

4.7 Electronic Control of Direct Current Motors

High-speed, reliable and inexpensive semiconductor devices have produced a dramatic change in the control of dc motors. In this chapter, we examine some of the basic principles of such electronic controls.

In describing the various methods of control, we shall only study the behavior of power circuits.

4.7.1 First Quadrant Speed Control

We begin our study with a variable speed drive for a dc shunt motor. We assume its operation is restricted to quadrant.

A gate triggering processor receives external inputs such as actual speed, actual current, actual torque, etc. These inputs are picked off the power circuit by means of suitable transducers. In addition, the processor can be set for any desired motor speed and torque. The actual values are compared with the desired values, and the processor automatically generates gate pulses to bring them as close together as possible. Limit settings are also incorporated so that the motor never operates beyond acceptable values of current, voltage and speed.

Four features deserve our attention as regards the start-up period: no armature resistors are needed; consequently, there are no losses except those in the armature itself; the power loss in the thyristors is negligible; consequently, all the active power drawn from the ac source is available to drive the load; even if an inexperienced operator tried to start the motor too quickly, the current-limit setting would override the manual command. In effect, the armature current can never exceed the allowable preset value.

The converter absorbs a great deal of reactive power when the motor runs at low speed while developing its rated torque. Furthermore, the reactive power diminishes continually as the motor picks up speed. As a result, power factor correction is difficult to apply during the start-up phase. Two-quadrant control -field reversal .

We cannot always tolerate a situation where a motor simply coasts to a lower speed. To obtain a quicker response, we have to modify the circuit so that the motor acts temporarily as a generator. By controlling the generator output, we can make the speed fall as fast as we please. We often resort to dynamic braking using a resistor. However, the converter can also be made to operate as an inverter, feeding power back into the 3-phase line. Such regenerative braking is preferred because the kinetic energy is not lost. Furthermore, the generator output can be precisely controlled to obtain the desired rate of change in speed.

To make the converter act as an inverter, the polarity of E_d must be reversed as shown in Fig. 4.5. This means we must also reverse the polarity of E_o. Finally, E_d must be adjusted to be slightly less than E_o to obtain the desired braking current I_d (Fig.4.5).

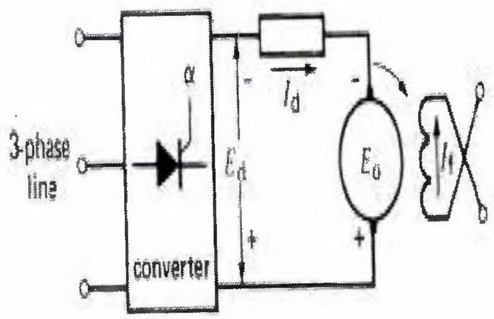


Figure 4.5 Motor Control by Field Reversal.

4.7.2 Two-Quadrant Control-Armature Reversal

In some industrial drives, the long delay associated with field reversal is unacceptable. In such cases, we reverse the armature instead of the field. This requires a high-speed reversing switch designed to carry the full armature current. The control system is arranged so that switching occurs only when the armature current is zero. Although this reduces contact wear and arcing, the switch still has to be fairly large to carry a current, say, of several thousand amperes.

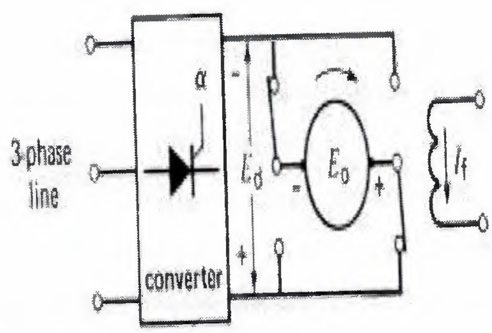


Figure 4.6 Motor Control by Armature Reversal.

4.7.3 Two-Quadrant Control -Two Converters

When speed control has to be even faster, we use two identical converters connected in reverse parallel. Both are connected to the armature, but only one operates at a given time, acting either as a rectifier or inverter. The other converter is on "standby", ready to take over whenever power to the armature has to be reversed. Consequently, there is no need to reverse the armature or field. The time to switch from one converter to the other is typically 10ms. Reliability is considerably improved, and maintenance is reduced. Balanced against these advantages are higher cost and increased complexity of the triggering source.

Because one converter is always ready to take over from the other, the respective converter voltages are close to the existing armature voltage, both in value and polarity. Thus, converter acts as a rectifier, supplying power to the motor at a voltage slightly higher than the cemf E_0. During this period, gate pulses are withheld from converter so that it is inactive. Nevertheless, the control circuit continues to generate pulses having a delay alpha2 so that E_{d2} would be equal to E_{d1} if the pulses were allowed to reach the gates.

4.7.4 Two-Quadrant Control - Two Converters With Circulating Current

Some industrial drives require precise speed and torque control right down to zero speed. This means that the converter voltage may at times be close to zero. Unfortunately, the converter current is discontinuous under these circumstances. In other words, the current in each thyristor no longer flows for 120°. Thus, at low speeds, the torque and speed tend to be erratic, and precise control is difficult to achieve.

To get around this problem, we use two converters that function simultaneously. They are connected back-to-back across the armature. When one functions as a rectifier, the other functions as an inverter, and vice versa. The armature current I is the difference between currents I_{d1} and I_{d2} flowing in the two converters. With this arrangement, the currents in both converters flow for 120°, even when I = 0. Obviously, with two converters continuously in operation, there is no delay at all in switching from one to the other. The armature current can be reversed almost instantaneously; consequently, this represents the most sophisticated control system available. It is also the most expensive. The reason is that when converters operate simultaneously, each must be provided with a large series inductor (L1, L2) to limit the ac circulating currents. Furthermore, the converters must be fed from separate sources, such as the isolated secondary windings of a 3-phase transformer. A typical circuit composed of a delta-connected primary and two wye-connected secondaries. Other transformer circuits are sometimes used to optimize performance, to reduce cost, to enhance reliability or to limit short-circuit currents

.

4.7.5 Two-Quadrant Control With Positive Torque

So far, we have discussed various ways to obtain torque-speed control when the torque reverses. However, many industrial drives involve torques that always act in one direction, even when the speed reverses. Hoists and elevators fall into this category because gravity always acts downwards whether the load moves up or down. Operation is therefore in quadrants 1 and 2.

Consider a hoist driven by a shunt motor having constant field excitation. The armature is connected to the output of a 3-phase, 6-pulse converter. When the load is being raised, the motor absorbs power from the converter. Consequently, the converter acts as a rectifier. The lifting speed depends directly upon converter voltage E_d. The armature current depends upon the weight of the load.

When the weight is being lowered, the motor reverses, which changes the polarity of E_o. However, the descending weight delivers power to the motor, and so it becomes a generator. We can feed the electric power into the ac line by making the converter act as an inverter. The gate pulses are simply delayed by more than 90°, and E_d is adjusted to obtain the desired current flow.

Hoisting and lowering can therefore be done in a stepless manner, and no field or armature reversal is required. However, the empty hook may not descend by itself. The downward motion must then be produced by the motor, which means that either the field or armature has to be reversed.

4.7.6 Four-Quadrant Control

We can readily achieve 4-quadrant control of a dc machine by using a single converter, combined with either field or armature reversal. However, a great deal of switching may be required. Four-quadrant control is possible without field or armature reversal by using two converters operating back-to-back. They may function either alternately or simultaneously, as previously described.

4.7.7 DC Traction

Electric trains and buses have for years been designed to run on direct current, principally because of the special properties of the dc series motor. Many are now being modified to make use of the advantages offered by thyristors. Existing trolley lines still operate on dc and, in most cases, dc series motors are still used. To modify such systems, high-power electronic choppers are installed on board the vehicle. Such choppers can drive motors rated at several hundred horsepower, with outstanding results. To appreciate the improvement that has taken place, let us review some of the features of the older systems.

A train equipped with, say, two dc motors, is started with both motors in series with an external resistor. As the speed picks up, the resistor is shorted out. The motors are then paralleled and connected in series with another resistor. Finally, the last resistor is shorted out, as the train reaches its nominal torque and speed. The switching sequence produces small jolts, which, of course, are repeated during the electric braking process. Although a jolt affects passenger comfort, it also produces slippage on the tracks, with consequent loss of traction. The dc chopper overcomes these problems because it permits smooth and continuous control of torque and speed. We now study some simple chopper circuits used in conjunction with series motors.

Figure 4.7 shows the armature and field of a series motor connected to the output of a chopper. Supply voltage Es is picked off from two overhead trolley wires. The inductor-capacitor combination L1C1 acts as a dc filter, preventing the sharp current pulses Is from reaching the trolley line. The capacitor can readily furnish these high current pulses. The presence of the inductor has a smoothing effect so that current I drawn from the line has a relatively small ripple.

As far as the motor is concerned, the total inductance of the armature and series field is large enough to store and release the energy needed during the chopper cycle. Consequently, no external inductor is required. When the motor starts up, a low chopper frequency is used, typically 50 Hz. The corresponding "on" time Ta is typically 500s. In many systems, Ta is kept constant while the switching frequency varies. The top frequency (about 2000 Hz) is limited by the switching and turn-off time of the thyristors.

Other choppers function at constant frequency, but with a variable "on" time Ta. In still more sophisticated controls, both the frequency and Ta are varied. In such cases, Ta may range from 20s to 800µ. Nevertheless, the basic chopper operation remains the same, Direct-current series motor driven by a chopper. The chopper is not a switch as shown, but a force-commutated SCR. no matter how the on-off switching times are varied.

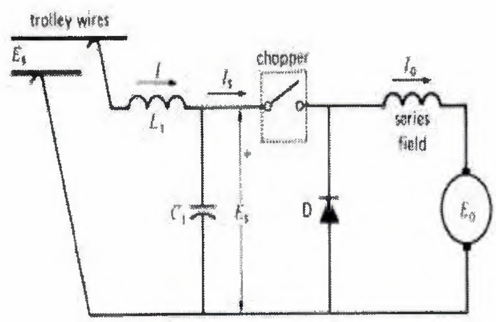


Figure 4.7 Direct-Current Series Motor Driven by a Chopper.

4.7.8 Current-Fed DC Motor

Some electronic drives involve direct current motors that do not look at all like dc machines. The reason is that the usual rotating commutator is replaced by a stationary electronic converter. We now discuss the theory behind these so-called "commutatorless" dc machines.

Consider a 2-pole dc motor having 3 independent armature coils, A, B, and C spaced at 120° to each other (Fig. 4-15). The two ends of each coil are connected to diametrically opposite segments of a 6-segment commutator. Two narrow brushes are connected to a constant-current source that successively feeds current into the coils as the armature rotates. A permanent magnet N, S creates the magnetic field.

With the armature in the position shown, current flows in coil A and the resulting torque causes the armature to turn counterclockwise. As soon as contact is broken with this coil, it is immediately established in the next coil. Consequently, conductors facing the N pole always carry currents that flow into the page, while those facing the S pole carry currents that flow out of the page (towards the reader). The motor torque is therefore continuous and may be expressed by:

$$T = kIB$$

where

T = motor torque (N-m)

I = current in the conductors (A)

B = average flux density surrounding the current-carrying conductors (T)

k = a constant, dependent upon the number of turns per coil, and the size of the armature

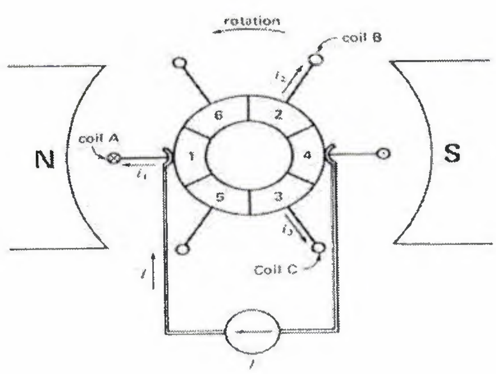


Figure 4.8 Special Current-Fed DC Motor.

If the current and flux density are fixed, the resulting torque is also fixed, independent of motor speed.

The commutator segments are 60° wide; consequently, the current in each coil flows in 60° pulses. Furthermore, the current in the coil reverses every time the coil makes half a turn (Fig.4.9). The alternating nature of the current is of crucial importance. If the current did not alternate, the torque developed by each coil would act first in one, then the opposite direction, as the armature rotates. The net torque would be zero, and so the motor would not develop any power.

Figure 4.9 shows that the ac currents in the 3 coils are out of phase by 120°. Consequently, the armature behaves as if it were excited by a 3-phase source. The only difference is that the current waveshapes are rectangular instead of sinusoidal. Basically, the commutator acts as a mechanical converter, changing the dc current from the dc source into ac current in the coils. The frequency is given by:

$$f = pn/120$$

where p is the number of poles and n the speed (r/min). The frequency in the coils is automatically related to the speed because the faster the machine rotates, the faster the commutator switches from one coil to the next. In effect, the commutator generates a frequency which at all times is appropriate to the instantaneous speed.

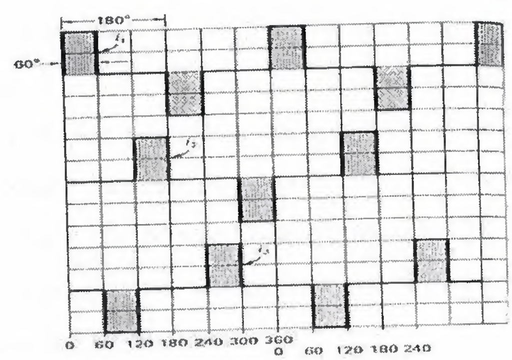


Figure 4.9 The dc current changes to ac current in the coils.

As the coils rotate, they cut across the magnetic field created by the N, S poles. An ac voltage is therefore induced in each coil, and its frequency. Furthermore, the voltages are mutually displaced at 120° owing to the way the coils are mounted on the armature. The induced ac voltages appear as a dc voltage between the brushes. The reason is that the brushes are always in contact with coils that are moving in the same direction through the magnetic field; consequently, the polarity is always the same.

If the brushes were connected to a dc voltage source E, the armature would accelerate until the induced voltage E_o is about equal to E. What determines the speed when the armature is fed from a current source, as it is in our case? The speed will increase until the load torque is equal to the torque developed by the motor. Thus, while the speed of a voltage-fed armature depends upon equilibrium between induced voltage and applied voltage, the speed of a current-fed armature depends upon equilibrium between motor torque and load torque. The torque of a mechanical load always rises with increasing speed. Consequently, for a given motor torque, a state of torque equilibrium is always reached, provided the speed is high enough. Care must be taken so that current-fed motors do not run away when the load torque is removed.

4.7.9 Commutator Replaced by Reversing Switches

Recognizing that each coil in Fig. 4.8 carries an alternating current, we can eliminate the commutator by connecting each coil to a pair of slip rings and bringing the leads out to a set of mechanical reversing switches (Fig. 4.10). Each switch has 4 normally open contacts.

Considering coil A, for example, switch contacts 7 and 8 are closed during the 60° interval when coil side 1 faces the N pole (Fig. 4-11). The contacts are then open for 120° until coil side 4 faces the N pole, whereupon contacts 9 and 10 close for 60°. Consequently, by synchronizing the switch with the position of coil A, we obtain the same result as if we used a commutator.

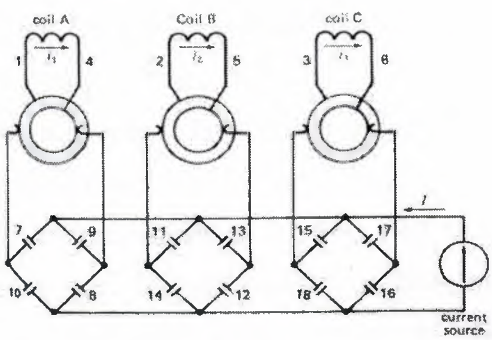


Figure 4.10 The Commutator Can Be Replaced by an Array of Mechanical Switches and a Set of Slip Rings.

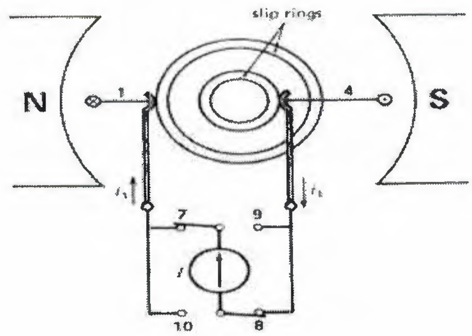


Figure 4.11 Circuit Showing How Current Is Controlled in Coil A.

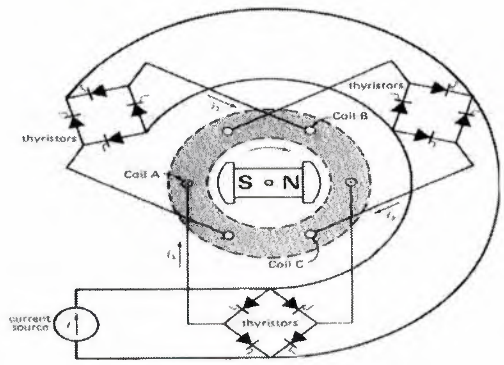


Figure 4.12 The Armature Is Now the Stator, and the Switches Have Been Replaced by Thyristors.

Coils B and C operate the same way, but they are energized at different times. Figure 4.10 shows how the array of 12 contacts and 6 slip rings are connected to the current source. The reversing switches really act as a 3-phase mechanical inverter, changing dc power into ac power. The slip rings merely provide electrical contact between the revolving armature and the stationary switches and power supply.

Clearly, the switching arrangement of Fig. 4.10 is more complex than the original commutator. However, we can simplify matters by making the armature stationary and letting the permanent magnets rotate. By thus literally turning the machine inside out, we can eliminate 6 slip rings. Then, as a final step, we can replace each contact by a thyristor (Fig. 4.12). The 12 thyristors are triggered by gate signals that depend upon the instantaneous position of the revolving rotor.

4.8.5 The Start Voltage Profile

The Start Profile can be a simple single slope from zero voltage to full voltage, or it can be a complex shape to more closely emulate a controlled current start. Like electromechanical starters, open loop soft starters cause the start voltage applied to the motor, to change with time irrespective of the motor and load conditions, eventually getting to full voltage, and under jammed load conditions, developing LRC and LRT until something trips or breaks.

4.8.6 Closed Loop Control

Closed Loop starters monitor an output characteristic or effect from the starting action and dynamically modify the start voltage profile to cause the desired response. The most common closed loop soft starter is the controlled current soft starter where the current drawn by the motor during start is monitored and controlled to give either a constant current, or a current ramp soft start. A much rarer closed loop format is the constant acceleration soft start where the motor speed is monitored by a tachogenerator or shaft encoder and the voltage is controlled to maintain a constant rate of acceleration or a linear increase in motor speed.

The controlled current soft starters are available with varying levels of sophistication. In the most basic systems, the soft starter is essentially a standard TVR soft starter with a ramp freeze option where the current on one phase is monitored and compared to a set point. If the current exceeds the set point, the ramp is frozen until the current drops below that set point. At the other end of the scale, a comprehensive closed loop soft starter will monitor the current on all three phases and dynamically change the output voltage to correct the start current to the required profile. This system is able to both increase and reduce the start voltage to suit the application.

A constant current starter will start initially at zero volts and rapidly increase the output voltage until the required current is delivered to the motor, and then adjust the output voltage while the motor is starting until either full voltage is reached, or the motor overload protection operates. Constant current starters are ideal for high inertia loads, or loads where the starting torque requirements do not alter.

The dc motor in Fig. 4.12 looks so different from the one in Fig. 4.8 that we would never suspect they have the same properties. And yet they do.

4.7.10 Synchronous Motor as a Commutatorless DC Machine

The revolving-field motor in Fig. 4.12 is built like a 3-phase synchronous motor. However, because of the way it receives its ac power, it behaves like a "commutatorless" dc machine. This has a profound effect upon its performance.

First, the "synchronous motor" can never pull out of step because the stator frequency is not fixed, but changes automatically with speed. The reason is that the gates of the SCRs are triggered by a signal that depends upon the instantaneous position of the rotor. For the same reason, the machine has no tendency to oscillate or hunt under sudden load changes.

Second, the phase angle between the ac current in a winding and the ac voltage across it can be modified by altering the timing of the gate pulses. This enables the synchronous motor to operate at leading, lagging, or unity power factor.

Third, because the phase angle between the respective voltages and currents can be fully controlled, the machine can even function as a generator, feeding power back to the dc current source. The thyristor bridges then operate as rectifiers.

Currents i1, i2, i3 in Fig. 4-12 flow only during 60 degree intervals, as they did in the original dc machine. In practice, the conduction period can be doubled to 120°, by connecting the coils in wye and exciting them by a 3-phase, 6-pulse converter (Fig. 4.13). This reduces the number of thyristors by half. Furthermore, it improves the current-carrying capacity of the windings because the duration of current flow is doubled.

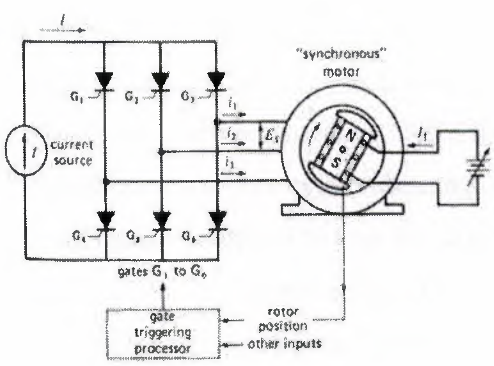


Figure 4.13 Commutatorless dc motor being driven by a converter.

series with the line voltage applied to the motor, or can be connected inside the delta loop of a delta connected motor, controlling the voltage applied to each winding.

4.8.1 Voltage Control

Voltage control is achieved by means of solid state A.C. switches in series with each phase. These switches comprise either:

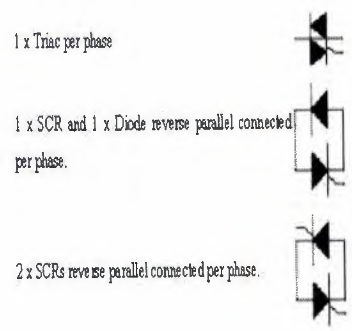


Figure 4.17 Voltage Control Switches.

4.8.2 Solid State Switches

These Solid State Switches are phase controlled in a similar manner to a light dimmer, in that they are turned on for a part of each cycle. The average voltage is controlled by varying the conduction angle of the switches. Increasing the conduction angle will increase the average output voltage. Controlling the average output voltage by means of solid state switches has a number of advantages, one of the major advantages being the vast improvement in efficiency relative to the primary resistance starter, due to the low on state voltage of the solid state switches. Typically, the power dissipation in the starter, during start, will be less than 1% of the power dissipated in a primary resistance starter during start. Another major advantage of the solid state starter is that the average voltage can be easily altered to suit the required starting conditions. By variation of the conduction angle, the output voltage can be increased or reduced, and this can be achieved automatically by the control electronics. The control electronics can be preprogrammed to provide a particular output voltage contour based on a timed sequence (open loop), or can dynamically control the output voltage to achieve an output profile based on measurements made of such characteristics as current and speed (closed loop).

4.8.3 Switching Elements

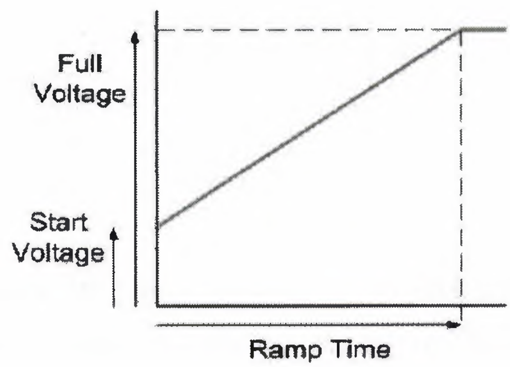


Figure 4.18 Switching Element Voltage-Time Graph.

The switching elements must be able to control the current applied to the motor at line voltage. In order to maintain a high level of reliability on a real industrial type supply, the switching elements need to be rated at least 3 times the line voltage. On a 400 volt supply, this means that the requirement is for 1200 Volt devices, and 600 Volt devices on a 200 volt supply. It is also important that the switching elements have a good transient current overload capacity.

1200 Volt triacs with good current transient overload characteristics are not readily available, and so the choice is really between the SCR-Diode and SCR-SCR. There are some triacs which are suitable for this operation, but they are not easily attainable.

The major differences between the SCR-SCR and the SCR-Diode options are price, and the harmonic content of the output voltage. The SCR-SCR method provides a symmetrical output which is technically desirable from the point of supply disturbances and harmonics, while the SCR-Diode method is inferior technically, it is commercially more effective and easier to implement.

Harmonics awareness and paranoia has drastically reduced the number of SCR-Diode type soft starters on today's market, but they do still exist. The technology is not flux depends upon the stator currents and the exciting current If. The flux is usually kept fixed; consequently, the induced voltage Es is proportional to the motor speed.

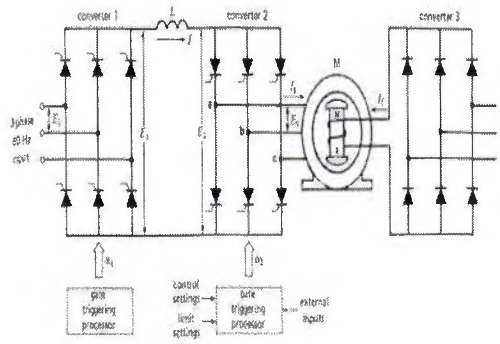


Figure 4.15 Commutatorless DC Motor Driven by a Converter with a DC Link. The Output Frequency can be Considerably Greater than 60 Hz, Thus Permitting High Speeds.

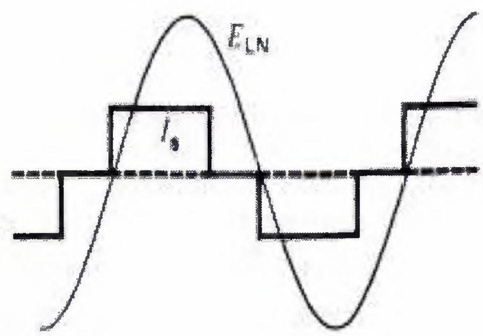


Figure 4.16 Typical Voltage and Current Waveshapes in Fig 4.15.

4.8 Soft Starters for Induction Motors

A soft starter is another form of reduced voltage starter for A.C. induction motors. The soft starter is similar to a primary resistance or primary reactance starter in that it is in series with the supply to the motor. The current into the starter equals the current out. The soft starter employs solid state devices to control the current flow and therefore the voltage applied to the motor. In theory, soft starters can be connected in

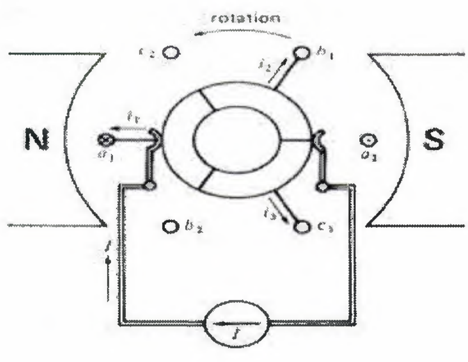


Figure 4.14 This elementary dc motor is equivalent to the entire circuit of Fig.4.13.

4.7.11 Standard Synchronous Motor and Commutatorless DC Machine

The machine shown in Fig. 4.13 can be made to function as a conventional synchronous motor by applying a fixed frequency to the SCR gates. Under these conditions, the input to the gate triggering processor no longer depends on rotor position or rotor speed.

4.7.12 Synchronous Motor Drive Using Current-Fed DC Link

Figure 4.15 shows a typical commutatorless dc motor circuit. It consists of two converters connected between a 3-phase source and the "synchronous" motor. Converter 1 acts as a controlled rectifier, feeding dc power to converter 2. The latter behaves as a naturally-commutated inverter whose ac voltage and frequency are established by the motor.

Readers familiar with feedback theory will recognize that the basic distinction between the two machines is that one functions on open loop while the other operates on closed loop.

A smoothing inductor L maintains a ripple-free current in the so-called dc link between the two converters. Current I is controlled by converter 1, which acts as a current source. A smaller bridge rectifier (converter 3) supplies the field excitation for the rotor.

Converter 2 is naturally-commutated by voltage Es induced across the terminals of the motor. This voltage is created by the revolving magnetic flux in the air gap. The always easily recognizable as such with terms such as three pulse technology being used to describe SCR-Diode systems as opposed to six pulse technology describing SCR-SCR systems.

4.8.4 Open Loop Control

Open Loop soft starters are soft starters producing a start voltage profile which is independent of the current drawn, or the speed of the motor. The start voltage profileis programmed to follow a predetermined contour against time. A very basic Timed Voltage Ramp (TVR) system operates by applying an initial voltage to the motor, and causing this voltage to slowly ramp up to full voltage. On basic systems, the initial start voltage is not adjustable, but the ramp time is. Commonly the voltage ramps time is referred to as the acceleration ramp time and is calibrated in seconds. This is not an accurate description as it does not directly control the acceleration of the motor. A lightly loaded motor can accelerate to full speed even with a sixty second ramp selected. More correctly this should be referred to as the voltage ramp time. On more comprehensive units, the start voltage is pre-setable, typically from 10% to 70% of full line voltage. This should be set to achieve at least breakaway torque for the motor at start. There is little advantage in the motor sitting, staining to start due to insufficient torque. this will only increase the heat dissipated in the motor. The start voltage setting is often referred to as the start torque setting and calibrated in percent. This is a nonsense, as although increasing the start voltage is going to increase the starting torque of the connected motor, the actual starting torque is a function of both the start voltage and the motor design. The starter does not know anything about the connected motor, and so is not able to deliver a prescribed amount of torque under open loop conditions. The actual start torque produced is initially equal to the LRT multiplied by the square of: (the start voltage divided by the line voltage). The LRT of the motor could vary from as low as 60% FLT to as high as 350% FLT which is a range of almost 6 to 1.

The current ramp soft starter operates in the same manner as the constant current soft starter except that the current is ramped from an initial start current to a current limit setting over a period of time. The initial start current, current limit, and the ramp time are all user adjustable settings and should be customize to suit the application. The current ramp soft starter can be used for a number of advantages over constant current in some applications. Machines which have a varying start torque requirement, such as on load conveyers, or applications requiring a reduced initial torque such as pumping applications, or genset applications where the relatively slow application of current load will allow the genset to track the load are examples of situation where the current ramp soft start can be used to advantage.

4.8.7 Starting Torque

To start a machine, the motor must develop sufficient torque over the entire speed range to exceed the work and loss torque of the driven load, and provide a surplus torque for accelerating the machine to full speed. The starting torque delivered by the motor at any speed, is equal to the full voltage starting torque at that speed, multiplied by the current or voltage reduction squared. Provided the full voltage speed/torque curves and the full voltage speed/current curves are available, the reduced voltage (or current) speed/torque curves can be calculated. This curve can be superimposed onto the load speed torque curve, and provided the torque developed at all speeds exceeds the load torque, the motor will accelerate to full speed. If the curves cross, the start current (or voltage) will need to be increased to increase the start torque developed by the motor. The difference between the torque developed and the load torque is essentially the acceleration torque that will accelerate the machine to full speed. A high acceleration torque may be desirable for a high inertia machine in order to minimize the starting time.

With a controlled current soft starter, the voltage reduction reduces as the motor impedance accelerates due to the rising motor impedance. As the motor approaches full speed, the voltage rises quickly (against speed) to full voltage. When the torque curve for a motor started by a constant current starter is compared with that of a constant voltage starter such as an auto transformer starter, it can be seen that there is an increase in the torque as the motor accelerates with a constant current start. This is ideal because as the motor and machine increase in speed, the actual load on the motor shaft will increase also. This characteristic will often enable a load to be started with a lower current on a soft starter than traditional starter methods.

4.8.8 Slip Ring Motors

Soft starters can be applied to many slip ring motors, however there are some where the application of a soft starter will not give satisfactory results. Slip ring motors are often employed for their ability to produce a very high torque across the entire speed range. The slip ring motor is able to do this at a very low start current. Another reason for the application of a slip ring motor is that it is able to offer a high degree of control.

If the slip ring motor is employed to give a very high start torque across the entire speed range, then the soft starter is not going to provide a satisfactory solution. This is because the application of a soft starter or any other primary starter, is going to reduce the torque available. Where the requirement is for a gentle start at reduced torque, the soft starter is of benefit.

A common misconception is that the slip ring starter can be converted to a cage type motor by shorting the slip rings and starting by the normal methods. If the secondary winding is shorted, the slip ring motor will exhibit a very high LRC (typically >1000%) and a very low LRT (typically < 100%). If a reduced voltage starter is applied under these conditions, the start torque will be very low and will not start a machine. To apply a reduced voltage starter to a slip ring motor, first ascertain that a reduced torque is going to start the machine, then fit resistors to the rotor circuit which will give curves similar to a high start torque cage motor. These resistors must then be bridged once the machine has reached full speed. The value of the resistance is dependent on the motor and the curve required, however the resistors must absorb a lot of energy, dependent on the inertia of the load. It is common to use the final stage resistance of the existing starter when available.

4.8.9 Ratings

As the rating of the starter is essentially thermal, there is a strong relationship between the start time, start current, start frequency, ambient temperature, OFF time, and the rating of the starter. Typically, there thermal inertia of the SCR Heatsink assembly is quite long so there is not a large variation in the rating between say a 10 second rating and a 30 second rating. - Semiconductor fuse curves do not follow the ratings curves for soft starters and only offer Short Circuit protection.

4.9 Summary

In this chapter, motor drives in AC and DC, their specific definition and control circuit, soft starters are given by their characteristics.

CHAPTER FIVE

POWER SIMULATION RESULTS

5.1 Overview

In this chapter, power simulation analysis with their circuit and PSIM simulation graphs will be given.

5.2 Thyristor System Circuit and Simulation

Development in the semiconductor technology replaced the old control systems (based on special hybrid transformers) with powerfull semiconductor (thyristor etc.) technique.Different types of thyristor controlled rectifiers are present in technical literature and it is possible to choose any suitable system depending upon the application: High efficiency and no inertia is necessary, good overloading ability, higher reliability, very small sized designed system.

In thyristor system we can obtain some important results for machine drives.Rectifier system is important to study all problems during design and study of different semiconductor system for automatic control of syncronous machine to be able to get full pictures of voltages and currents that may appear at any circumstances at any regime for very low to overload on the shaft.

Thyristor system has advantages on the motor drives. In this system, we can control the speed, torque, voltage and current on the motor at very extensively time duration. The efficieny is very high and usefull for motor drives.

5.3 Psim Modelling Analysis and Results

The thyristor system is analyzed by PSIM modelling software and the graphs in figures.The simulations were performed with PSIM software.

The diagram of proposed design is given in figure 5.1.

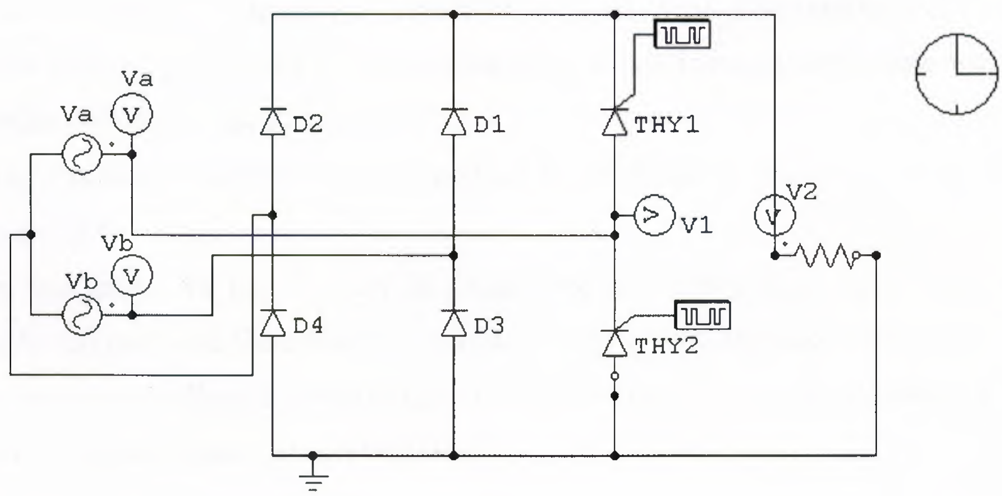


Figure 5.1 Diagram of Thyristor System.

This circuit consists of two sinusoidal source like a phase, which have voltage peak amplitude 110 volts.Their frequecy values are 60 hz.It consists of four diodes bridge that is connected between phase B and neutral but two additional thyristors parallel to bridge are connected to phase A from the common point.

The designed diagram works as the follows:

At the first, it looks obvious that when thyristors THY1 and THY2 are fully open when the phase angle equal to zero bridge of D3, D1, THY1, THY2 works (under line voltage) .

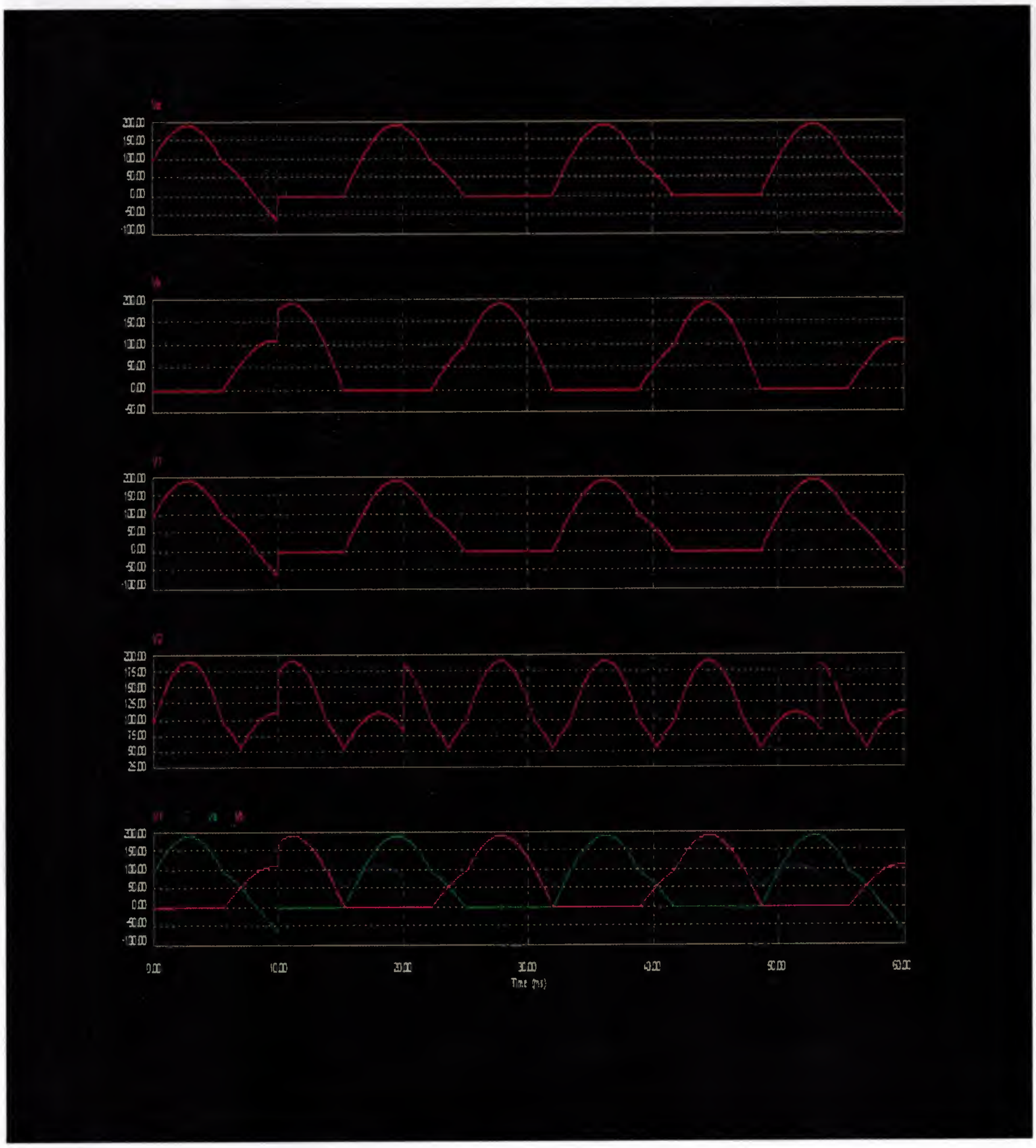
Up to some level of the torque, the exciting coil of the synchronous machine gets power from only one phase B bridge. As the torque and the current of the machine rise. Internal stator current sensor starts to form a signal to switch on thyristors THY1 and THY2. This signal depends on the amplitude of the synchronous current. It changes the phase angle of the thyristors and in this way it controls the voltage value on the exciting coil proportional to the synchronous machine current.

5.4 Simulations

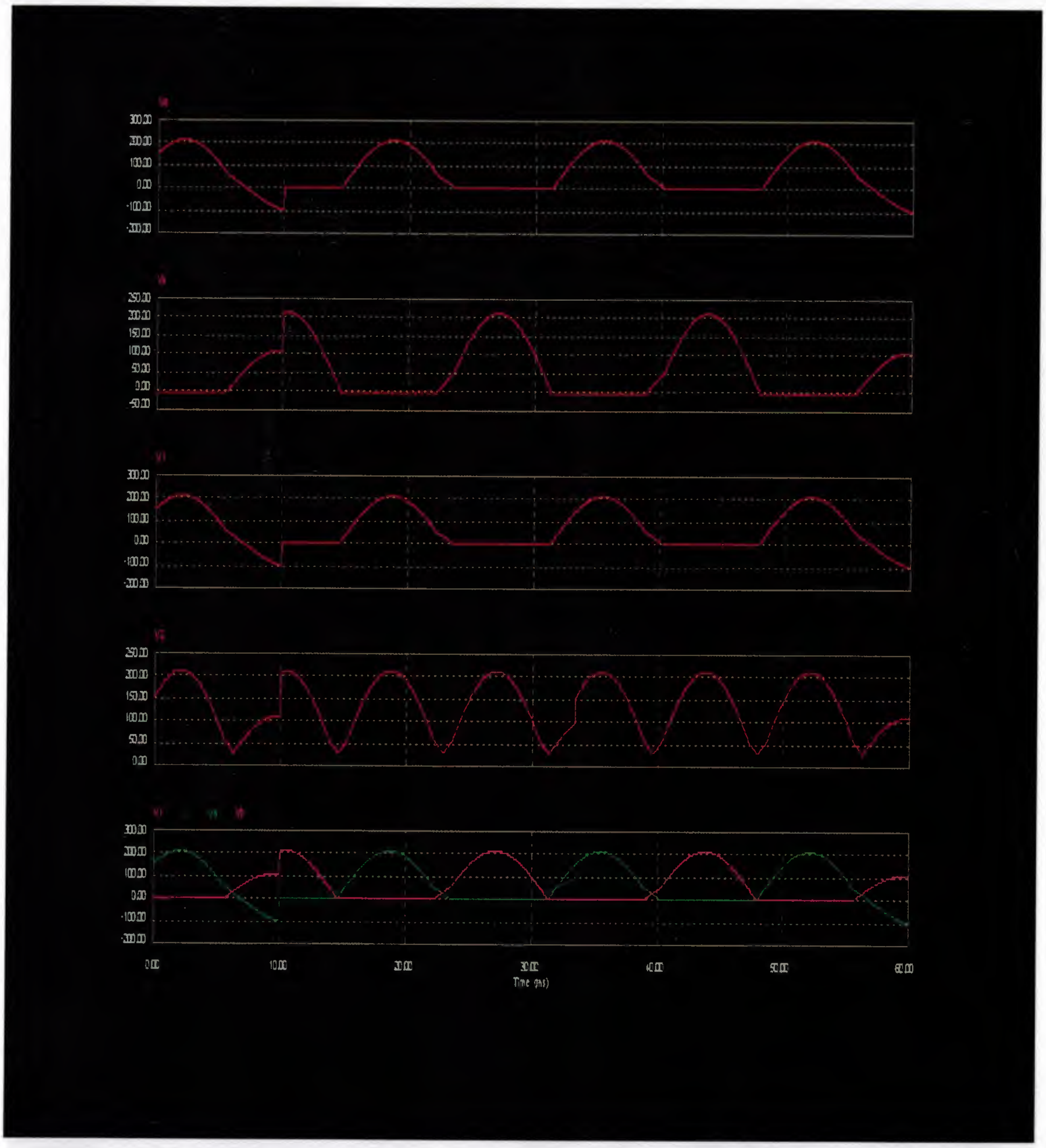
The simulations were performed with POWER SIMULATION software. A selected DC load in simulation is a resistor which have 100 Ω value. When the phase angle changes, the switching time changes as well; it produces together with current reference oscillation small non-sinusoidal periods to the load steps. The oscillation of the DC current reference is caused by speed controller, which is tuned concerning mainly on transient situation of speed reference.

The simulation control device is used for the limitation of the time, and it controls the time step.

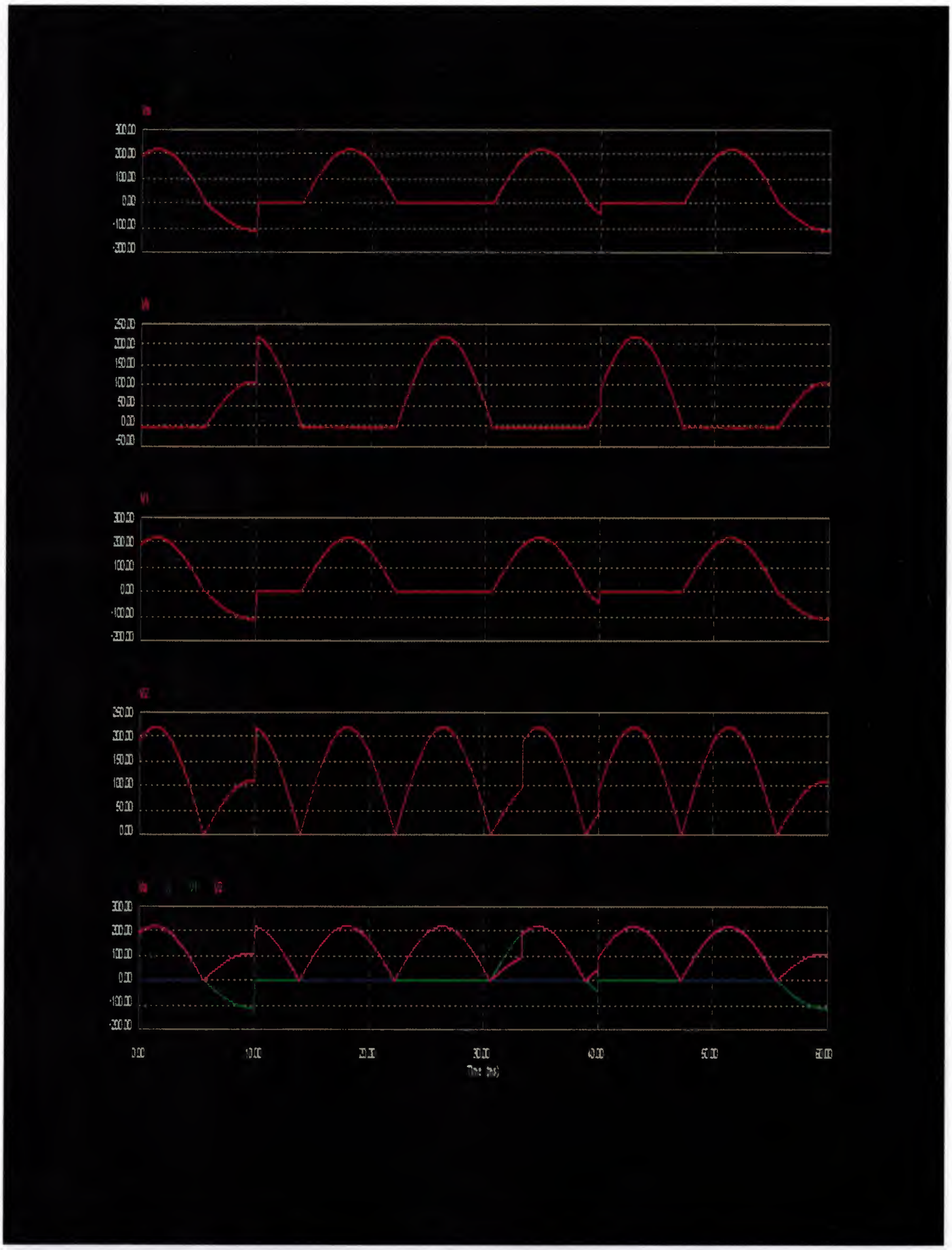
In first graph Va has 0^{o} phase angle and V_b has -120^{o} phase angle. Time step is 0.00001 minutes and total time is 1 minute.Rectifiers is in the nonsensitive zone as in this zone phase voltage is always higher than line voltage even if the thyristors are fired current prefers the path through D1, D4.



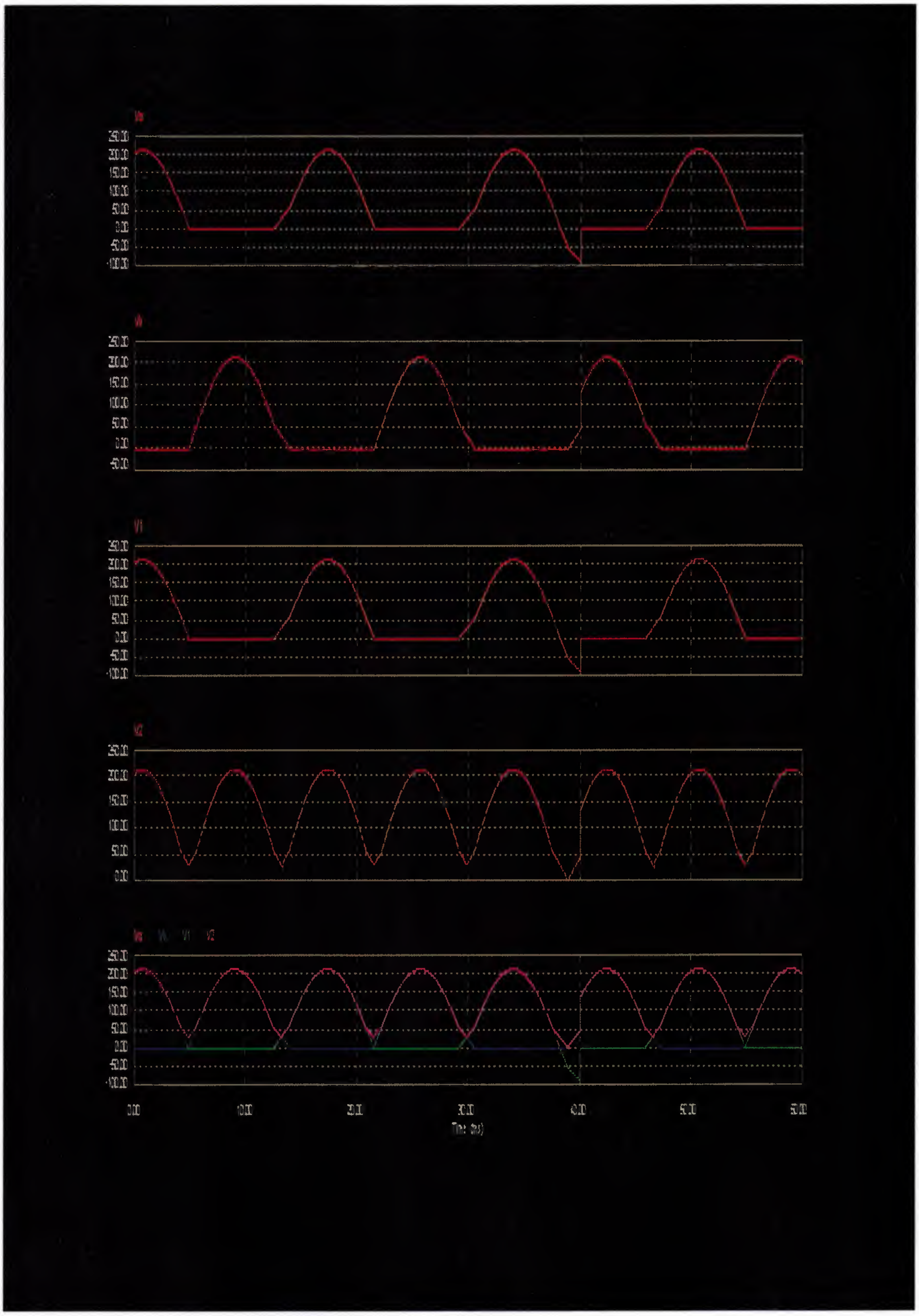
In this graph Va has 30^{0} phase angle. From the point where voltage V_b crosses line voltage. The thyristors operate and the current passes through the path THY1 and D3. The thyristor THY2 does not operate as negative voltage is applied to it for this duration.



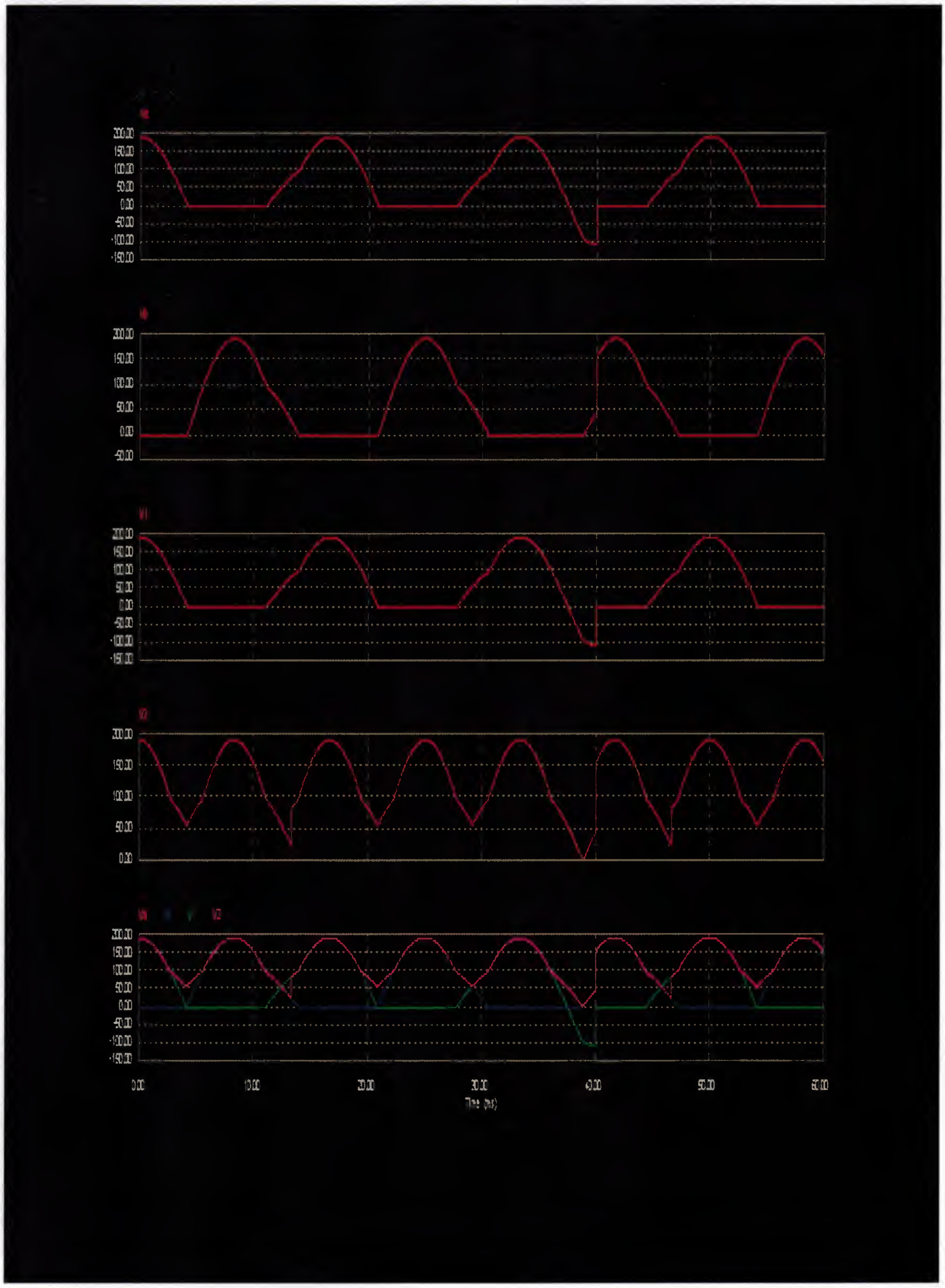
In this graph V_a has 60^o phase angle.The thyristors still operate and the current passes through the path THY2 and D3.



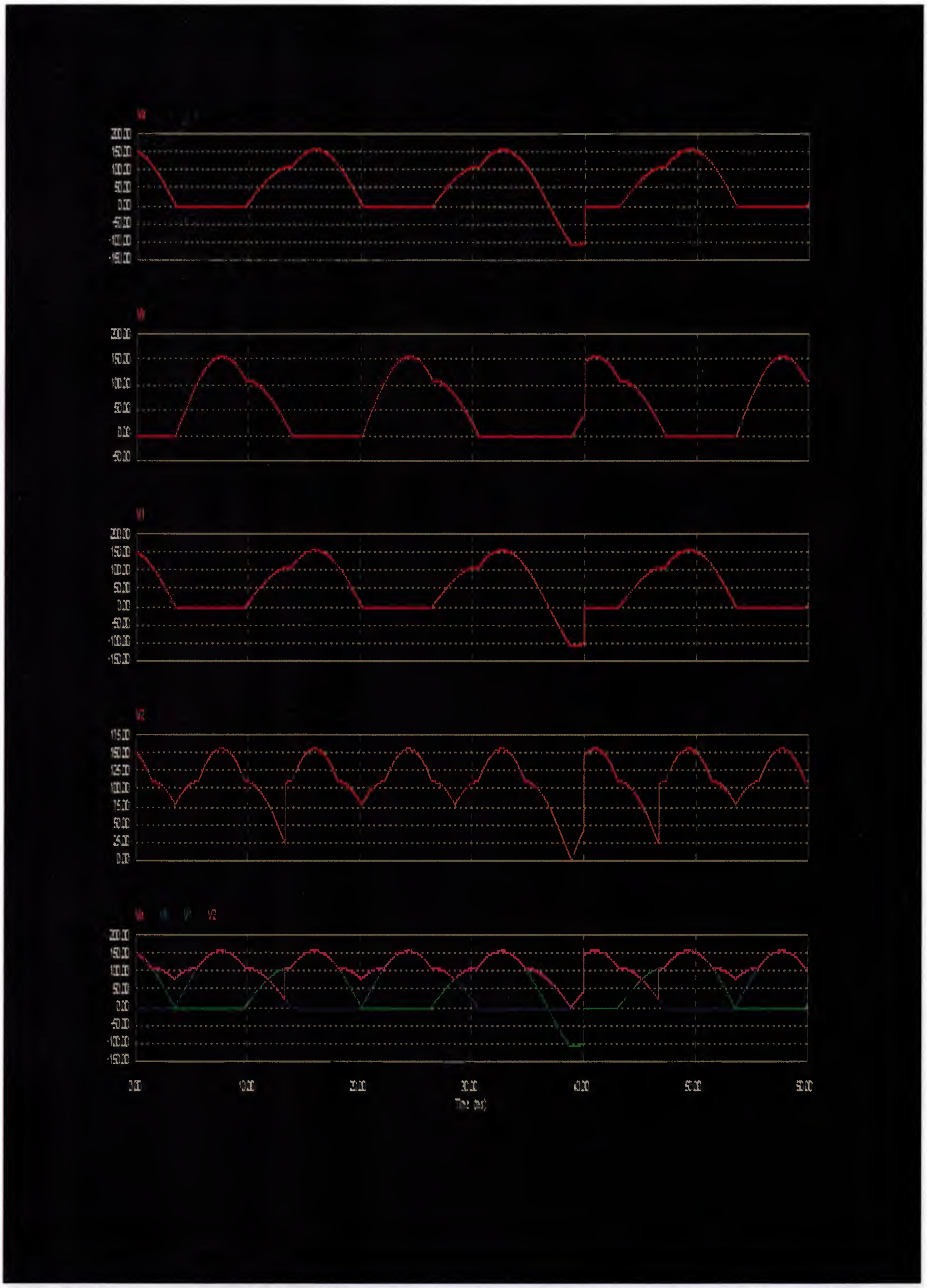
For 90° phase angle, THY2 and D2 starts to operate together from where voltage V_b crosses voltage V_a.



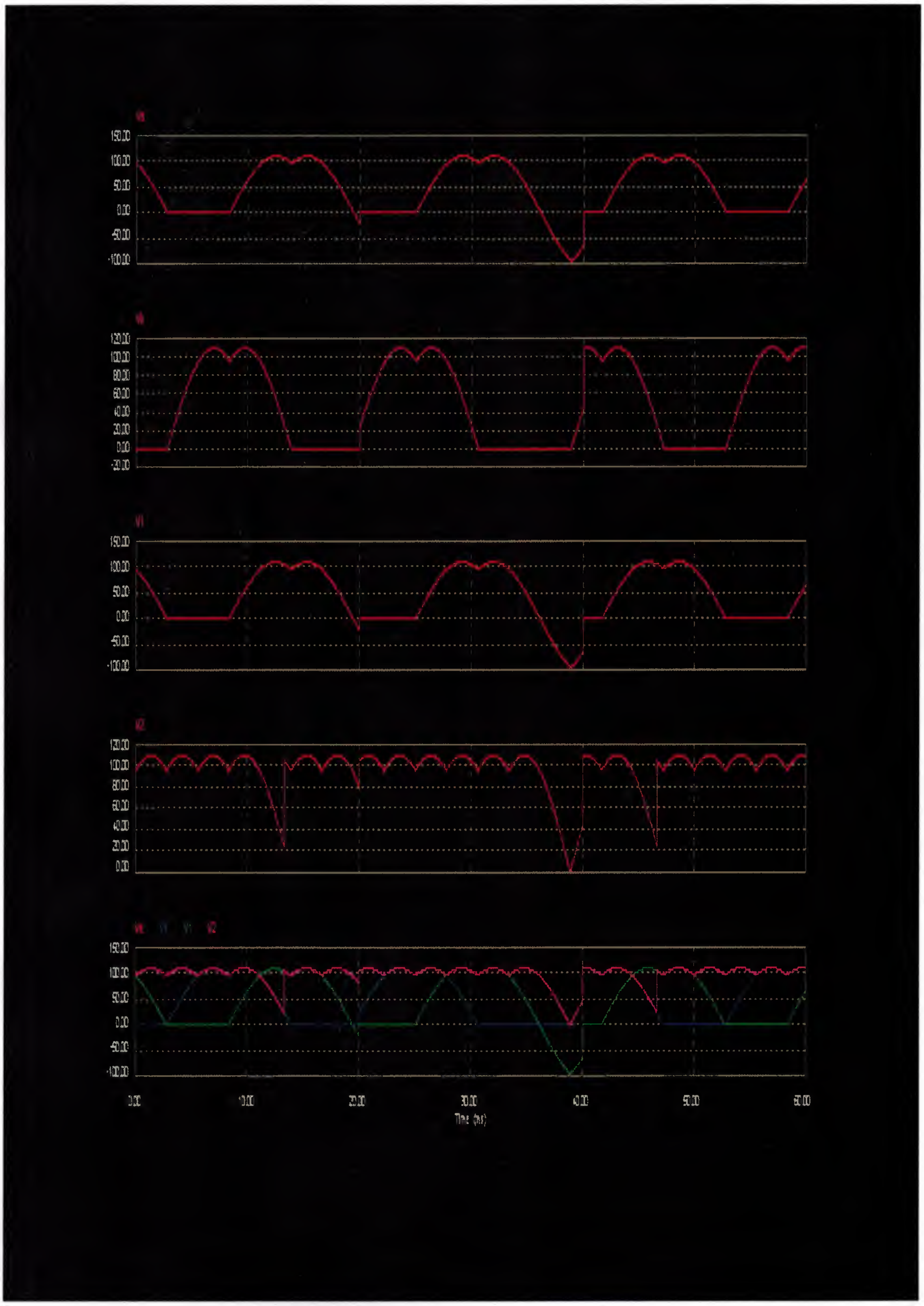
For 120^{o} phase angle, THY2 and D2 operate together where voltage V_{ab} crosses voltage V_a.



For 150° phase angle, both phase voltages are operating.



For 180° phase angle, both thyristors are blocked and only phase voltage V_b is present.



5.5 Summary

The designed and analyzed converter is simple to construct, cheap, and low distortion converter. It is suitable for smooth up to % 15 exciting voltage control. The advantage over two phase control rectifier is the possibility of better zone control of synchronous motors and generators.

The proposed low distortion rectifier can be implemented in common electrical applications such as: chemical electrolysis in the industry, DC motors, battery chargers etc.

This system helps to raise the stability of the synchronous machine, specially wind and subsea generating machine where naturally torque on the axe is unstable and various from time to time even abruptly.

For this system quick instantaneous rise (or fall down) excitation current can help to keep dynamic stability of the synchronous machine.

CONCLUSION

The diode-clamped converter has become the most used and analyzed multilevel topology for the last decades. Nevertheless, there are still some aspects that require further insight. And this type converter is very usefull for motor controlling in the real-life applications. Because of these reasons, this project has focused on the analysis and simulations of the converter.

Power electronic devices and the analysis of these devices with their properties and applications areas were discussed.

In chapter 1, definition of power electronics, application areas of power electronics and introduction to the power electronic devices were discussed.

In chapter 2, power electronic devices, their types, structure, characteristics, protection ways, and schematic and circuit diagrams with their operation were presented.

In chapter 3, definition of most popular power electronic converters, their circuit diagram, their operation in the circuit, classifications, structural analysis, their waveforms and their maximum efficiency conditions were developped.

In chapter 4, motor drives, motor drives properties, components of motor drives, their failures, speed control, voltage control, current control, switching conditions, starters for motors, and switching elements were explained.

In chapter 5, power simulation results using different variables such as voltage, current, time interval different results on different phase angle in the converter and PSIM modelling results with their PSIM simulation results were discussed.

Using PSIM program as a tool, comparison criteria is created. Analysis and the results are compared with each other giving decision upon the design method.

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