UNIVERSITY OF NAIROBI

PRJ-076
DESIGN OF INVERTER DRIVE FOR AC INDUCTION MOTOR.

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I am grateful to my parents, and family who have been a good source of inspiration and support.

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Finally, great thanks to the Almighty Father for having seen me through it all and enabling me to accomplish this task.

Dedication

To my parents: Your love, support, patience and encouragement, gave me the will and determination to complete this work.

Abstract

The emerging applications in home appliances, HVAC drives, professional hand tools, automobile accessories, fans, pumps, and process drives can be attributed to demand for variable speed drives. Almost all of these applications do not require precise positioning or speed control and hence of low performance types. They are very cost sensitive as the make or break decision about their introduction is primarily decided by the cost only, in spite of their advantages in terms of energy savings and control and operational flexibilities. The inverter Controlled induction motor drive is a strong competitor for such applications due to the low cost of the motor and due to its inherent features such as the simplicity of control requiring no sensors in these low performance applications, brushless as it has no slip rings thus making it truly almost maintenance-free, robust and proven construction and operation. But most of the variable speed induction motor drive's cost comes from the inverter and its controller.
A principle of v/f control of the output voltage in PWM voltage-source inverters is described.

The proposed approach is a modification of the sinusoidal modulation technique, and consists of open-loop control of three-phase inverter fed induction motor that is complementary commutated, with a constant dc link converter, based on uncontrolled rectifiers.

This project addresses issues of design and development of a three-phase inverter to drive an induction machine from a 200V DC bus. The various issues addressed in detail are: Specifications, appropriate control strategy and drive configuration, design of control circuits, , tuning and ease of manufacturability.

Theoretical considerations, computer simulations, and experimental results are presented to correlate the key design aspects and design specifications.

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CHAPTER ONE

INTRODUCTION

1.1 GENERAL INTRODUCTION

Induction motors are the most widely used motors for appliances, industrial control, and automation; hence, they are often called the workhorse of the motion industry. They are robust, reliable, and durable. When power is supplied to an induction motor at the recommended specifications, it runs at its rated speed. However, many applications need variable speed operations. For example, a washing machine may use different speeds for each wash cycle. Historically, mechanical gear systems were used to obtain variable speed.

Recently, electronic power and control systems have matured to allow these components to be used for motor control in place of mechanical gears. These electronics not only control the motor's speed, but can improve the motor's dynamic and steady state characteristics. In addition, electronics can reduce the system's average power consumption and noise generation of the motor.

Induction motor control is complex due to its nonlinear characteristics. While there are different methods for control, Variable Voltage Variable Frequency (VVVF) or V/*f* is the most common method of speed control in open loop. This method is most suitable for applications without position control requirements or the need for high accuracy of speed control. Examples of these applications include heating, air conditioning, fans and blowers. *V/f control can be implemented by using low cost PIC micro*(*microcontrollers*, rather than using costly digital signal processors (DSPs) .

In this project open loop control of motor is simulated and then implemented using timers (LM 555), and forms the control circuit from which a Pulse Width Modulated (PWM) inverter is driven. The PWM inverter will then supply the induction motor with the correct voltage, frequency and phase.

The induction motor that is to be controlled is a squirrel cage induction motor, which produces 2.2 kW.

The squirrel cage motor has a rotor with a winding consisting of conducting bars embedded in slots in the rotor iron and short-circuited at each end by conducting end rings.

An inverter converts dc voltage from the input to ac voltage at the output. The PWM inverter output ac voltage can be controlled in both magnitude and frequency. This control of voltage and frequency is needed as it allows the user to vary the current, torque and speed of the induction motor at various loads.

The complete system consists of an ac voltage input that is put through a diode bridge rectifier to produce a dc output which across a shunt capacitor, will in turn, feed the PWM inverter. The PWM inverter is controlled to produce a desired sinusoidal voltage at a particular frequency, which is filtered by the use of an inductor in series and capacitor in parallel and then through to the squirrel cage induction motor. The voltage and frequency that the inverter supplies, is controlled by the control system which takes its input from the induction motor parameters to produce required speed. The system diagram is shown in Figure 1.0

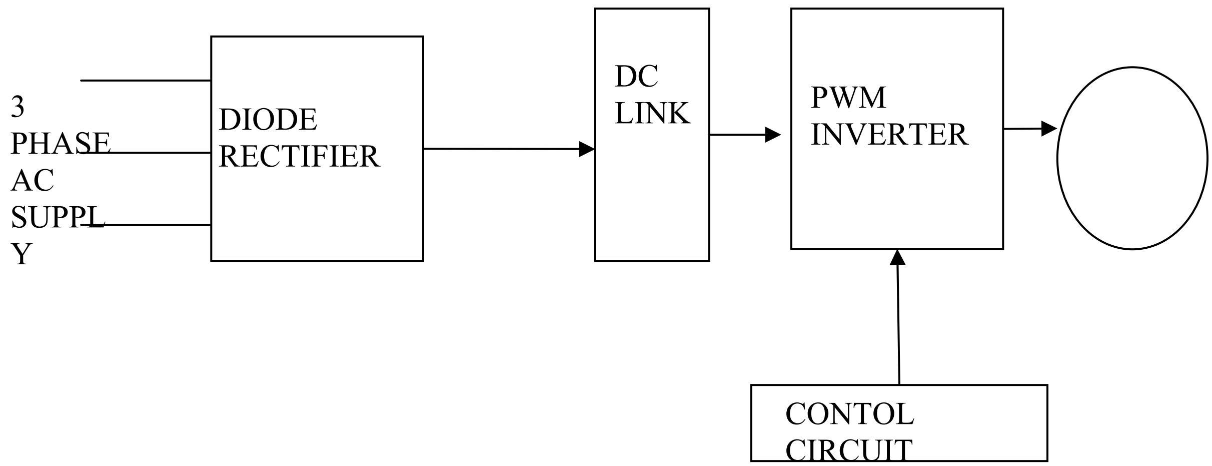


Figure 1.0 main components of ac variable speed system.

This project utilized electronics to measure the line currents and motor speed, and a combination of timers, gates and commutation techniques to control the switching within the PWM inverter so that the appropriate voltage and frequency is applied to the induction motor. In order to achieve this, a good understanding of PWM inverter characteristics and control theory along with solid understanding of squirrel cage induction motor function and parameters, was needed before commencement of the design process.
In order to simulate the circuits and to validate the design process **MULTISIM** simulation software was used. MULTISIM is a design software that allows the design and simulation of electronic systems and electronics components. It allows the viewing of output graphs of any features in the system including internal component parameters. This software was used to simulate the induction motor and its characteristics under different conditions as well as simulation of the PWM circuit. This report is therefore divided into six sections as listed below,

- Chapter One: Introduction
- Chapter Two: Literature review
- Chapter Three ac drives
- Chapter Four: Design
- Chapter Five: Results, analysis and discussion
- Chapter Six: Conclusion and recommendations

1.2 OBJECTIVES

This project aimed at:
• Gaining understanding of Squirrel cage Induction motor characteristics and parameters.
• Gaining Understanding of Pulse Width Modulated (PWM) Inverter.
• Understanding control techniques of a PWM fed induction motor, in particular v/f control.
• Simulation of PWM inverter and induction motor
• Implement control of induction motor through the use of an inverter drive.
• Implementation of the inverter drive and connection of control circuitry to PWM inverter and induction motor.
• Testing of system.

CHAPTER TWO

LITERATURE REVIEW

2.1 Introduction

AC motors are of great use in industry due to their low cost, robustness and precise controllability, however, only over the past few years that the full potential of the controllability of these motors has been reached. This is due to the development of more powerful microprocessors that can compute long algorithms much faster, which has led to the full control of an induction motor using PWM inverters.

2.2.0 Three-Phase Induction Motor

2.2.1 Introduction

3-phase AC induction motors can be operated either directly from the mains or from adjustable frequency drives. The applications for these motors cover almost every stage of manufacturing and processing. Applications also extend to commercial buildings and the domestic environment. They are used to drive pumps, fans, compressors, mixers, agitators, mills, conveyors, crushers, machine tools and cranes. It is increasingly common practice to use 3-phase squirrel cage AC induction motors with *variable voltage variable frequency (VVVF) converters* for variable speed drive (VSD) applications.
The reliability of squirrel cage AC induction motors, compared to DC motors, is high. The only parts of the squirrel cage motor that can wear are the bearings.

2.2.2 Basic construction

The AC induction motor comprises 2 electromagnetic parts:
• Stationary part called the *stator*
• Rotating part called the *rotor*, supported at each end on bearings

The stator and the rotor are each made up of:
• An *electric circuit*, usually made of insulated copper or aluminum, to carry current
• A *magnetic circuit*, usually made from laminated steel, to carry magnetic flux

i) The stator
The *stator* is the outer stationary part of the motor, which consists of:
• The **outer cylindrical frame** of the motor, which is made either of welded sheet steel, cast iron or cast aluminum alloy. This may include a flange for mounting.
• The **magnetic path**, which comprises a set of slotted steel laminations pressed into the cylindrical space inside the outer frame. The magnetic path is laminated to reduce eddy currents, lower losses and lower heating.

• A set of **insulated electrical windings**, which are placed inside the slots of the laminated magnetic path. The cross-sectional area of these windings must be large enough for the power rating of the motor. For a 3-phase motor, 3 sets of windings are required, one for each phase.

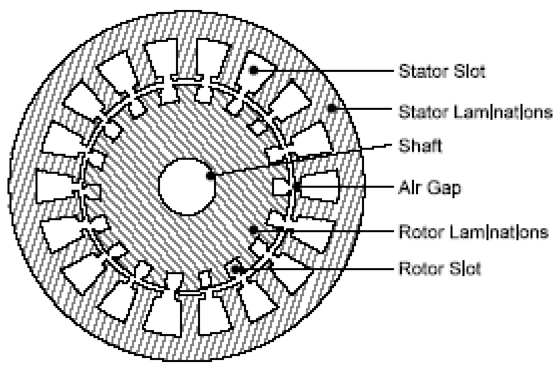


Figure 2.0 Stator and rotor laminations

ii) The rotor

This is the rotating part of the motor. As with the stator above, the rotor consists of a set of slotted steel laminations pressed together in the form of a cylindrical magnetic path and the electrical circuit. The electrical circuit of the rotor can be either:

• **Wound rotor type**, which comprises 3 sets of insulated windings with connections brought out to 3 slip rings mounted on the shaft. The external connections to the rotating part are made via brushes onto the slip rings. Consequently, this type of motor is often referred to as a *slip-ring motor*.

• **Squirrel cage rotor type**, which comprises a set of copper or aluminum bars installed into the slots, which are connected to an end-ring at each end of the rotor. The construction of these rotor windings resembles a '*squirrel cage*'. Aluminum rotor bars are usually die-cast into the rotor slots, which results in a very rugged construction. Even though the aluminum rotor bars are in direct contact with the steel laminations, practically all the rotor current flows through the aluminum bars and not in the laminations.

iii) The other parts

The other parts, which are required to complete the induction motor, are:

• *Two end-flanges* to support the two bearings, one at the drive-end (DE) and the other at the non drive-end (NDE)
• *Two bearings* to support the rotating shaft, at DE and NDE
• *Steel shaft* for transmitting the torque to the load
• *Cooling fan* located at the NDE to provide forced cooling for the stator and rotor
• *Terminal box* on top or either side to receive the external electrical connections

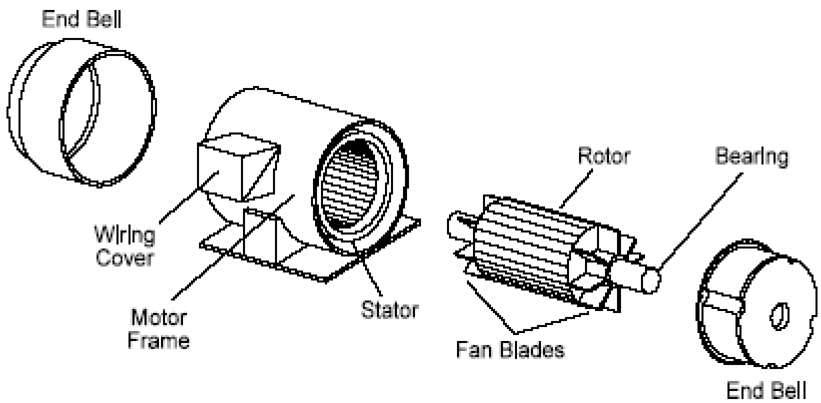


Figure 2.1 Assembly details of a typical AC induction motor

2.3 Principles of operation

When a three-phase AC power supply is connected to the stator terminals of an induction motor, three-phase alternating currents flow in the stator windings. These currents set up a changing magnetic field (flux pattern), which rotates around the inside of the stator. The speed of rotation is in synchronism with the electric power frequency and is called the *synchronous speed*.

In the simplest type of three-phase induction motor, the rotating field is produced by three fixed stator windings, spaced 120° apart around the perimeter of the stator. When the three stator windings are connected to the three-phase power supply, the flux completes one rotation for every *cycle* of the supply voltage. On a 50 Hz power supply, the stator flux rotates at a speed of 50 revolutions per second, or $50 \times 60 = 3000$ rev per minute.

The rotor magnetic field interacts with the rotating stator flux to produce a rotational force. The magnetic field in a normal induction motor is induced across the rotor air-gap as described below.

The rotating field only passes three stator windings for each power supply cycle, and will therefore rotate at half the speed of the above example, 1500 rev/min.

Consequently, induction motors can be designed and manufactured with the number of stator windings to suit the base speed required for different applications:

- 2 pole motors, stator flux rotates at 3000 rev/min
- 4 pole motors, stator flux rotates at 1500 rev/min
- 6 pole motors, stator flux rotates at 1000 rev/min
- 8 pole motors, stator flux rotates at 750 rev/min

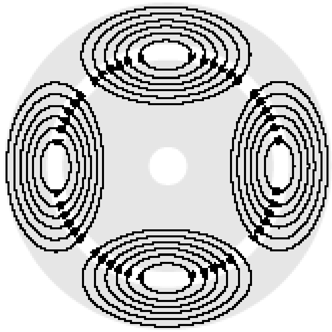


Figure 2.2 Flux distribution in a 4 pole machine at any one moment

The speed at which the stator flux rotates is called the *synchronous speed* and, as shown above, depends on the number of poles of the motor and the power supply frequency.

$$n_o = \frac{120f}{p} rev/min \qquad (1)$$

Where *n*o = Synchronous rotational speed in rev/min
f = Power supply frequency in Hz
p = Number of motor poles

Since the rotor bars are short circuited by the end-rings, current flows in these bars will set up its own magnetic field. This field interacts with the rotating stator flux to produce the rotational force.

To produce torque, the rotor must rotate at a speed slower (or faster) than the synchronous speed. Consequently, the rotor settles at a speed slightly less than the rotating flux, which provides enough torque to overcome bearing friction and windage.

Induction motors are also referred to as *asynchronous motors* because the rotor speed is not in synchronism with the rotating stator flux. The amount of slip is determined by the *load torque*, which is the torque required to turn the rotor shaft.
At no-load, the rotor torque is required to overcome the frictional and windage losses of the motor. As shaft load torque increases, the slip increases and more flux lines cut the rotor windings, which in turn increases rotor current, which increases the rotor magnetic field and consequently the rotor torque. Typically, the slip varies between about 1% of synchronous speed at no-load to about 6% of synchronous speed at full-load.

Actual rotational speed is,

$$slip = s = \frac{(n_o - n)}{n} \quad \text{p.u} \qquad (2)$$

And actual rotational speed is,

$$n = n_o(1 - s) \text{ Rev/min} \qquad (3)$$

Where n_o= Synchronous rotational speed in rev/min
n = Actual rotational speed in rev/min
s = Slip in per-unit

The direction of the rotating stator flux depends on the phase sequence of the power supply connected to the stator windings. If two supply connections are changed, the phase sequence would result in a reversal of the direction of the rotating stator flux and the direction of the rotor.

2.4 The equivalent circuit

This helps clarify what happens in the motor when stator voltage and frequency are changed or when the load torque and slip are changed.
The stator current *I*S, which is drawn into the stator windings from the AC stator supply voltage *V*, can then be predicted using this model.

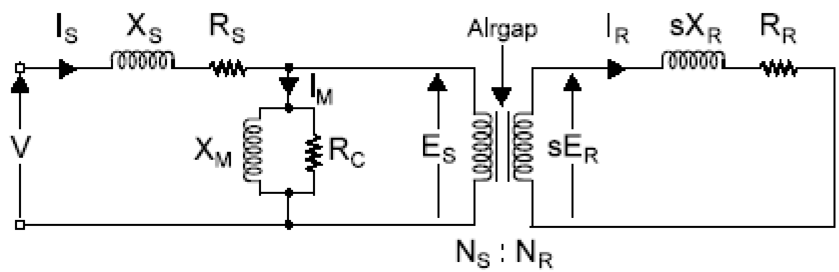


Figure 2.3 The equivalent circuit of an AC induction motor

Where *V* = Stator supply voltage *R*S = Stator resistance
*E*S = Stator induced voltage *X*S = Stator leakage reactance at 50 Hz
*E*R = Rotor induced voltage *R*R = Rotor resistance
*N*S = Stator turns *X*R = Rotor leakage reactance
*N*R = Rotor turns *X*M = Magnetizing inductance
*I*S = Stator current
*I*R = Rotor current
*I*M = Magnetizing current
*R*C = Core losses, bearing friction, windage losses.
The main components of the motor electrical *equivalent circuit* are:
• **Resistances** represent the *resistive losses* in an induction motor and comprise,
– Stator winding resistance losses (*R*S)
– Rotor winding resistance losses (*R*R)
– Iron losses, which depend on the grade and flux density of the core steel
– Friction and windage losses (*R*C)

• **Inductances** represent the *leakage reactance*. These are associated with the fact that not all the flux produced by the stator windings cross the air-gap to link with the rotor windings and not all of the rotor flux enters the air-gap to produce torque.
– Stator leakage reactance (*X*S)
– Rotor leakage reactance (*X*R)
– Magnetizing inductance (*X*M which produces the magnetic field flux)

In contrast with a DC motor, the AC induction motor does not have separate field windings. As shown in the equivalent circuit, the stator current therefore serves a double purpose:
• It carries the current (*I*M) which provides the rotating magnetic field
• It carries the current (*I*R) which is transferred to the rotor to provide shaft torque.

The equivalent circuit can be simplified even further to represent only the most significant components, which are:
• Magnetizing inductance (*X*M)
• Variable rotor resistance ($\frac{R_R}{s}$)

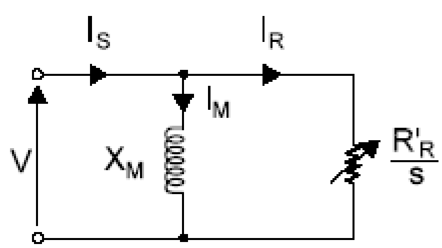


Figure 2.4 The very simplified equivalent circuit of an AC induction motor

The total stator current IS represents the vector sum of:

• The **reactive magnetizing current** *I*M, which is largely independent of load and generates the rotating magnetic field. This current lags the voltage by 90° and its magnitude depends on the stator voltage and its frequency. To maintain a constant flux in the motor, the *V*/*f* ratio should be kept constant.

$$X_M = jwL_M = j(2\pi f)M \tag{4}$$

And

$$I_M = \frac{V}{j(2\pi f)L_M} \tag{5a}$$

$$I_M = K\left(\frac{V}{f}\right) \tag{5b}$$

Where k=constant

• The **active current** *I*R, which produces the rotor torque depends on the mechanical loading of the machine and is proportional to slip. At no-load, when the slip is small, this current is small. As load increases and slip increases, this current increases in proportion. This current is largely in phase with the stator voltage.

The figure 2.5 below shows the current vectors for low-load and high-load conditions.

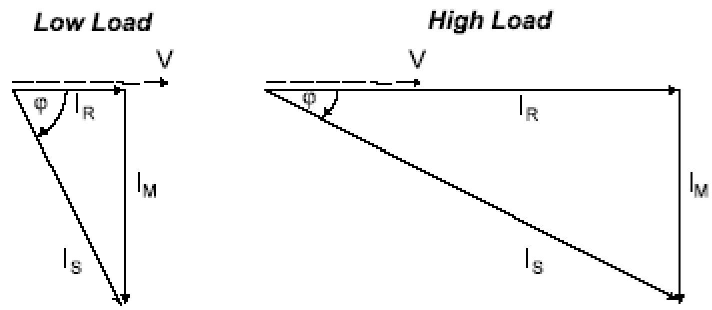


Figure 2.5 Stator current for low-load and high-load conditions

2.5 Electrical and mechanical performance

The angle between the two main stator components of voltage V and current I_S, known as the power factor angle represented by the angle φ and can be measured at the stator terminals. The stator current is the vector sum of the magnetizing current *I*M, which is in quadrature to the voltage, and the torque producing current *I*R, which is in phase with the voltage. Consequently, the total apparent motor power *S* also comprises two components, which are in quadrature to one another,

$$S = P + jQ \text{ KVA} \quad (6)$$

• *Active power P* can be calculated by

$$P = \sqrt{3} \times V \times I_R \text{ KW} \quad (7)$$

Or

$$P = \sqrt{3} \times V \times I_S \times cos\emptyset \ \ kW \quad (8)$$

• *Reactive power Q*, can be calculated by

$$Q = \sqrt{3} \times V \times I_M \ kVAr \quad (9)$$

Or

$$Q = \sqrt{3} \times V \times I_S \times sin\emptyset \ kVAr \quad (10)$$

Where S = Total apparent power of the motor in kVA
P = Active power of the motor in kW
Q = Reactive power of the motor in kVAr
V = Phase-phase voltage of the power supply in kV

IS = Stator current of the motor in amps
Ø= Phase angle between V and IS (power factor =sin ∅)

Not all the electrical input power PI emerges as mechanical output power PM. A small portion of this power is lost in the stator resistance $(3I^2R_S)$ and the core losses $(3{I_M}^2R_C)$ and the rest crosses the air gap to do work on the rotor. An additional small portion is lost in the rotor $(3I^2R_R)$. The balance is the mechanical output power PM of the rotor.

The magnetizing path of the equivalent circuit is mainly inductive. At no-load, when the slip is small (slip s ⇒ 0), the equivalent circuit shows that the effective rotor resistance $\frac{R_R}{S}$ tends to infinity. Therefore, the motor will draw only no-load magnetizing current. As the shaft becomes loaded and the slip increases, the magnitude of $\frac{R_R}{S}$ decreases and the current rises sharply as the output torque and power increases. This affects the phase relationship between the stator voltage and current and the power factor cos∅. At no-load, the power factor is low, which reflects the high component of magnetizing current. As mechanical load grows and slip increases, the effective rotor resistance falls, active current increases and power factor improves.

The torque–speed curve can be derived from the equivalent circuit and the equations above. The output torque of the motor can be expressed in terms of the speed as follows:

$$T_M = \frac{3 \times s \times V^2 \times R_R{}'}{\left[(R_S+R_R{}')^2 + s(X_S+X_R{}')^2\right] n_o} \tag{11}$$

This equation and the curve in Figure 2.6 below, shows how the motor output torque TM varies when the motor runs from standstill to full speed under a constant supply voltage and frequency. The torque requirements of the mechanical load are shown as a dashed line.

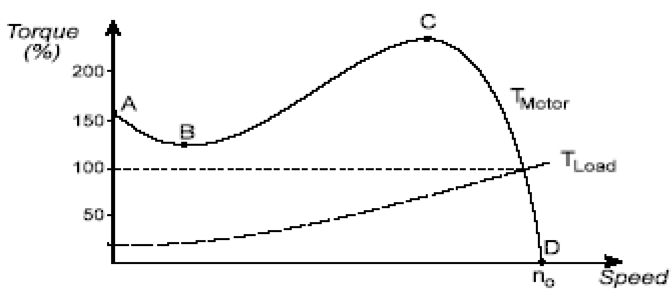


Figure 2.6 Torque-speed curve for a three-phase AC induction motor

A: is called the breakaway starting torque
B: is called the pull-up torque
C: is called the pull-out torque (or breakdown torque or maximum torque)
D: is the synchronous speed (zero torque)

At starting, the motor will not pull away unless the starting torque exceeds the load breakaway torque. Thereafter, the motor accelerates if the motor torque always exceeds the load torque. As the speed increases, the motor torque will increase to a maximum *T*Max at point C.
On the torque–speed curve, the final drive speed (and slip) stabilizes at the point where the *load torque* exactly equals the *motor output torque*. If the load torque increases, the motor speed drops slightly, slip increases, stator current increases, and the motor torque increases to match the load requirements.

The range CD on the torque–speed curve is the stable operating range for the motor. If the load torque increased to a point beyond *T*Max, the motor would stall because, once the speed drops sufficiently back to the unstable portion ABC of the curve, any increase in load torque requirements *T*L and any further reduction in drive speed, results in a lower motor output torque.
The relationship between stator current *I*S and speed in an induction motor, at its rated voltage and frequency, is shown in figure 2.7 below.

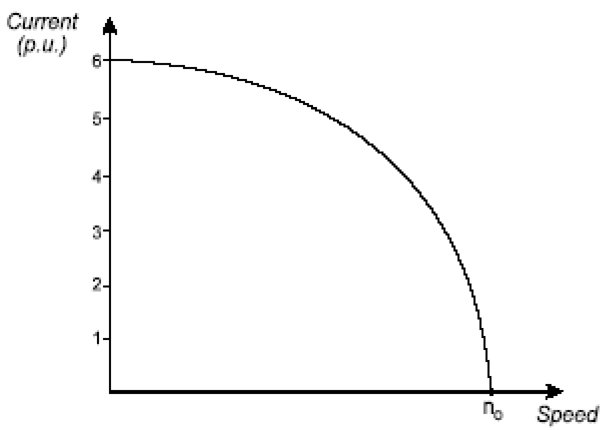


Figure 2.7 current speed characteristics of a three-phase induction motor.

CHAPTER THREE

AC DRIVES

3.1 Introduction

Ac motors require control of frequency, voltage and current for variable speed application. The power converters, inverters and ac voltage controllers, can control the frequency, voltage and/or current to meet the drive requirements. However, they are relatively complex and more expensive, and require advanced feedback control techniques such as adaptive control, sliding mode control and field vector control.

3.2 Induction motor drives

Three-phase induction motors are used in adjustable speed drives and have three phase stator and rotor windings. Stator winding are supplied with balanced three-phase ac voltages which induce voltages in the rotor windings due to transformer action.
The speed and torque o induction motors can be varied by one of the following means;

- Stator voltage control
- Rotor voltage control
- Frequency control
- Stator voltage and frequency control
- Stator current control
- Voltage, current and frequency control

To meet the torque-speed duty cycle of a drive, the voltage, current and frequency control are used.

i. **Stator voltage control**

From equation (11), torque is proportional to the square of stator supply voltage and a reduction in stator voltage causes a reduction in speed, if terminal voltage is reduced to bV_s , equation (11) gives the developed torque as,

$$T_d = \frac{3R_{r(bV_S)^2}}{sn_{o\left[\left(R_S+\frac{R_r}{s}\right)^2+(X_S+X_r)^2\right]}} \qquad (12)$$

Where $b \leq 1$

The figure 3.0 below shows the torque- speed characteristics for various values of b.
Points of intersection with load line define stable operating points.

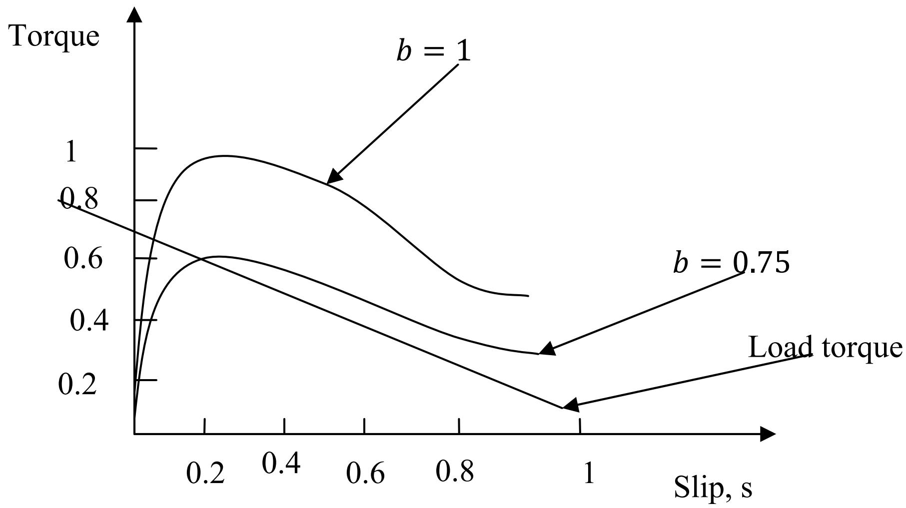


Figure 3.0 torque-speed characteristics with variable stator voltage

In a magnetic circuit, induced voltage is proportional to flux and frequency, hence rms air gap is given by,

$$V_a = bV_s = K_m \omega \emptyset \tag{13}$$

Or

$$\emptyset = \frac{V_a}{K_m \omega} \tag{14}$$

Where K_m is a constant that depends on number of turns of stator windings.asa reduction in stator voltage reduces the air gap flux and torque. This type of control is therefore not suitable for constant torque-load and is thus used in applications requiring low starting torque and a narrow range of speed.

Stator voltage can be varied by;

- Ac voltage controllers
- Voltage fed variable dc-link inverters
- PWM inverters

Ac voltage controllers have limited speed range. They are very simple but have high harmonic contents and low input power factor. Thus, they are used in low power applications.

ii. Rotor voltage control

Used in wound rotor motor where an external three phase resistor is connected to the slip rings. The developed toque is varied by varying the resistance R_x. referring Rx to the stator winding and adding to Rr, developed torque may be determined from equation (12).

This method increases starting torque while limiting the starting current. It is an inefficient method as there would be imbalance in voltage and currents if the resistances in the rotor circuit are not equal.

iii. Frequency control

When torque and speed are controlled using this method, then at rated voltage and frequency, the flux will be the rated value. If voltage is maintained fixed at the rated value while the frequency is reduced below its rated value, the flux increases and causes saturation of air gap flux and motor parameters are invalidated in determining torque sped characteristics. At a low frequency, the reactance will decrease and motor current may be too high. Hence this method is seldom used.

iv. Voltage and frequency control

If ratio of voltage to frequency is kept constant, the flux in equation (14) remains constant and thus maximum torque, which is independent of frequency, can be maintained constant. At a low frequency, however, the air gap flux is reduced due to drop in stator impedances and voltage has to be increased to maintain the torque level. This type of control is therefore known as volts/hertz control.
Torque-speed characteristics for volts/hertz control are as shown in the figure 3.1 below.

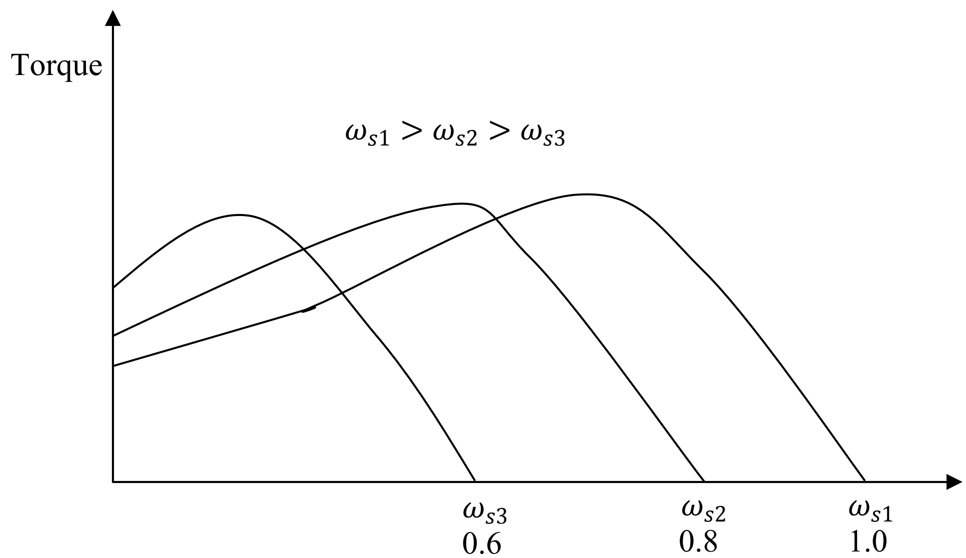


Figure 3.1 Torque-speed characteristics with volts/hertz control

Three possible circuit arrangements, for obtaining variable voltage and frequency in three-phase inverters are;

a. Fixed dc and PWM inverter drive.
b. Variable dc and inverter drive.
c. Variable dc from dual converter and inverter.

In case **a**, the dc voltage remains constant and the PWM techniques are applied to vary both the voltage and frequency within the inverter. Due to the diode rectifier, regeneration is not possible and the inverter generates harmonics into the ac supply.

In case **b**, the chopper varies the dc voltage to the inverter and the inverter controls the frequency. The chopper reduces harmonic injection into the ac supply.

In case c, the voltage is varied by the dual converter and frequency is controlled within the inverter. It permits regeneration but the input power factor of converter is low (especially at high delay angle).

3.3 Thyristor control of ac motors

The thyristor or SCR is a semiconductor device that is capable of controlling large currents. The thyristor converter is used as a variable speed drive. The static variable frequency ac drive uses a cage rotor induction motor or synchronous reluctance motor powered by a static frequency converter. This gives a versatile and robust variable speed machine. Which has the advantages over conventional variable speed drives of higher accuracy-better reliability, reduced maintenance and higher efficiency.

3.3.1 Generation of variable frequency ac power

i. Rotating frequency converters

They are used principally in multimotor mill drives and in special a[applications where a high operating frequency is chosen in order to permit the use of compact ac motors.

ii. Static frequency converters

They improve the performance and reliability if used in place of rotating frequency converters. To obtain high frequency it is essential to use static switching devices which are either on off. In the on condition, the device approximates to an ideal closed switch having zero voltage drop across it and a current which is determined by the external circuit.
In the off condition, the device approximates an ideal open switch which has infinite impedance and blocks the flow of current in the circuit.
If the solid state switch can be triggered from the off condition to on condition by a low power control signal, the device can be used in converter circuits for generation of variable frequency alternating voltages.

The thyristor is ideal because of

1. Its availability in high power ratings.
   A thyristor can be triggered into conduction by an external signal on the gate electrode and can also be switched off by an external signal. Turn off or commutation can be achieved by interrupting the anode current.
   If the thyristor operates on ac supply, natural commutation occurs. In dc operated circuits, current does not go to zero naturally and forced commutation must be used. This technique employs auxiliary charged capacitors to force the current in a conducting thyristor to zero.

2. The thyristor is a more efficient switching device since the voltage drop in the on condition is approximately 1volt.
3. Switching times of thyristor are of order of magnitude lower than those of thyratron. The reduced turn- off times permits a significant reduction in auxiliary apparatus necessary to achieve forced commutation.
4. The thyristor is also more rugged and durable device and is more compact, even with its associated heat sink. Being a semiconductor, it has no warm up time and it eliminates the filament heating supply of thyratron.

3.4.0 PULSE WIDTH MODULATED (PWM) INVERTERS.
3.4.1 Introduction

An inverter is a DC-to-ac converter, whose function is to change a dc input voltage to a symmetrical ac output voltage of desired magnitude and frequency. Output voltage could be fixed or variable at a fixed or variable frequency.

A variable output voltage is obtained by varying the input ac voltage and maintaining the gain of the inverter constant; or the dc voltage is fixed and a varying output voltage obtained by varying the gain of the inverter which is accomplished by pulse-width-modulation (PWM) control within the inverter. The inverter gain may be defined as the ratio of ac output voltage to dc input voltage.

Output voltage waveforms of ideal inverter are sinusoidal but practical waveforms of practical inverters are non-sinusoidal and contain certain harmonics.
For low and medium power applications, square wave or quasi square wave voltages may be acceptable; and for high power applications, low distorted sinusoidal waveforms are required. With the availability of high-speed power semiconductor devices, harmonic content of output voltage waveforms can be minimized.

Inverters, single and three- phase, use controlled turn-on and turn–off devices (BJT's, MOSFET's, IGBT's, MCT's, GTO's, et cetra) or forced commutated thyristors. These inverters use PWM control signals for producing an ac output voltage.

A voltage fed inverter (VFI) is one with a constant input voltage, while a current fed inverter (CFI) is one with a constant input current. A variable dc linked inverter is one with controllable input voltage.

3.4.2 Three-phase inverters

They are used for high power applications and a three-phase output can be obtained from a configuration of six switching devices as shown in the figure 3.3 below

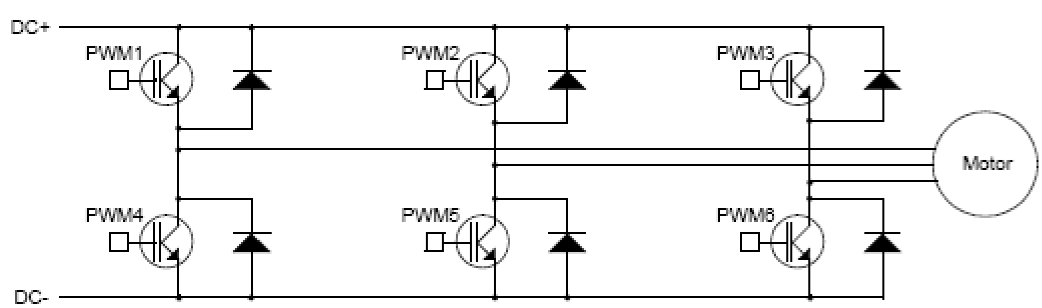


Figure 3.3 Three-phase inverter

Two types of control signals can be applied to the switching devices;

- 180° conduction
- 120° conduction

i. **180-Degree conduction**.

Each switching device conducts for 180°. Three switches remain ON at any instant of time. When Q_1 is switched on, terminal, a, is connected to the positive terminal of dc input voltage. When Q_4 is switched on, terminal a, is brought to the negative terminal of input dc voltage. Six modes of operation exist and each lasts 60° in conduction. The switches are numbered in the sequence of firing or gating (123, 234, 345, 456, 561, and 612). The gating signals are shifted from each other by 60° as shown in the figure 3.4 below.

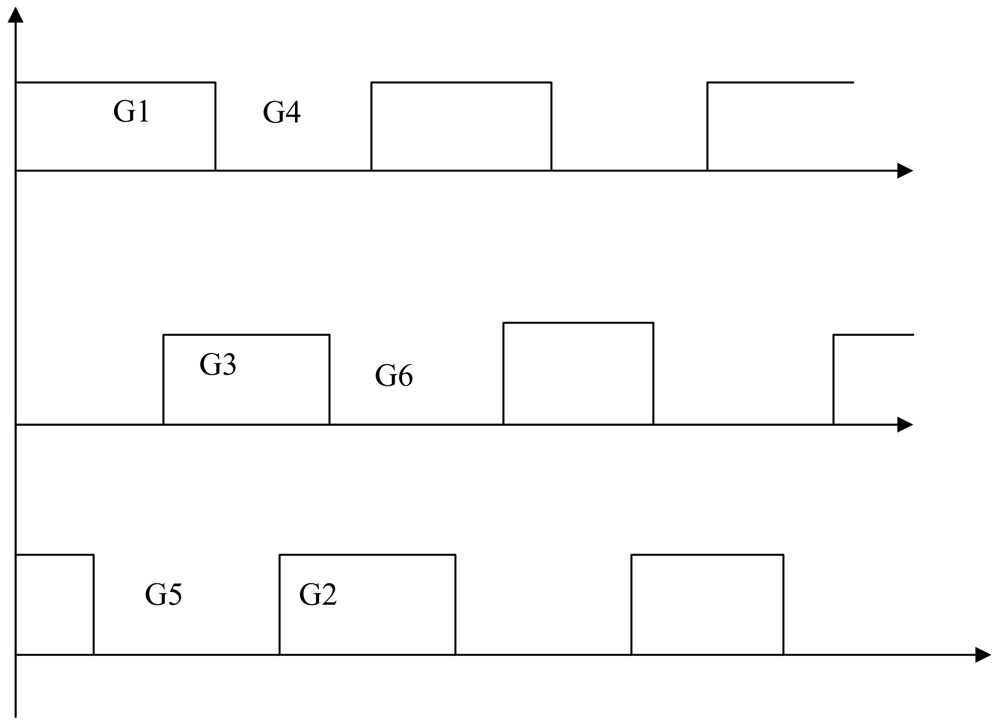


Figure 3.4 gating signals for 180° conduction

The load may be connected in wye or delta. For a delta connected load, the phase currents are obtained directly from the line-to-line voltages. With phase currents known, line currents can be determined. For wye-connected load, the line to neutral voltages must be determined to find the line (or phase) currents. Three modes of operation exist in a half cycle,

Mode 1 for $0 \leq \omega t \leq \frac{\pi}{3}$

During this mode,

$$R_{eq} = R + \frac{R}{2} = 1.5R \tag{15}$$

$$i_1 = \frac{V_s}{R_{eq}} = \frac{2V_s}{3R} \tag{16}$$

$$v_{an} = v_{cn} = \frac{i_1 R}{2} = \frac{V_S}{3} \quad (17)$$

$$v_{an} = -i_1 R = \frac{-2V_S}{3} \quad (18)$$

Mode 2 for $\frac{\pi}{3} \leq \omega t \leq \frac{2\pi}{3}$

During this mode,

$$R_{eq} = R + \frac{R}{2} = 1.5R \quad (19)$$

$$i_2 = \frac{V_S}{R_{eq}} = \frac{2V_S}{3R} \quad (20)$$

$$v_{an} = i_2 R = \frac{2V_S}{3} \quad (21)$$

$$v_{bn} = v_{cn} = \frac{-i_2 R}{2} = \frac{-V_S}{3} \quad (22)$$

Mode 3 for $\frac{2\pi}{3} \leq \omega t \leq \pi$

$$R_{eq} = \frac{3R}{2} \quad (23)$$

$$i_3 = \frac{V_S}{R_{eq}} = \frac{2V_2}{3R} \quad (24)$$

$$v_{an} = v_{bn} = \frac{i_3 R}{2} = \frac{V_S}{3} \quad (25)$$

$$v_{cn} = -i_3 R = \frac{-2V_S}{3} \quad (26)$$

The phase voltages are shown in figure b. The instantaneous line-to-line voltage, v_{ab}, expressed in Fourier series, with the recognition that v_{ab} is shifted by, $\frac{\pi}{6}$ rad, and that the even harmonics being zero,

$$v_{ab} = \sum_{n=1,3,5,..}^{\infty} \frac{4V_S}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t + \frac{\pi}{6}\right) \quad (27)$$

v_{bc} and v_{ca} are obtained by phase shifting v_{ab} in the equation(27), above by 120° and 240° respectively:

$$v_{bc} = \sum_{n=1,3,5,..}^{\infty} \frac{4V_S}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t - \frac{\pi}{2}\right) \quad (28)$$

$$v_{ca} = \sum_{n=1,3,5,..}^{\infty} \frac{4V_S}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t - \frac{7\pi}{6}\right) \quad (29)$$

Note:
Triplen harmonics (n=3, 9, 15….) will be zero, from equations (28) and (29) of the line-to-line voltages.

The line-to-line rms voltage,

$$V_L = \left[\frac{2}{2\pi}\int_0^{2\pi/3} V_s{}^2 d(\omega t)\right]^{\frac{1}{2}} = \sqrt{\frac{2}{3}V_s = 0.8165V_s} \tag{30}$$

The rms n^{th} component of line voltage is,

$$V_{Ln} = \frac{4V_s}{\sqrt{2}\, n\pi}\cos\frac{n\pi}{6} \tag{31}$$

Which for n=1 gives,

$$V_{L1} = \frac{4V_s \cos 30°}{\sqrt{2}\,\pi} = 0.7797V_s \tag{32}$$

The rms value of phase voltages is,

$$V_p = \frac{V_L}{\sqrt{3}} = \frac{\sqrt{2}V_{\ s}}{3} = 0.4714V_s \tag{33}$$

ii. 120-degree conduction

In this type of control, each switching device conducts for 120°. Only two switches remain on at a time. The gating signals are as shown in figure 3.6 below.

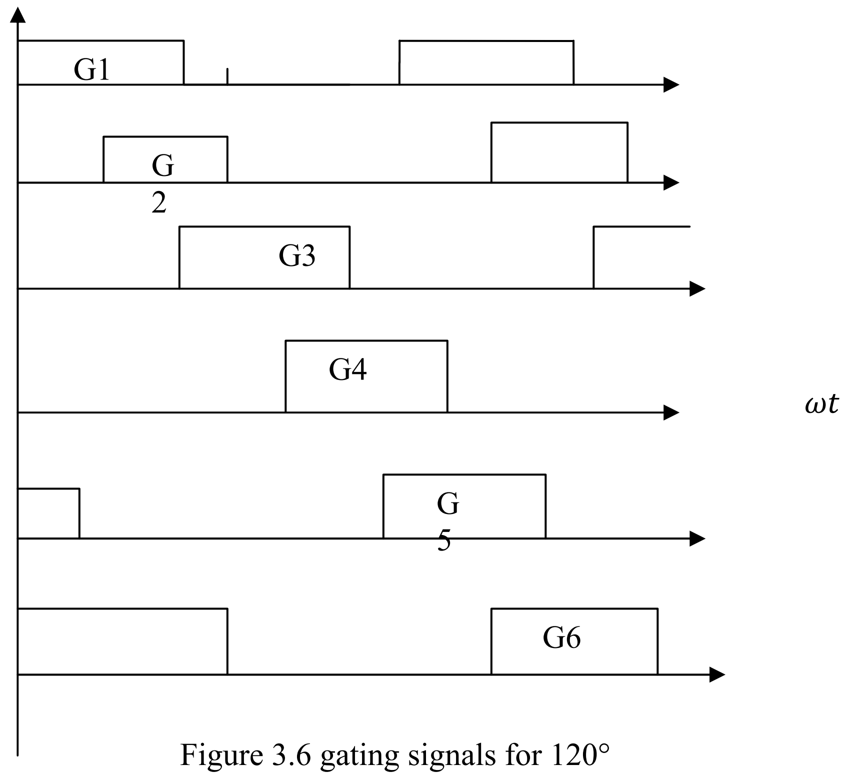


Figure 3.6 gating signals for 120° conduction

There are three modes of operation as well, and the gating sequence is, 61, 12, 23, 34, 45, 56, and 61.

Mode 1 for $0 \le \omega t \le \frac{\pi}{3}$

Switching devices 1 and 6 are on.

$v_{an} = \frac{V_s}{2}$, $v_{bn} = \frac{-V_s}{2}$, $v_{cn} = 0.$

Mode 2 for $\frac{\pi}{3} \le \omega t \le \frac{2\pi}{3}$

Switches 1 and 2 are on.

$v_{an} = \frac{V_s}{2}$, $v_{bn} = 0$, $v_{cn} = \frac{-V_s}{2}.$

Mode 3 for $\frac{2\pi}{3} \le \omega t \le \pi$

Switches 2 and 3 are on.

$v_{an} = 0,\ v_{bn} = \frac{V_s}{2}\ ,\ \ v_{cn} = \frac{-V_s}{2}$.

The Phase voltages expressed in Fourier series are,

$$v_{an} = \sum_{n=1,3,5..}^{\infty} \frac{2V_s}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t + \frac{\pi}{6}\right) \tag{34}$$

$$v_{bn} = \sum_{n=1,3,5..}^{\infty} \frac{2V_s}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t - \frac{\pi}{2}\right) \tag{35}$$

$$v_{cn} = \sum_{n=1,3,5..}^{\infty} \frac{2V_s}{n\pi} \cos\frac{n\pi}{6} \sin n\left(\omega t - \frac{7\pi}{6}\right) \tag{35}$$

The line a-to-b voltage, $v_{ab} = \sqrt{3}\, v_{an}$, with a phase advance of 30°. There is a delay of *π/6 rad* between turn-off of Q_1 and turn-on of Q_4.

3.4.3 Pulse Width Modulation (PWM) switching scheme

To produce a sinusoidal output voltage waveform at a desired frequency, a sinusoidal control signal at the desired frequency is compared with a triangular waveform. The frequency of triangular waveform,$f_{s,}$ establishes the inverter switching frequency which is kept constant with its amplitude, v_{tri}.
The control signal,v_{con}, is used to modulate the switching duty ratio and has a frequency, f, which is the desired fundamental frequency of inverter output voltage.
Amplitude modulation ratio is given by,

$$m_a = \frac{Vcont}{Vtri} \tag{35}$$

Frequency modulation ratio is given by,

$$m_f = \frac{f_s}{f} \tag{36}$$

Thus, the peak amplitude of the fundamental frequency component is given by,

$$V_o = m_a \times \frac{V_d}{2} \tag{37}$$

Therefore the amplitude of the fundamental component varies linearly with m_a (provided$m_a \leq$ 1).
For $m_f \leq 9$ (which is always the case except in high power ratings), harmonic amplitudes are almost independent ofm_f, though m_f defines the frequency at which they occur. That is,

$$f_h = \left(jm_f \pm k\right)f \tag{38}$$

Where,

$$h = j\left(m_f\right) \pm k$$

The fundamental frequency corresponds to h=1. For odd values of j, the harmonics exist only for odd values of k.

3.4.3 Selection of switching frequency and frequency modulation ratio.

Because of the relative ease in filtering harmonic voltages at high frequencies, it is desirable to use a high switching frequency as possible, except for one drawback; switching losses in inverter switches increase proportionally with the switching frequency f_s. Normally f_s, is either less than 6 kHz or greater than 20 kHz.

In 50-60 Hz type applications such as ac motor drives,$m_f \leq 9$, for switching frequencies less than 2 kHz. m_f will be larger than 100 for f_s> 20kHz.
The desirable relationship between v_{tri} and v_{con} waveforms are dictated by how large m_f is.
For synchronous PWM, triangular waveform frequency varies with desired inverter frequency.

For example;
If f=65.42 Hz and m_f=15, then, $f_s = 15 \times 65.42 = 981.3\ Hz$

3.5 THREE-PHASE RECTIFIERS.

Three phase rectifier circuits are preferred to single phase circuits because of their lower ripple content in the waveforms and a higher power handling capability. A filter is usually connected at the dc side of the rectifier.
Idealized circuit with Ls=0 is as shown in figure 3.7 below;

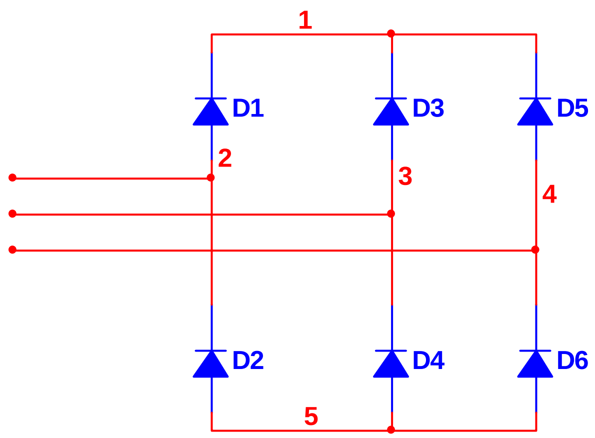


Figure 3.7 uncontrolled three-phase rectifier.

The current i_d flows through one diode in the top group and one from the bottom group. The diode with its anode at the highest potential will conduct (from the top group), and the other two become reverse biased. In the bottom group, the diode with its cathode at the lowest potential will conduct and the other two become reverse biased.

is the voltage at point P with respect to ac voltage neutral point n. is the voltage at the negative dc terminal N.

Since flows continually, and can be obtained in terms of one of the ac input voltages and

Applying KVL;

(39)

The instantaneous waveform of consists of six segments per cycle of line frequency. Hence it is termed as a six pulse rectifier. Each segment belongs to one of the six, line-to -line voltage combinations.

Each diode conducts for 120 degrees as depicted in the figure 3.9 below.

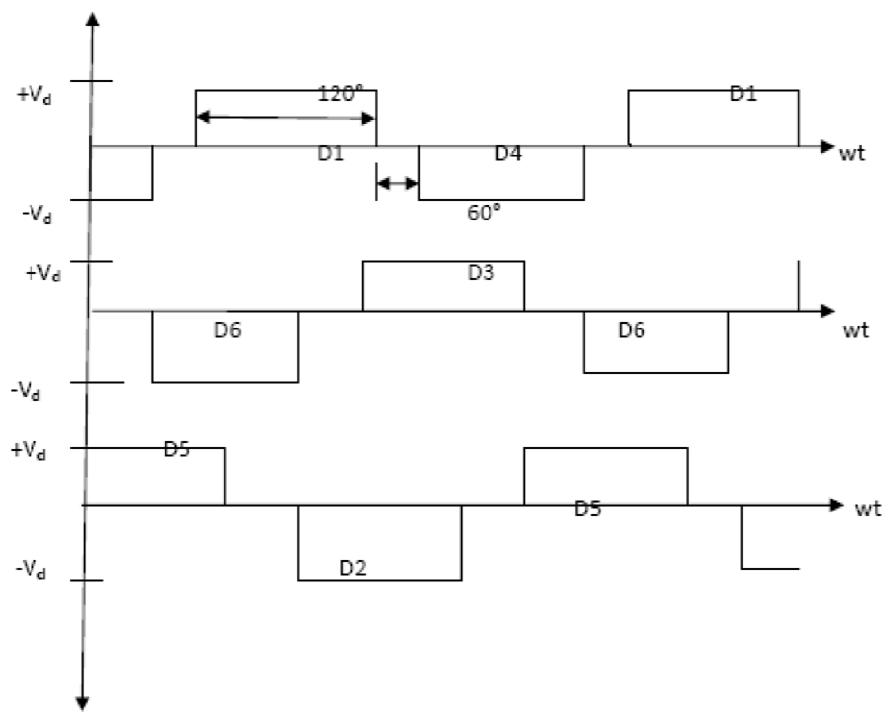


Figure 3.9 diode conduction

Considering the phase *a* current waveform;

, when diode D1 conducts and

, when D4 conducts

,when neither conducts.

Commutation of current from one diode to the next is instantaneous based on the assumption that Ls =0.

To obtain average values of dc output voltage, only one of the six segments is considered and its average value over 60° ($\pi/3$ rad) interval. Taking the reference time (t_o) at $v_{abmax.}$, then,

$$v_d = v_{ab} = \sqrt{2}\, V_{LL} \cos \omega t \;\; d(\omega t), \text{ for } \;\; -\pi/6 < wt < +\pi/6 \qquad (40)$$

V_{LL} =rms value of line–to-line voltage.
Integrating v_{ab} over the given interval yields the area A=$\sqrt{2}\, V_{LL}$.
Therefore,

$$V_{do} = \frac{3A}{\pi} = \frac{3}{\pi} \int_{\frac{-\pi}{6}}^{\frac{\pi}{6}} \sqrt{2}\, V_{LL} \cos \omega t \; d(\omega t) = 1.35 V_{LL} \qquad (41)$$

The rms value of line current is given as,

$$I_s = \sqrt{\frac{2}{3}}\, I_d = 0.816\, I_d. \qquad (42)$$

By means of Fourier analysis of i_s, the fundamental frequency component i_{s1} ha an rms value $I_{s1} = \frac{1}{\pi} \sqrt{6}\, I_d = 0.78 I_d$

The harmonic components I_{sh} can be expressed in terms of the fundamental frequency component as,

$$I_{sh} = \frac{I_{s1}}{h} \text{ ; h= 5, 7, 11, 13…} \qquad (43)$$

The even and triplen harmonics are zero.

Since i_{s1} is in phase with the utility phase voltage,

DPF = 1.0
and

PF = 1

Line frequency diode rectifiers are used to convert 50-60 Hz ac input into dc voltage in an uncontrolled manner.

CHAPTER FOUR

4.0 DESIGN

4.1 RECTIFIER DESIGN

Three-phase full-wave rectification was chosen, and the choice of diodes was governed by the expected load current $I_{dc} = 20\ amperes$, the peak inverse voltage, and the line to neutral voltage which was assumed to be 120V at supply frequency 50 Hz. The average current of a diode in the three-phase rectifier is,

$$I_d = \frac{I_{dc}}{3} = \frac{20}{3} = 6.7 \approx 7.0\ amps. \qquad (44)$$

The rms current is,

$$I_{rms} = \frac{I_{dc}}{\sqrt{3}} = \frac{20}{\sqrt{3}} = 11.55\ amps \qquad (45)$$

The peak inverse voltage, PIV, is the maximum voltage that can be safely applied across the diode when reverse biased. Therefore,

$$PIV = \sqrt{3}\, V_m = \sqrt{3} \times \sqrt{2} \times v_{an} = 294\text{V} \qquad (46)$$

The output ripple voltage V_{ac} is found approximately from equation, (47) below. In this design, the ripple voltage was considered negligible.

$$V_{ac} = \frac{V_{rms(pp)}}{2\sqrt{2}} \qquad (47)$$

The ripple factor RF is found from,

$$RF = \frac{V_{ac}}{V_{dc}} \qquad (48)$$

From the above specifications, the diode used is 1N5404, which had the following properties,

$$I_{rrm} = 300A$$
$$V_{rrm} = 400V$$
$$V_F = 0.6V$$

4.2 Inverter and commutation design

Selection of thyristors for the inverter was based on,

- Required load current, I_L=20A.
- Peak inverse voltage to be applied during commutation, PIV=294V.

Thyristor BT 152 with the specifications below was chosen,

$$I_{on} = 20A$$
$$PIV = 400V$$
$$I_{gt} = 32mA$$

$$V_{gt} = 1V$$

4.2.1 Commutation circuit design

Factors considered include,

- Energy loss per commutation must be low.
- Trapped energy should be low.
- Commutation circuit turn-off time.
- Performance was predicted based on calculation.

The main commutating components are the L-C circuit; the diodes provide a path for return of reactive power to the supply and provide the necessary reverse voltage for turning off the thyristors.

$$t_{off\ max} = \frac{\pi}{3}\sqrt{2L_m C_m} \qquad (49)$$

Therefore, choosing Cm to be 100uF, Lm is calculated to be,

$$L_m = \left(\frac{3}{\pi}\ t_{off}\right)^2 \div 2C_m = 10\mu H \qquad (50)$$

From the calculated values of L and C, performance was predicted as below;
During resonance, it was important that the duration of time for which current through the L-C circuit exceeds load current, should be greater than thyristor turn-off time. The excess current flows through the diode, providing the necessary reverse voltage across the thyristor.

4.2.2 Frequency control and generation of gating signals circuitry

Pulse width modulation was achieved by using two 555 timers configured as shown in the figure 4.0 below.

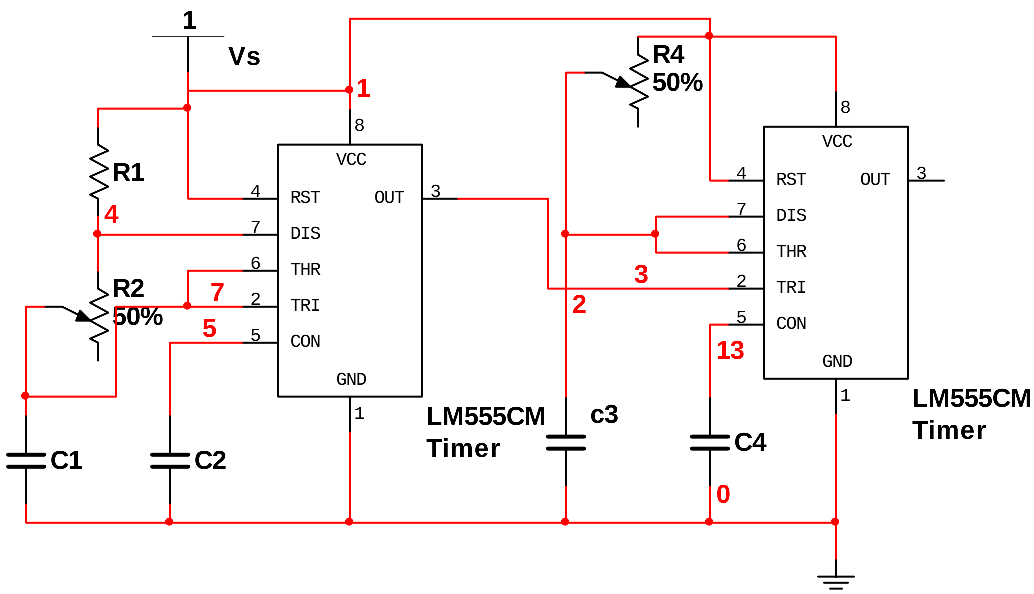


Figure 4.0 Pulse width modulation circuit using 555 timers.

The waveforms in the figure 4.1 below help to explain the way in which this method of control operates. In each case the signal has maximum and minimum voltages of 5V and 0V.
In waveform 1a, the signal has a mark-space ratio of 1:1, with the signal at 5V for 50% of the time.

In waveform 1b, the signal has a mark-space ratio of 3:1, which means that the output is at 5V for 75% of the time.

In waveform 1c, the signal has a mark-space ratio is 1:3, giving an output signal that is 5V for just 25% of the time
By varying the mark-space ratio of the signal over the full range, it is possible to obtain any desired average output voltage from 0V to 5V.

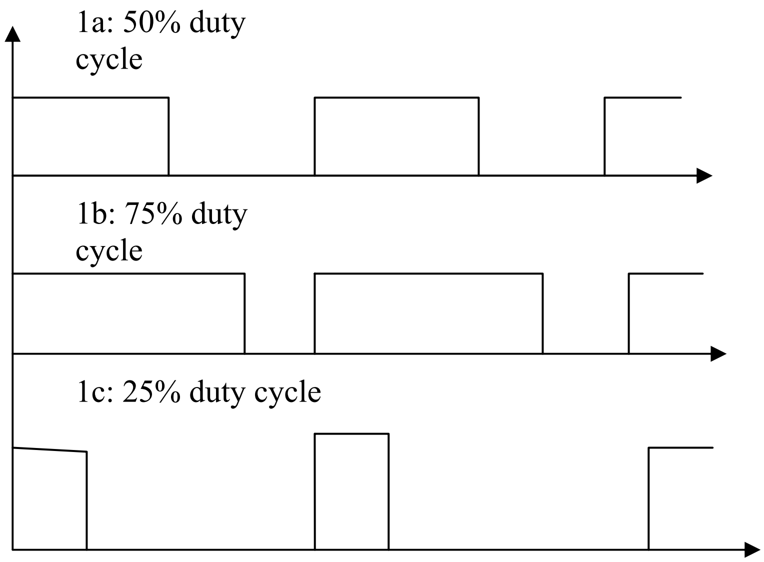


Figure 4.1 PWM waveforms

4.2.3 Designing a PWM Circuit

The concept of PWM requires timing. Two 555 timer ICs and some potentiometers are used to generate a PWM signal, and since PWM provides a digital, on/off signal, it is also easy to use a PC or micro-controller to create the signal.

The circuit in figure 4.0, uses two 555 ICs and is a combination of two types of circuit. The first is a free running multivibrator (astable) with an adjustable frequency. The output of this circuit then triggers a pulse shaping (monostable) circuit which adjusts the width of the pulse. The circuit produces a duty cycle in the range of approximately 0.3% to 97%.

The frequency preset (R2) of the 555 astable circuit allows the frequency of the signal to be adjusted so that the duty ratio potentiometer (R4) can achieve its full range. Modulation frequency is measured at pin 3 of the 555 astable.

The output of the monostable timer was then used to clock a five stage Johnson counter, which is then connected as in the figure below, to generate three signals, 120 degrees out of phase. The other three signals are the inverted outputs of the first three signals.

The outputs of the counter were combined in such a way that the firing of the thyristors was as shown in the table 4.0 below,

T1	1	1	1	0	0	0
T1'	0	0	0	1	1	1
T2	0	0	1	1	1	0
T2'	1	1	0	0	0	1
T3	1	0	0	0	1	1
T3'	0	1	1	1	0	0

Table 4.0 required firing logic state of thyristors gates.

The count was made up to six, since six gating signals were required.

For T1 to be on, the MSB of the count was always required to be a 1. Possible combinations for these are, 101, 100, and 110, whose decimal equivalents are 4, 5, and 6.
The count combination for T2 is thus, 6, 1, and 2 and for T3, the count combination is 2, 3, and 4. The combination for the other signals is the inverted count of their respective thyristors gate logic.

4.2.4 Clock pulse generator

Timer configuration used is as shown in the figure 4.2 below.

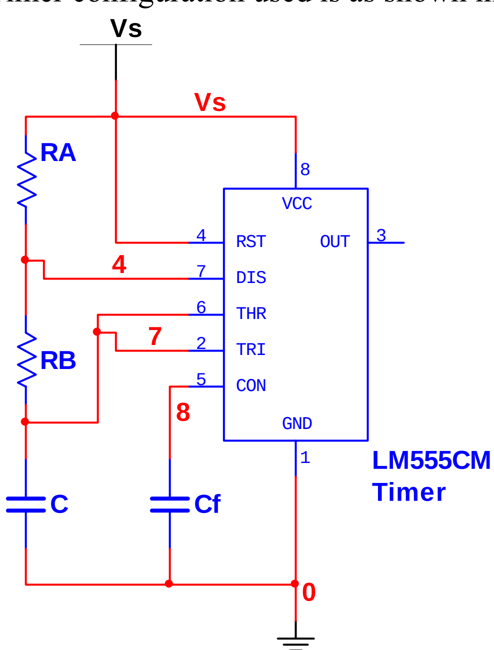


Figure 4.2 clock pulse generator

To maintain the phase displacement angle at 60°, a clock pulse was fed to the counter and the frequency of the clock determines the output frequency of the inverter.

The output waveform is as shown in the figure 4.1 above.

Where,

$$T1 \approx 0.7RC(R_A + R_B)$$
$$T2 \approx 0.7R_B. \tag{51}$$

The output frequency,

$$f = \frac{1}{T} = \frac{1}{T1+T2} \tag{52}$$

Thus,

$$f = \frac{1.44}{C(R_A+2R_B)} \tag{53}$$

The duty cycle,

$$D = \frac{R_A+R_B}{R_A+2R_B} \tag{54}$$

By making RB variable, output frequency can be made variable.
Fixing the output frequency of inverter at maximum frequency of 100 Hz, and a minimum frequency of 20 Hz, time period of clock pulse for 100 Hz inverter frequency is,

$$T_1 = \frac{1}{3} \times \frac{1}{4\times100} = 0.83\ mS \tag{55}$$

Therefore maximum timer frequency,

$$f_1 = \frac{1}{T_1} = \frac{10^3}{0.83} = 1200\ Hz \tag{56}$$

For minimum inverter frequency of 20 Hz, time period of clock pulse,

$$T_2 = \frac{1}{3\times4\times20} = 4.17\ mS \tag{57}$$

Therefore minimum frequency of timer is given by,

$$f_2 = \frac{10^3}{4.17} = 240\ Hz \tag{58}$$

At f =1200 Hz, let RB = 0 and C = 0.1μF, then,

$$R_A = \frac{1.4}{f\times C} = \frac{1.4\times10^6}{0.1\times1200} = 11.6\ k\Omega \tag{59}$$

At the minimum frequency of 240 Hz,

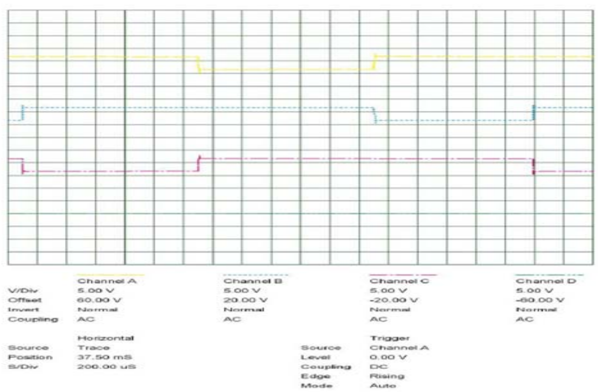
$R_A + R_B = \frac{1.4\times10^6}{0.1\times240} = 58.3\ k\Omega$ (60

Implying that $R_B = \frac{1}{2}(58.3 - 11.6)k\Omega = 23.35\ k\Omega$
Thus, RB can be varied from 0-24 k**Ω.**

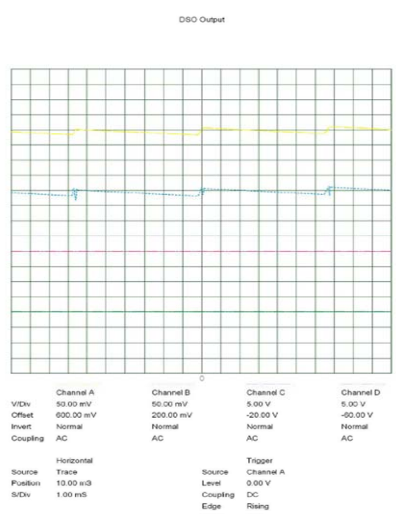
CHAPTER FIVE

5.0 RESULTS

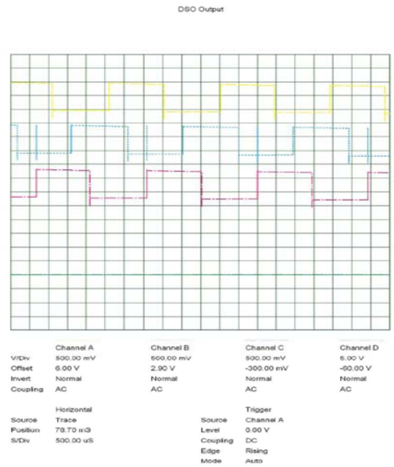
RESULTS



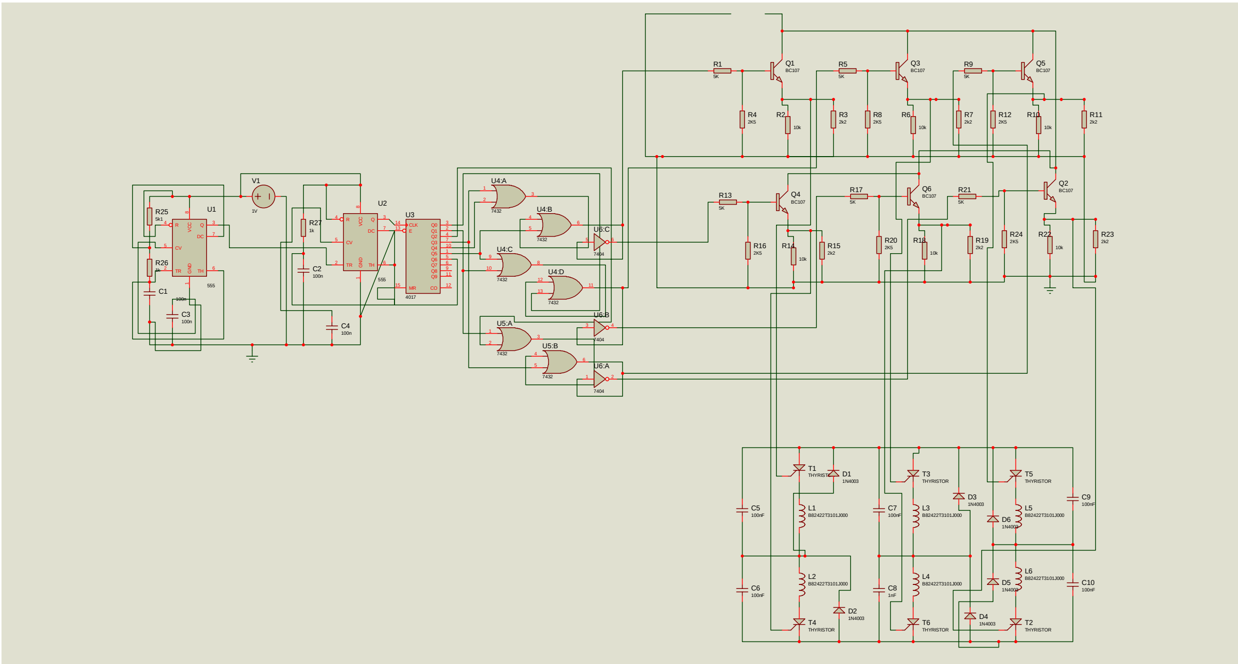
PWM inverter output voltage waveforms.



Filtered rectifier output waveforms



Gating signals



Complete circuit diagram.

CHAPTER SIX

6.0 CONCLUSION AND RECOMMENDATIONS.

Due to insufficient resources, the entire project goals were not achieved especially commutation in the inverter.
The control circuitry was well designed and behaved as expected

Recommendations

Control of the inverter can be improved even further to incorporate feedback loops by using micro controllers which can be programmed to handle complex algorithms.

Instead of using thyristors, power MOSFETs or IGBTs can be used with their appropriate gate drivers. This helps to eliminate the problem of commutation in thyristor based inverters.

Chapter seven

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