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Addis Ababa Institute of Technology(AAiT)
Electrical and Computer Engineering Department

DESIGN AND SIMULATION OF FUZZY LOGIC BASED SELF-TUNING PI CONTROLLER FOR PERMANENT MAGNET SYSCHRONOUS MOTORS

By

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Declaration

I, the undersigned declare that this thesis is my original work, and has not been presented for a degree in this or other university, and all sources of materials used for the thesis have been fully acknowledged.

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Addis Ababa,Ethiopia May ,2017

Place Date Of Submission

This thesis has been submitted with my approval as a university advisor

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Advisor name signature

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LIST OF ABBREVIATIONS

PMSM	Permanent magnet synchronous motor
AC	Alternate current
PID	Proportional, integral and derivative
PM	Permanent magnet
DC	Direct current
MRAC	Model reference adaptive control
VSC	Variable structure control
FLC	Fuzzy logic control
MATLAB	Matrix laboratory
BDCM	Brushless dc motor
EMF	Electromotive force
DSP	Digital signal processor
PWM	Pulse width modulation
SVPWM	Space vector pulse width modulation
RPM	Revolution per minute
SM	Synchronous machine
SPMSM	Surface mounted permanent magnet synchronous machine
IPMSM	Interior permanent magnet synchronous machine
SPM	Surface permanent magnet motor
IPM	Interior permanent magnet motor
BLDC	Brushless dc motor
FOC	Field oriented control
SPMSM	Surface permanent magnet synchronous machine
MOSFET	Metal oxide sulfate field effect transistor
IGBT	Integrated gate bipolar transistor

ABSTRACT

This thesis presents the methods used in speed control of permanent magnet synchronous motor drives. PI control is used as speed controller for permanent magnet synchronous motor in high performance drive system. For the linear system, PI controller is easily tuned and perfectly controls the system with good performance. PMSM is non-linear system by nature. However, PI controller is sensitive to uncertainty of the model, parameter variation and load disturbance. The conventional approach to these issues is to tune the proportional and integral gains manually by observing the response of the system. The well-known conventional approach, Ziegler-Nichols method to tune the coefficient of a PI controller is very simple but cannot guarantee to be always effective to avoid tedious tasks in manual control. The tuning of PI parameters should also be done on-line for better performance.

This thesis presents design of self-tuning PI controller schemes based on fuzzy logic approach. The performance of proposed controller was tested at different operating conditions of reference speeds as well as with load variations through simulations using matlab/Simulink software package. The simulation results reveal, for no-load torque over shoot is 34.85 % and 3.38 % for PI-controller and proposed controller respectively. When 0.25Nm load is applied at 0.02sec, over shoot is 32.9 % and 3.56 % for PI and Fuzzy logic based self-tuning PI controller. The simulations results also show that the speed response of proposed controller is not altered by small change in load torque and parameter variations. When the values of stator resistance and inductance is increased by 50%, speed response shows 31.97% and 34.64% overshoot with conventional PI, where as 3.84% and 0.99% overshoot with proposed controller respectively. Further, the designed controller works well even when the reference speed is changed to more complex trajectory which means that, the developed controller speed regulation is unaltered by the nature of reference speed trajectory. These shows remarkable transient response performance improvement compared to conventional PI controller.

Keywords: PMSM, Self-tuning PI-controller, PI-controller, Fuzzy logic,Matlab/simulink.

CHAPTER ONE

Introduction

1.1. Background theory

The synchronous machine is constant speed machine which always rotate at synchronous speed, which depend on the supply frequency and on the number of poles. The existence of modern permanent magnets (PM) with considerable energy density led to the development of DC machines with PM field excitation in the 1950s. Introduction of PM to replace electromagnets, which have windings and require an external electric energy source, resulted in compact dc machines. The synchronous machine, with its conventional field excitation in the rotor is replaced by the PM excitation; The slip rings and brush assembly are dispensed with the advent of switching power transistor and silicon – controlled-rectifier devices in later part of 1950s, the replacement of the mechanical commutator with an electronic commutator in the form of an inverter was achieved. These two developments contributed to the development of PM synchronous and brushless dc machines. The armature of the dc machine need not be on the rotor if the machine commutator is replaced by its electronic version. Therefore, the armature of the machine can be on the stator, enabling better cooling and allowing higher voltages to be achieved: significant clearance space is available for insulation in the stator. The excitation field that used nothing but ‘an inside out dc machine’ with the field and armature interchanged from the stator to rotor and rotor to stator, respectively [1][2][3].

Permanent magnet synchronous motors (PMSM) are widely used in AC servo drives because of its high torque to inertia ratio, high power density, high efficiency and power factor as compare to the other motors used in drives.

For many years d.c motors were used extensively in areas where variable speed operation was required since their flux and torque could be controlled independently and easily by the field and armature current. Particularly, the separately excited d.c motor has been used mainly for applications where there was a requirement of fast response and four-quadrant operation with high performance near zero speed. However, due to the existence of the commentator and the brushes, d.c motors have certain disadvantage. That is, they required periodic maintenance; they cannot be used in explosive or corrosive environments. These problems can be overcome by the application of a.c motors, because of its simplicity, efficiency, low cost, compactness, and economical and volume manufacturing advantages. Some of the reasons for dealing with AC motor drives rather than DC motor drives are:

- PM synchronous motors comparable and frequently better efficiency than the equivalent DC motor.
- Commutator and brushes do not exist in PMSM which can result in higher speed operations.
- A dc motor must be regularly taken out of service to replace brushes and at less frequent interval to resurface the commutator.
- Brushless construction does not produce any sparking and eliminates the need for brush maintenance
- Since there are no field windings, there are no field copper losses resulting in improved efficiency.
- They are able to maintain a constant speed throughout the operation.
- Has lower rotor inertia when compared to Induction Motors because of the absence of a rotor cage. This allows for faster response

1.2. Problem of the statement

The demand for the use of permanent magnet synchronous motor drives has increased in many industrial applications. Because, the inherent advantages of these machines include high power density, low inertia, high speed capabilities, and high efficiency. Despite of the above advantage, the permanent magnet synchronous motor is a coupled nonlinear multivariable control structure which needs a complex nonlinear design. This makes the control performance of permanent magnet synchronous motor drives in highly sensitive to external load and system parameters variation [1][4]. Since this variation affect the performance of permanent magnet synchronous motor drive system. High performance speed control against parameter variations and load disturbances are required. Most of the time, vector controlled permanent magnet synchronous motor drive systems use a fixed gain proportional-integral (PI) controllers. However, the control performance of the permanent magnet synchronous motor drive is still influenced by unpredictable plant parameter variations, external load disturbances, and nonlinear dynamics of the plant or uncertainty of the system model, which influence performance of transient response such as: maximum overshoot, rise time, steady state error and settling time. Thus, the controller parameters must be tuned on-line continually. To overcome the above problem, many modern control theory (adaptive control techniques) such as model reference adaptive control (MRAC), Variable structure control (VSC), and self-tuning PI controllers have been developed. The design of all of the above controllers depends on the exact mathematical model of the drive system which is difficult to achieve practically.

Addition of artificial intelligence techniques like fuzzy logic control (FLC) can yield interesting solution in the field of speed control of drives. Fuzzy logic control does not depend on the mathematical model of the drive system.

In this thesis, self-tuning PI controller is proposed based on fuzzy logic approaches, which when included in the controller structure, ensures robustness against variations of some parameters of the system.

1.3. Objective of the thesis

General objectives: The general objective of the thesis is to design and simulate fuzzy logic based self-tuning PI controller for permanent magnet synchronous motor.

Specific objectives: The specific objectives are:

- To study the characteristics and mathimatical modeling of the Permanent magnet sysncronous motor drive system.
- To control the speed of Permanent magnet sysncronous motor by using self-tuning PI controller based on fuzzy logic approaches for the outer loop
- To design PI controller for controlling the inner loop.
- To simulate the system with the proposed controller using MATLAB/SIMULINK

- To precisely control the permanent magnet sysncronous motor to give the desired performance such as: Minimum overshoot,zero steady state and fast settling time.

1.4. Methodology

The methodology of the thesis involves a number of different tasks that are performed to lead towards completions of the work. In the first task, statement of the problem was described and the objective of the research was defined. This was followed by an extensive review of literature, publications and journals where all the theoretical information's regarding permanent magnet synchronous motor drives were gathered and comparison of previous similar work was studied. This was followed by studying characteristics and modeling of PM motor drives. Due to the non-linear dynamic model property of the permanent magnet synchronous motor drive, self-tuning PI controller was developed based on the fuzzy logic approach. Next, a brief description on the fuzzy control theory was presented. Then the rule base and membership function has been developed for tuning PI controller parameters based on fuzzy logic approach. Here after, by using proposed controller, the simulation has been carried out by using matlab/Simulink software package. The solver options, in the model configurations parameter was set to fixed step to fast the simulation time. The simulation studies are carried out for deferent controllers at different conditions to shows the advantages of the proposed controller compared to the conventional speed controller. The final stage is conclusion based on the research finding.

1.5. Scope and Limitation

The scope of this thesis is to design appropriate speed controller for PMSM based on fuzzy logic approaches. The motor parameter specification has been taken from matlab documentation.

1.6. Literature reviews

Speed control of the permanent magnet motor drives has been a topic of interest for last years. Different papers have been published reporting different controller design for such drives. Pillay and Krishnan, R. [5] in 1988, presented views on PM motor drives and classified them into two types. These are permanent magnet synchronous motor drives and brushless dc motor (BDCM) drives. The permanent magnet synchronous motor had a sinusoidal back emf(electromotive force) and required sinusoidal stator currents which produced constant torque while the BDCM had a trapezoidal back emf, required rectangular stator currents for producing constant torque. In 2008 Mutasim Nour, Omrane Bouketir & Ch'ng Eng Yong [6] presented a self-tuning of PI speed controller gains using fuzzy logic controller for permanent magnet synchronous motor as test bed to adapt the controller gains to speed changes, load disturbances and parameters variations. Hence, an online self-tuning scheme using fuzzy logic controller (FLC) is proposed in this paper. The performance of the developed controller is tested through a wide range of speeds as well as with load and parameters variations through simulation using MATLAB/SIMULINK. The simulation results show that the developed controller can well adapt to speed changes as well as sudden speed reduction besides fast recovery from load torque and parameters variation and these show improvement compared to conventional PI controller performance. In this Paper the controller scheme needs tuning of the input/output scaling factors which is additional job for the processor which increases the time of response of the system. In 2007 Yanpeng Dou, Zhang Ze [7] discussed design and realization of fuzzy self-tuning PID speed controller based on TMS320F2812 DSPs to apply fuzzy self-tuning PID controller to AC-speed adjustable system, and special attention is given on how to realize this controller using fuzzy search table method in C++ language based on TMS320F2812, which is the newest 32-bit fixed point DSP. And the results show that the controller improves both the dynamic characteristics and static characteristics of drive system. The paper only compares the speed response of the proposed controller and PID controller without any load torque and parameter variation. In 2007 Limei Wang, Mingxiu Tian and Yanping Gao [8] reviewed fuzzy self-adapting PID Control of permanent magnet synchronous motor servo system to adjust PID parameters adaptively on-line using fuzzy inference method for the varying state of the system. Because the traditional self-adapting control is very difficult to realize, which need on-line identification of the system parameters and the real-time change of control strategy. The proposed control scheme was verified by simulation using MATLAB and shows better performance than conventional PID controller in the permanent magnet synchronous motor drive system. This paper compares only the speed responses of the proposed controller and PID controller for step input reference speed without considering any other speed trajectories, load disturbances, and parameter variations. In 2006 Rajesh Kumar, R. A. Gupta, and Bhim Singh presented a critical analysis of intelligent tuned PID controllers for the permanent magnet synchronous motor drive so that PID control strategies, based on fuzzy logic, neural network and genetic algorithms are reviewed. Different tuning algorithms and their effect on dynamic performance of permanent magnet synchronous motor drive in the real time frame are illustrated in terms of starting and speed reversal time, steady state error in speed, overshoot, performance parameters and speed and torque ripple. Good agreements between different methods has been observed and presented. A fuzzy-logic-based speed control of vector controlled permanent magnet synchronous motor can also be designed that completely replaces the traditional PI controller. The fuzzy logic controller (FLC) is more robust and, hence, found to be a suitable replacement of the PI controller for the high performance drive systems. But this controller needs a highly experienced operator to design it and the control system is completely replaced which is still additional cost for the PMSM drive in industrial systems [9]. A composite control of fuzzy and PI or fuzzy and PID is also possible for permanent magnet synchronous motor (PMSM) drives to achieve fast dynamic response and minimum steady state error. A switching algorithm is created to switch between fuzzy control and PI(PID) control and results showed that the controller had a quick dynamic response, and eliminated basic steady-state error, dynamic characteristics and stability. Using this composite control of fuzzy and PI for PMSM drive in Industrial processes cause changing of the overall system topology which increases the cost of optimization of the control system[10][11]. A number of other papers also present advantages of fuzzy-logic based controller techniques over PI and verified to implement sensored and sensorless control of permanent magnet synchronous motor in normal and field weakening region.

1.7. Thesis organization

The thesis is organized into five chapters.

Chapter one presents the introduction part, objective of the studies, method to be followed, literature review, organizations of the studies are presented.

The second chapter describes the theory and operation principle of permanent magnet synchronous motor. Different permanent magnet synchronous motor construction details, difference with synchronous motors, and its different applications, dynamic modeling of the permanent magnet synchronous motor drive with field oriented control based on co-ordinate transformation.

The third chapter deals with the control of permanent magnet synchronous motor drives. It has two main parts. The first one deals with the drive system component such as: perma nent magnet synchronous motor, inverter, pulse width modulations, space vector pulse width modulation and implementations of SVPWM are discussed while the second part deal with controller design. Under this, speed controller, self-tuning PI controller based on fuzzy logic approach and current controller (PI), for controlling i_q *and* i_d are discussed.

The fourth chapter discusses simulation of the drive system using matlab/Simulink, where as the fifth chapter contains the conclusions from the work done in this thesis, recommends and suggestions for future work.

CHAPTER TWO

General characteristics of PMSM and its principle of operation

2.1. Introduction

Permanent magnet machines are electromechanical devices using magnets to produce a magnetic flux in the air gap. There are two major classifications of ac motors. Thus are: Induction motors which is electrically connected to power source through electromagnetic coupling, the rotor and the stator fields interact, creating rotation without any other power source. The second one is synchronous motors that have fixed stator windings that are electrically connected to the ac supply with a separate source of excitation connected to field windings when the motor is operating at synchronous speed. Among the synchronous motor types the permanent magnet synchronous motor (PMSM) is one possible design of three phase synchronous machines. The stator of a permanent magnet synchronous motor has conventional three phase windings. In the rotor, PM materials have the same function of the field winding in a conventional synchronous machine. The use of a PM to generate substantial air gap magnetic flux makes it possible to design highly efficient PM motors [1][2][12].

Synchronous motors rotate at the speed of stator revolving field, which is called synchronous speed. The synchronous speed is determined by the frequency of the stator supply and the number of stator pole pairs P.

$$N_s = \frac{120 * f_s}{p} \quad \text{(2.1)}$$

where

N_s= synchronous speed in (rpm)

f_s=frequency (Hz)

p=number of poles

For commonly used alternator, for 50Hz and with 4 poles, speed of 1,500rpm is calculated

unlike the induction motor, the rotor also has p-pole pairs, excited by a separate DC or permanent magnet (PM) source. The stator of a three-phase synchronous motor normally has a sine distributed three-phase winding. When excited with a three phase balanced supply, a rotating magnetic field develops.

2.2. Over view of permanent magnet synchronous machine.

For the permanent magnet synchronous motor, the three phase stator winding has sinusoidal winding distributed. Therefore the generated voltages are also sinusoidal. The key reason for the development of the permanent magnet synchronous motor was to remove the foregoing disadvantages of the synchronous machine (SM) by replacing its field coil, dc power supply and slip rings with a permanent magnet. The permanent magnet synchronous motor therefore, has a sinusoidal induced emf and requires sinusoidal currents to produce constant torque just like the SM.

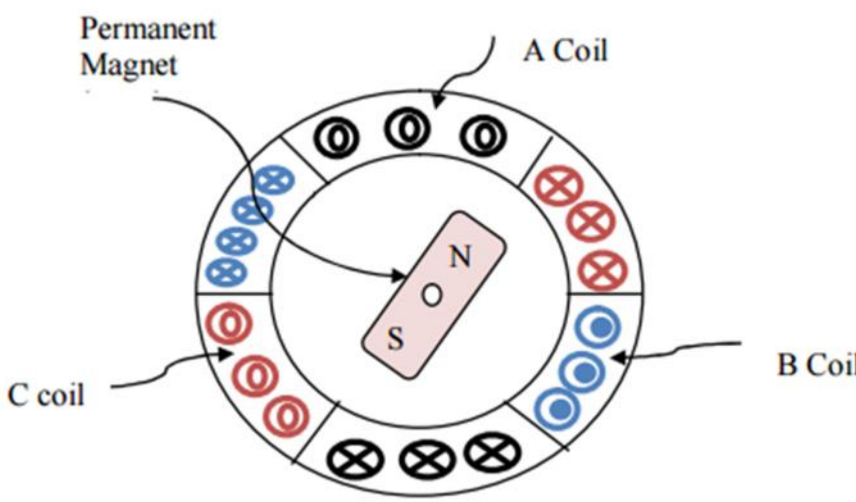

Figure 2.1: Cross sectional view of PMSM [21]

2.2.1. Permanent magnet materials

The properties of the permanent magnet material will affect directly the performance of the motor and proper knowledge is required for the selection of the materials and for understanding PM motors. Magnets made from steel were easily magnetized. However, they could hold very low energy and it was easy to demagnetize. In recent years other magnet materials such as Aluminum Nickel and Cobalt alloys (ALNICO)Strontium Ferrite or Barium Ferrite (Ferrite) Samarium Cobalt (First generation rare earth magnet)(SmCo) and Neodymium Iron-Boron (Second generation rare earth magnet) (NdFeB) have been developed and used for making permanent magnets. The rare earth magnets are categorized into two classes: Samarium Cobalt (SmCo) magnets and Neodymium Iron Boride (NdFeB) magnets. SmCo magnets have higher flux density levels but they are very expensive. NdFeB magnets are the most common rare earth magnets used in motors these days[13].

2.2.2. Classification of permanent magnet synchronous machines

Based on the directions of the field flux, permanent magnet synchronous motor is classified as:

- Radial field: flux direction is along the radius of the machine.
- Axial field machines: the flux direction is parallel to the rotor shaft.

The radial field machines are common; axial field machines are coming into prominence in a small number of applications because of their higher power density and acceleration. But these are very desirable features in high performance applications. Here radial field permanent magnet synchronous machines are discussed.

2.2.3. Radial field permanent magnet synchronous machines

There are basically three types of permanent magnet synchronous machines.

- Surface mounted permanent magnet synchronous machines(SPMSM)
- Inset permanent magnet synchronous machines
- Interior permanent magnet synchronous machines(IPMSM)

2.2.4. Surface mounted permanent magnet synchronous machine

Based on the placement, they are called surface permanent magnet motor (SPM) or interior permanent magnet (IPM) synchronous motor. For surface PM, each of the PM is mounted on the surface of the outer periphery of rotor laminations. This arrangement provide the highest air gap flux density as it directly faces the air gap without the interruption of any other medium such as part of rotor laminations. These motors are considered to have small saliency, thus having practically equal inductances in both quadrature and direct axes. For a surface permanent magnet motor, Ld = Lq. Figure 2.2 shows the placement of the magnet

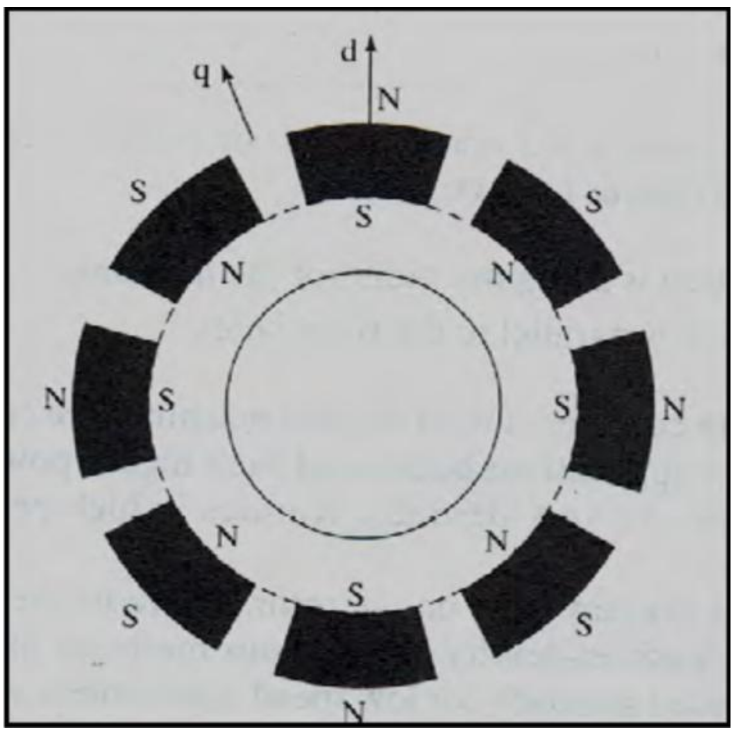

Figure 2.2: Surface mounted permanent magnet synchronous motor [IJCSE]

Machines with this arrangement are known as surface mounted permanent magnet synchronous machines. Surface mounted permanent magnet synchronous machines with radial field versions are used for low speed applications.

2.2.5. Interior permanent magnet synchronous machine

For Interior permanent magnet, each permanent magnet is mounted inside the rotor. The interior PM rotor construction is mechanically robust and therefore suited for high-speed applications. The manufacturing of this arrangement is more complex than the surface mount. By designing a rotor magnetic circuit such that the inductance varies as a function of rotor angle, the reluctance torque can be produced in addition to the mutual reaction torque of synchronous motors. These motors are considered to have saliency with q axis inductance (Lq) greater than the d axis inductance (Ld) (Lq > Ld).

In this thesis SPM radial flux machine has been chosen due to the following reasons:

- The SPM drive without reluctance torque is simpler to analyze and design than an IPM drive.
- The topology of a radial flux machine with classical winding and lamination has been chosen because of the well-known and established technology

Regardless of manner of mounting the permanent magnets the basic principle of operation is same. An important consequence of method of mounting the rotor magnets is the difference between the direct and quadrature axes inductance values. It is explained as follows. The rotor magnetic axis is called the direct axis and the principal path of flux flow is through the magnets. The permeability of high flux density permanent magnets is almost that of air. This results in magnetic thickness becoming an extension of air gap by that amount. The stator inductance when direct axis or magnets are aligned with the stator winding is known as direct axis inductance. By rotating the magnets from aligned position by 90^0, the stator flux sees inter polar area of rotor containing only the iron path and inductance measured in this position is referred to as quadrature axis inductance. The direct axis reluctance is greater than quadrature axis reluctance, because the effective air gap of direct axis is multiple times that of air gap seen by the quadrature axis.

Therefore $L_q > L_d$ for IPMSM and $L_d = L_q$ for surface mounted permanent magnet synchronous motor.

where

Ld is the inductance along the magnetic axis.

Lq is the inductance along the axis in quadrature with the magnetic axis.

Now a brief discussion is given about various types of permanent magnet synchronous machines.

PM motors are also classified on the basis of the flux density distribution and the shape of current excitation as permanent magnet synchronous motor and BLDC. The permanent magnet synchronous motor has a sinusoidal-shaped back emf and is designed to develop sinusoidal back emf waveforms.

Permanent magnet synchronous motor have the following features:

1. Sinusoidal distribution of magnet flux in the air gap

2. Sinusoidal current waveforms

3. Sinusoidal distribution of stator conductors.

BLDC has a trapezoidal-shaped back emf and is designed to develop trapezoidal back emf waveforms. They have the following:

1. Rectangular distribution of magnet flux in the air gap

2. Rectangular current waveform

3. Concentrated stator windings

Figure 2.3 show the classification of electrical machines.

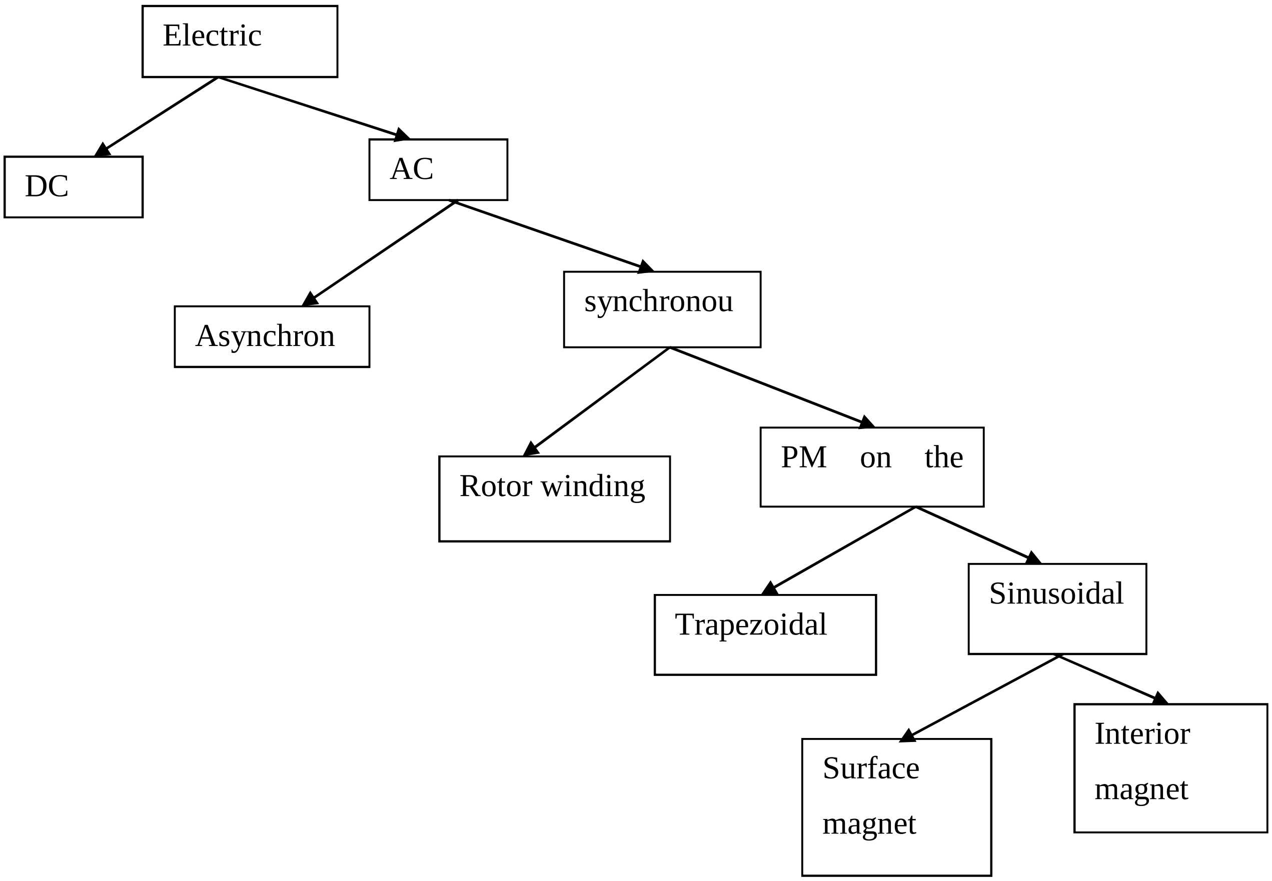

Figure2.3: Summary of classification of machines

2.2.6. Comparison of permanent magnet synchronous motor and BLDC motors

There are primarily two main types of brushless motors, namely- The brushless DC motor (BLDC), and The permanent magnet synchronous motor (PMSM). They both share the same basic construction, consisting of permanent magnets on the rotor and windings on the stator. However, one principal difference is that the coils in the stator are evenly wound in a BLDC motor, and in the case of permanent magnet synchronous motor they are wound in a symmetrical fashion. As a result, the back-EMF generated in a BLDC motor is trapezoidal in nature as shown in Figure 2.4, and Sinusoidal in nature in permanent magnet synchronous motor. Consequently, the permanent magnet synchronous motor are more efficient and produce less noise and torque ripple. However, the inherent advantages of a permanent magnet synchronous motor do necessitate more complex control algorithms.

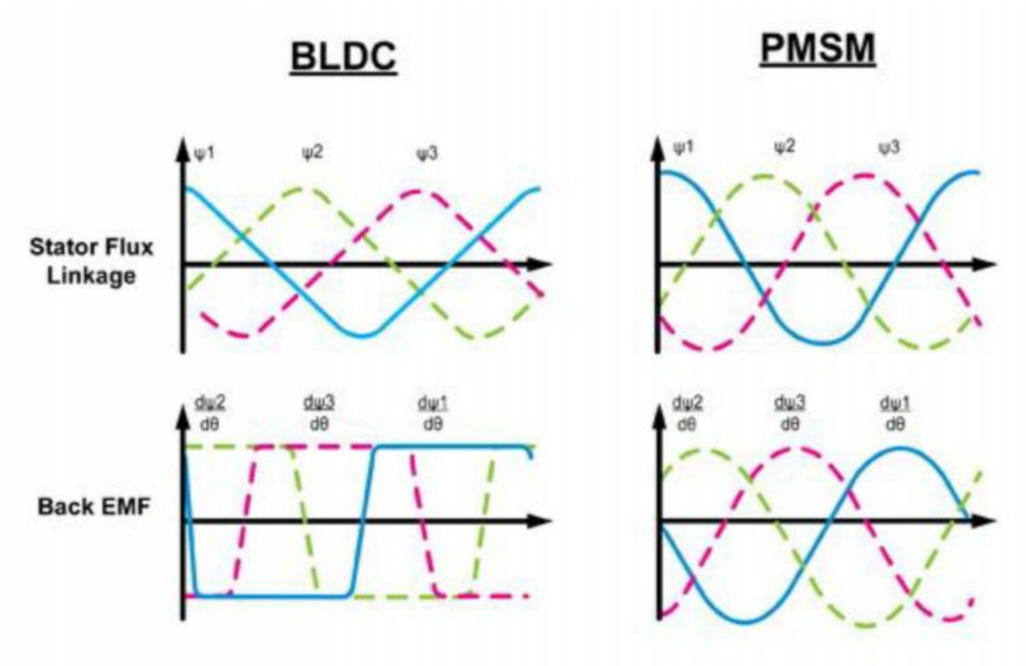

Figure 2.4: Comparison of BLDC and PMSM [36]

2.2.7. Control Strategy

For any position of the rotor, there is an optimal direction for the net stator field which maximizes the torque produced. If the net stator field and the rotor field are in the same direction then no torque will be produced. This is because the forces produced by the interaction of the fields act in line with the axis of rotation of the rotor and only leads to compression of the motor bearings, rather than rotation. If the stator field is perpendicular to the rotor field, then maximum torque is produced [14].

Any stator field can be represented as a resultant of two vector components, the orthogonal (quadrature) component which is responsible for producing torque, and the parallel (direct) component which produces undesired heat and compression forces. Hence, an ideal drive should aim to minimize the parallel component while maximizing the quadrature component.

The stator field is produced by current flow in three equally spaced windings that are mechanically located 120^0 apart. Hence, they each produce a field vector component that is oriented 120^0 from each other. The mathematical models for permanent magnet motors are in terms of winding currents rather than stator magnetic field since they are easier to measure. Current space vectors are used to model the stator fields in terms of winding currents. The current space vector for a given winding has a magnitude proportional to the current flowing through the winding and a direction of the field produced by that winding. This allows us to represent the total stator field as a current space vector that is the vector sum of three current space vector components of the windings. To put this simply, the resultant current space vector is the current that would need to flow in a single fictitious winding in order to produce the same stator field direction and magnitude as the combination of three real currents through real stator windings.

Just like the stator field, the current space vector too can be broken down into quadrature and direct axis components. The orthogonal (quadrature) current component produces a field at right angles to the rotor magnet and therefore results in torque, while the parallel (direct) current component produces a field that is aligned with the rotor magnet and produces no torque. Hence, the aim of a control algorithm is to reduce the direct current components so as to minimize losses. The current space vector is illustrated in Figure 2.5 where the three windings are named A,B& C.

To produce constant smooth torque, the stator current space vector should ideally be constant in magnitude and should turn with the rotor so as to always be in the quadrature direction, irrespective of rotor angle and speed. To achieve this, various control algorithms have been experimented and developed with varying degrees of success.

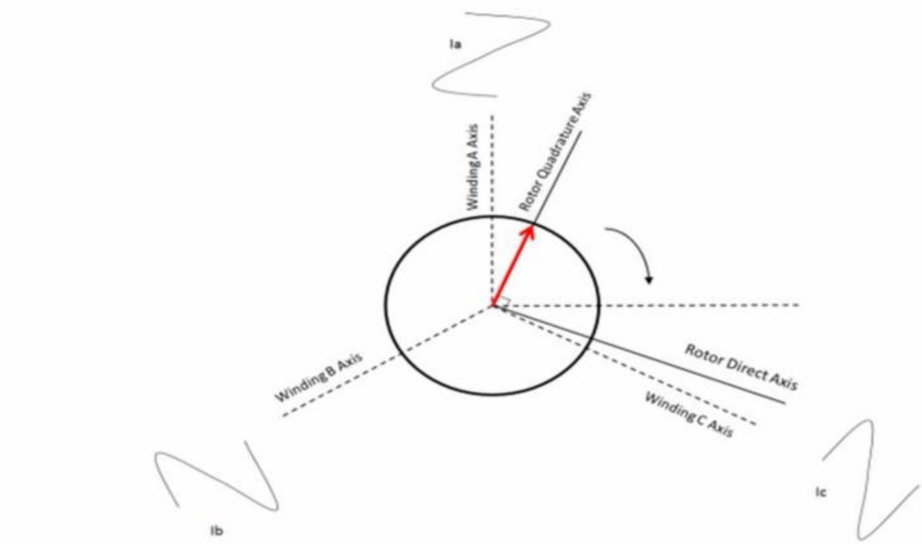

Figure 2.5: Ideal current space vector[36]

2.2.8. Field oriented control

The most prevalent control strategies for brushless motors are trapezoidal control, Sinusoidal control and field oriented control. Trapezoidal control is relatively simple and offers smooth operation at high speed but causes torque ripple at low speeds. Sinusoidal Control eliminates torque ripple and provides efficient operation at low speeds but the limitations of a PI controller make it unsuitable for high speed applications. Field oriented control (FOC) combines the best aspects of the trapezoidal and sinusoidal control methods, offering smooth and efficient operation with fast dynamic response at both low and high speeds. Due to this reason, field oriented control is opted to implement in the drive system.

The main drawback of sinusoidal control arises due to the fact that this control scheme tries to control the motor currents whose magnitude and direction varies with time. As the speed and frequency increase, the PI controllers are incapable of handling the operation due to their limited bandwidth. This problem can be solved by representing and controlling the current space vector in the two axis d-q frame of reference.

This control scheme revolves around clarke and park transformations, and their inverse. By applying these transformations, we can transform the 3-Φ currents of the stator into the rotating frame of the rotor. By using clarke transformation, three-phase quantities are translated from the three-phase reference frame to the two-axis orthogonal stationary reference. However, the quantities are still in a stationary reference frame while the rotor reference frame is constantly rotating. Park's transformation converts these quantities into an orthogonal reference frame consisting of the direct and quadrature axis. The three reference frames are illustrated in Figure 2.6.

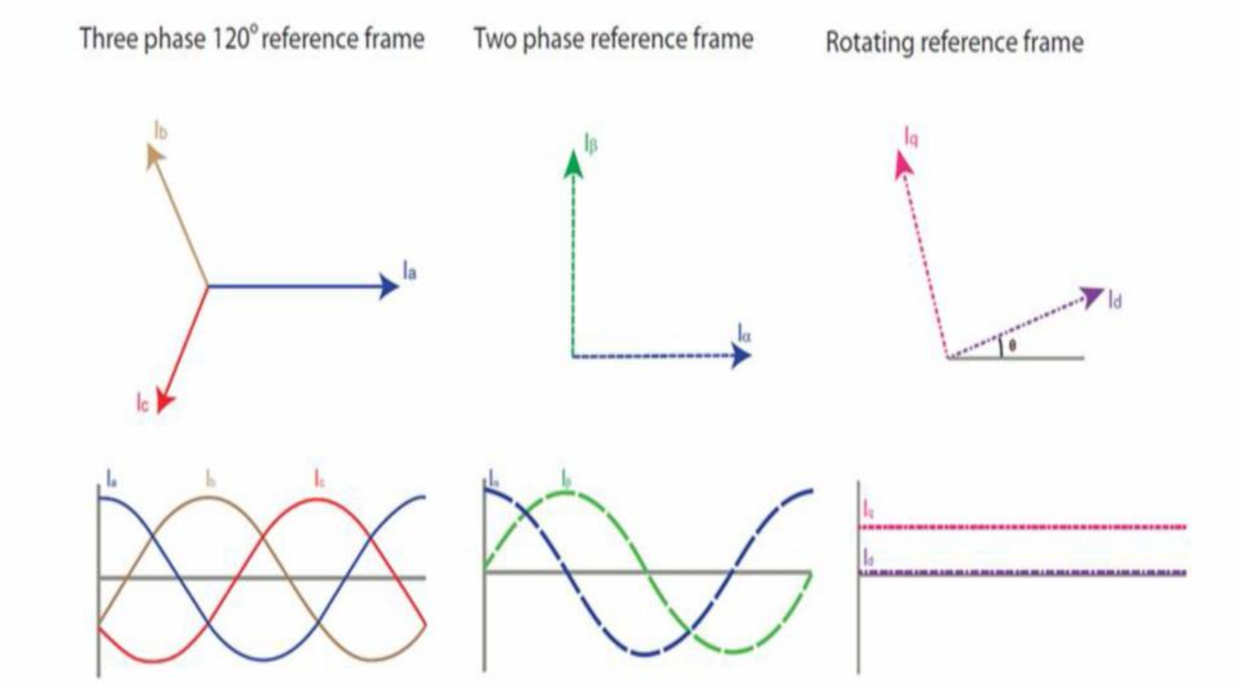

Figure 2.6: The three reference frames[36]

As mentioned earlier, the direct axis component produces useless torque while the quadrature axis component produces the torque responsible for rotation. In an ideal case, the current space vector in the d-q frame is fixed in magnitude and direction (quadrature) with respect to the rotor, irrespective of rotation. Since the current space vector in the d-q frame of reference is static, the PI controllers now have to operate on DC quantities rather than sinusoidal signals, thus greatly simplifying the control structure. This isolates the controllers from the time variant winding currents and voltages, therefore eliminating the limitation of controller frequency response and phase shift on motor torque and speed[15]

By isolating the PI controllers from the time varying currents and voltages, the FOC algorithm is able to offer numerous advantages, such as:

- High Efficiency
- Smooth operation at low and high speeds resulting in a wide range of speed
- Fast dynamic response and good transient and steady state performance
- Transformation of a complex and coupled AC model into a simple linear system.

2.2.9. Application area of permanent magnet synchronous motor

Permanent magnet synchronous motors have many applications. Among this;

Robotics and factory automation (servo drives)

- Pick and place robot (motion control)
- Positioning tablets
- Automatic guided vehicles

Appliance

- Washers
- Blowers
- Compressors
- Heating
- Ventilations and air conditioning

Computer and office equipment

- Copier and microfilm
- Printer
- Plotter
- Tape driver.

2.3. Mathematical model of permanent magnet synchronous motor

Traditionally per-phase equivalent circuits have been widely used in steady-state analysis and design of AC machines. It is not appropriate to predict the dynamic performance of the motor. In order to understand and analyze vector control of AC motor drives, a dynamic model is necessary. As the application of AC machines has continued to increase over this century, new techniques have been developed to aid in their analysis. The significant breakthrough in the analysis of three-phase machines was the development of the reference frame theory. Using these techniques, it is possible to transform the machine model to another reference frame. By careful choice of the reference frame, it is possible to simplify vastly the complexity of the mathematical machine model. While these techniques were initially developed for the analysis and simulation of AC machines, they are now important tools in the digital control of such machines.

From a construction point of view, the permanent magnet synchronous motor is very similar to the conventional synchronous motor and as a result, its dynamic model can be derived from the well-known equations of the classical synchronous machine. This model can describe the machine in a real abc stationary frame or in a dq rotating frame. Typically, by using the abc reference frame, the obtained equations are non-linear, meaning that they are relatively complex, and the variables are time-variant. If the machine is described using a dq rotor reference frame, the obtained equations are much more simple, allowing to accelerate the numerical calculations involved in the computational simulations. Furthermore, under steady-state operation the obtained variables do not vary in time, making easier additional calculations and simplifying the control system implementation.

According to this, the dq model in the rotor reference frame is the most common and widely used model for proper simulation of the system. This model assumes that the saturation is neglected, the back-EMF is sinusoidal, that there is no cage in the rotor and that the eddy currents and hysteresis losses are negligible. Taking into account these restrictions, the resulting equations become very simple, allowing predicting the machine behavior with a reasonable accuracy.

2.3.1. Dynamic model

As said before, for describing and analyzing a permanent magnet synchronous motor, the most convenient way is to consider a synchronous rotating reference frame fixed to the rotor. The transformation of the state variables (voltages, currents and fluxes) from the abc stationary frame into the rotating dq0 coordinates, is accomplished by using the amplitude-invariant transformation matrix defined as:

$$\begin{bmatrix} x_d \\ x_q \\ x_o \end{bmatrix} = \begin{bmatrix} \cos(\theta) & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ -\sin(\theta) & -\sin\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (2.2)$$

where θ is the angle between the stator phase *a* and the rotor flux. On the contrary, the state variables in the *abc* stationary frame can be obtained from the rotating *dq*0 components using the inverse amplitude-invariant transformation matrix:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \begin{bmatrix} \cos(\theta) & -\sin(\theta) & \frac{1}{2} \\ \cos\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta - \frac{2\pi}{3}\right) & \frac{1}{2} \\ \cos\left(\theta + \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) & \frac{1}{2} \end{bmatrix} \begin{bmatrix} x_d \\ x_q \\ x_o \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (2.3)$$

Detailed modeling of PM motor drive system is required for proper simulation of the system. The d-q model has been developed on rotor reference frame as shown in Figure 2.7. At any time t, the rotating rotor d-axis makes an angle θ with the fixed stator phase axis. Figure 2.7 illustrates a conceptual cross-sectional view of a 3-phase, permanent magnet synchronous motor along with two reference frames. The stator reference axis for the a-phase is chosen to the direction of maximum mmf when a positive a-phase current is supplied at its maximum level. Reference axis for b- and c- stator frame are chosen

120° and 240° (electrical angle) ahead of the a-axis, respectively. Following the convention of choosing the rotor reference frame, the direction of permanent magnet flux is chosen as the d-axis, while the q-axis is 90 degrees ahead of the d-axis. The angle of the rotor d-axis with respect to the stator a-axis is θ. As the machine turns, the d-q reference frame is rotating at a speed of $\omega = \frac{d\theta}{dt}$, while the stator a-b-c axes are fixed in space. The choice of this rotating frame greatly símplífies the dynamic equations of the model.

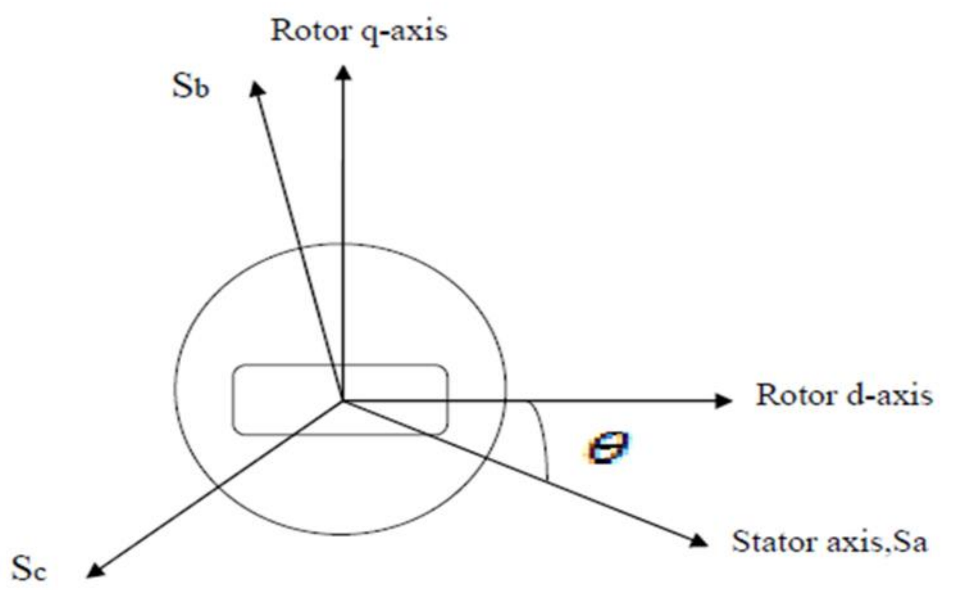

Figure 2.7: Three - Phase permanent magnet synchronous motor with one pole pair permanent magnet[35]

2.3.2. Coordinate transformation

To make a transformation, there are three coordinates whose relations are shown by Figure 2.8, that are a-b-c,α-β and d-q. a-b-c is three phase coordinate, α-β is stationary coordinate and d-q is rotating coordinate which rotates at ω speed

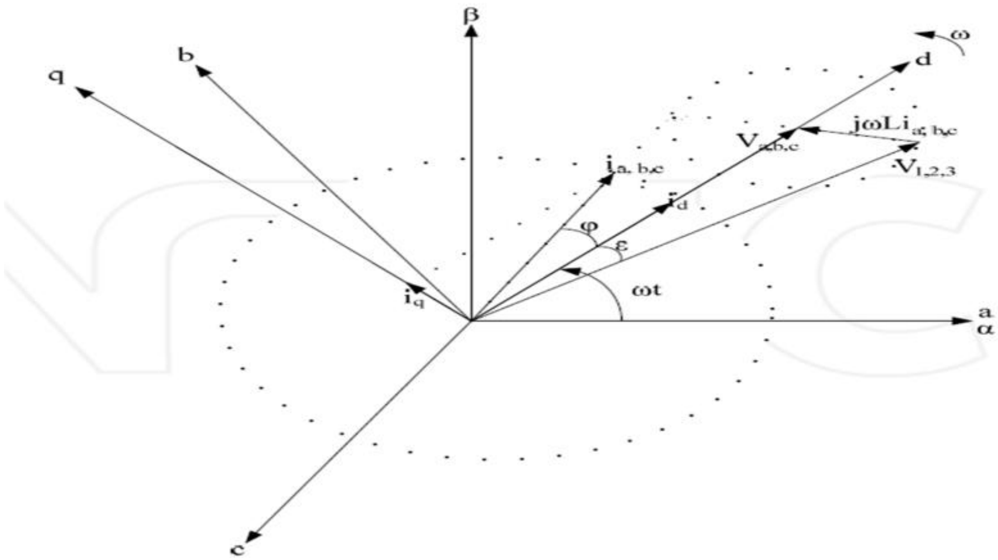

Figure 2.8: Coordinates diagram of a-b-c, α-β and d-q[1]

The d-q transformation is a transformation of coordinates from the three-phase stationary coordinate system to the d-q rotating coordinate system. A representation of a vector in any n-dimensional space is accomplished through the product of a transpose n-dimensional vector (base) of coordinate units and a vector representation of the vector, whose elements are corresponding projections on each coordinate axis, normalized by their unit values[1][15][16].

Assuming a balanced three-phase system, a three-phase vector representation transforms to d-q vector representation (zero-axis component is 0) through the transformation matrix T as given in (2.4).

$$T = \frac{2}{3}\begin{bmatrix} \cos(\omega t) & \cos\left(\omega t - \frac{2\pi}{3}\right) & \cos\left(\omega t + \frac{2\pi}{3}\right) \\ -\sin(\omega t) & -\sin\left(\omega t - \frac{2\pi}{3}\right) & -\sin\left(\omega t + \frac{2\pi}{3}\right) \end{bmatrix} \qquad (2.4)$$

where

w = the fundamental frequency of three-phase variables.

T=the transformations matrix

The transformation from X_{abc}(three-phase coordinates) to X_{dq}(d-q rotating coordinates) is called park transformation, which is obtained through the multiplication of the vector X_{abc} by the matrix T, as in (2.5).

$$x_{dq} = T * x_{abc} \quad \ldots \ldots (2.5)$$

The inverse transformation matrix (from d-q to a-b-c) is defined as in (2.6)

$$T^{-1} = \begin{bmatrix} \cos(\omega t) & \sin(\omega t) \\ \cos\left(\omega t - \frac{2\pi}{3}\right) & \sin\left(\omega t - \frac{2\pi}{3}\right) \\ \cos\left(\omega t + \frac{2\pi}{3}\right) & \sin\left(\omega t + \frac{2\pi}{3}\right) \end{bmatrix} \quad \ldots \ldots (2.6)$$

The inverse transformation is calculated as in (2.7).

$$x_{abc} = T^{-1} * x_{dq} \quad \ldots \ldots (2.7)$$

These is all about transformations used to transform from three phase stationary (stator) reference frame (X_{abc}) to two phase rotating (rotor) reference frame X_{dq} for convenience operations.

where: X_{abc} and X_{dq} are voltage, or current

2.3.3. Mathematical description of permanent magnet synchronous motor

The electrical dynamic equation in terms of phase variables can be written as follows in the stator phase quantities.

$$V_a = R_s i_a + \frac{d\Psi_a}{dt} \quad \ldots \ldots (2.8)$$

$$V_b = V_b i_b + \frac{d\Psi_b}{dt} \quad \ldots \ldots (2.9)$$

$$V_c = V_c i_c + \frac{d\Psi_c}{dt} \quad \ldots \ldots (2.10)$$

For the model, input power P_i can be given by

$$p_i = v_a i_a + v_b i_b + v_b i_b \quad \ldots \ldots (2.11)$$

The mathematical models in phase quantities given above are not usually convenient for control. It is usually convenient to transform the phase quantities in the stationary reference frame to rotating reference frame quantities, which is discussed above.

2.3.4. Voltage equation

The two axis voltage equations for the stator winding which are of surface permanent magnet synchronous motor where Ld and Lq have the same value) are given by:

$$V_q = R_S i_q + \frac{d}{dt}(\Psi_q) + \omega_r \Psi_d \dots \dots (2.12a)$$

$$V_d = R_S i_d + \frac{d}{dt}(\Psi_d) - \omega_r \Psi_q \dots \dots (2.12b)$$

where

$$\Psi_d = L_d i_d + \Psi_m \dots \dots (2.13a)$$

$$\Psi_q = L_q i_q \dots \dots (2.13b)$$

Substituting equations (2.13a and 2.13b) in to (2.12a and 2.12b) the dynamic equation gives a more convenient equation as in (2.14a and 2.14b)

$$V_q = \left(R_s + L_q \frac{d}{dt}\right) i_q + \omega_r L_d i_d + \omega_r \Psi_m \dots \dots (2.14a)$$

$$V_d = \left(R_s + L_d \frac{d}{dt}\right) i_d - \omega_r L_q i_q \dots \dots (2.14b)$$

In the matrix notation equations (15) can be written as:

$$\begin{bmatrix} V_q \\ V_d \end{bmatrix} = \begin{bmatrix} R_s & L_q \frac{d}{dt} \\ R_s & L_d \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_q \\ i_d \end{bmatrix} + \begin{bmatrix} \omega_r L_d i_d + \omega_r \Psi_m \\ -\omega_r L_q i_q \end{bmatrix} \dots \dots (2.15)$$

where

V_d and V_q are the stator voltages in dq-axis;

i_d and i_q = the stator currents in dq-axis;

Ψ_d and Ψ_q= the stator flux linkages in dq-axis;

Ψ_m= the permanent-magnet flux linkage;

T_e= the electromagnetic torque;

W_r = the angular velocity of rotor;

R_s=the stator resistance;

L_s=the stator inductance;

P = the number of poles;

Equivalent circuit of permanent magnet synchronous motor

The equivalent circuit of the permanent magnet synchronous motor can be derived for the stator q-axis and d-axis coordinates. During steady state operation, the *d-q* axis currents are constant quantities. From the dynamic equation (2.15), the equivalent circuit can be reduced to the steady state circuit shown in Figure 2.9 and Figure 2.10 respectively.

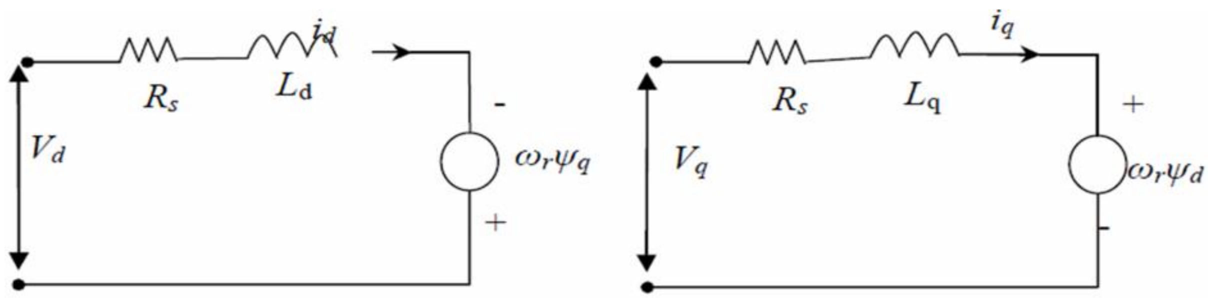

Figure 2.9: PMSM Dynamic stator q-axis and d-axis equivalent circuit

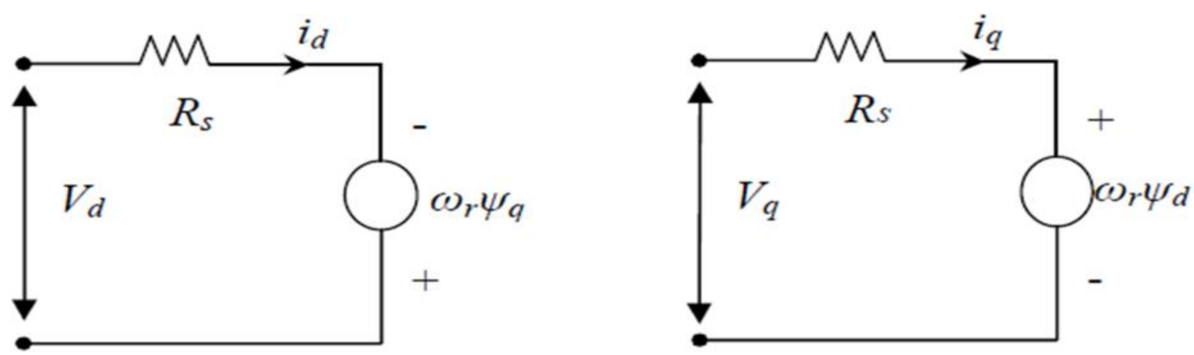

Figure 2.10: PMSM equivalent circuits from steady state equations

2.3.5. Power equivalence

The power input to the three-phase machine has to be equal to the power input of the two Phase machine to have meaningful interpretation in the modeling, analysis, and simulation

The three-phase instantaneous power input is given by:

$$P_{in} = [V_a \quad V_b \quad V_c] \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \qquad (2.16)$$

$$P_{in} = V_a i_a + V_b i_b + V_c i_c \qquad (2.17)$$

Using equation (2.6) the inverse transformation from two phase rotating reference frame to three phase stationary reference frame for voltage and current components is given as equation (2.18a and 2.18b).

$$V_{abc} = [T_{abc}]^{-1} V_{qd} \qquad (2.18a)$$

$$i_{abc} = [T_{abc}]^{-1} i_{qd} \qquad (2.18b)$$

By substituting equation (2.18a and 2.18b) into (2.17) results in (2.19)

$$P_{in} = \left([T_{abc}]^{-1} V_{qd}\right)^t [T_{abc}]^{-1} i_{qd} \qquad (2.19)$$

$$P_{in} = \frac{3}{2} \; V_q i_q + V_d i_d] \qquad (2.20)$$

Input power remains constant for all reference frames

2.3.6. Electromagnetic torque equation

The electromagnetic torque is the most important output variable that determines the mechanical dynamics of the machine such as the rotor position and speed. Therefore, its importance cannot be overstated in all the simulation studies. It is derived from the machine matrix equation by looking at the input power and its various components such as resistive losses, mechanical power, and the rate of change of stored magnetic energy. Elementary reasoning leads to the fact that there cannot be a power component due to the introduction of reference frames. Similarly, the rate of change of stored magnetic energy could only be zero in steady state. Hence, the output power is the difference between the input power and the resistive losses in a steady state.

Note that dynamically, the rate of change of stored magnetic energy need not be zero. Based on these observations, the derivation of the electromagnetic torque is made as follows [1] [17] [18].

Substituting equation (2.11) to the power equation (2.19) result

$$P_{in} = \frac{3}{2}\left[\left(R_s i_d^2 + R_s i_q^2\right) + \left(i_d \frac{d}{dt}\Psi_d + i_q \frac{d}{dt}\Psi_q\right) + \omega_r\left(\Psi_d i_q - \Psi_q i_d\right)\right] \ldots\ldots\ldots (2.20)$$

The first term of the above equation is the power loss in the conductors, the second term is the time rate of change of stored energy in the magnetic fields and the third term is the energy conversion from electrical to mechanical energy. The effective torque can be derived from the third term of the power equation and written as:

$$P_m = \omega_n T_{em} = \frac{3}{2}\omega_r\left(\Psi_d i_q - \Psi_q i_d\right) \ldots\ldots\ldots (2.21)$$

The relationship between mechanical speed and electrical speed is given by:

$$\omega_r = \omega_n P \ldots\ldots\ldots (2.22)$$

where

P is number of pole pairs

Substituting equation (2.22) into (2.21) result in

$$T_{em} = \frac{3}{2}P\left(\Psi_d i_q - \Psi_q i_d\right) \ldots\ldots\ldots (2.23)$$

By substituting the equation (2.13) in to equation (2.23) the torque equation can also be Expressed as(2.24) which is equivalent to the mechanical equation given in (3.25)

$$T_{em} = \frac{3}{2}P\left[\Psi_m i_q + \left(L_d - L_q\right) i_q i_d\right] \ldots\ldots\ldots (2.24)$$

$$T_{em} = T_l + J\frac{d}{dt}\omega_m + B\omega_m \ldots\ldots\ldots (2.25)$$

The first term in equation (2.24) is called “mutual reaction torque” occurring between i_q and the permanent magnet, while the second term corresponds to “reluctance torque” due to the difference in d-axis and q-axis reluctance (or inductance).

The two axis voltage equations for the stator winding are derived for interior permanent magnet synchronous motor (but are the same for surface permanent magnet synchronous motor where L_d and L_q have the same value). The motor used in this thesis is surface mounted permanent magnet synchronous motor which means that $L_d = L_q$, due to the same reluctance paths in rotor d-axis and q-axis, and therefore the “reluctance torque” is equal to zero, so the torque expression for surface permanent magnet synchronous motor is given as:

$$T_{em} = \frac{3}{2} P \Psi_m i_q \quad (2.26)$$

Since the number of pole pairs and the magnetic flux linkages are constant, then the torque is directly proportional to q-axis current i_q.

$$T_{em} = K_t i_q \quad (2.27)$$

where

$K_t = \frac{3}{2} P \Psi_m$ is called torque constant

2.3.7. Field oriented control of permanent magnet motor

The permanent magnet synchronous motor control is equivalent to that of the dc motor by a decoupling control known as field oriented control or vector control. The vector control separates the torque component of current and flux channels in the motor through its stator excitation. The vector control of the PM synchronous motor is derived from its dynamic model.

Detailed modeling of PM motor drive system is required for proper simulation of the system. The d-q model has been developed on rotor reference frame. At any time t, the rotating rotor d-axis makes an angle Θ_r with the fixed stator phase axis and rotating stator mmf makes an angle α with the rotor d-axis. Stator mmf rotates at the same speed as that of the rotor.

The model of permanent magnet synchronous motor has been developed on rotor reference frame using the following assumptions:

a) Saturation is neglected.
b) The induced EMF is sinusoidal.
c) Eddy currents and hysteresis losses are negligible.
d) There are no field current dynamic

Steady state torque characteristics of the permanent magnet synchronous motor highlights the relationship between the torque components and the torque angle α. Assume the set of balanced polyphone current as the input to the permanent magnet synchronous motor: To analyze the steady state torque characteristics of surface permanent magnet synchronous motor, consider a set of balanced three phase currents as input to the stator windings.

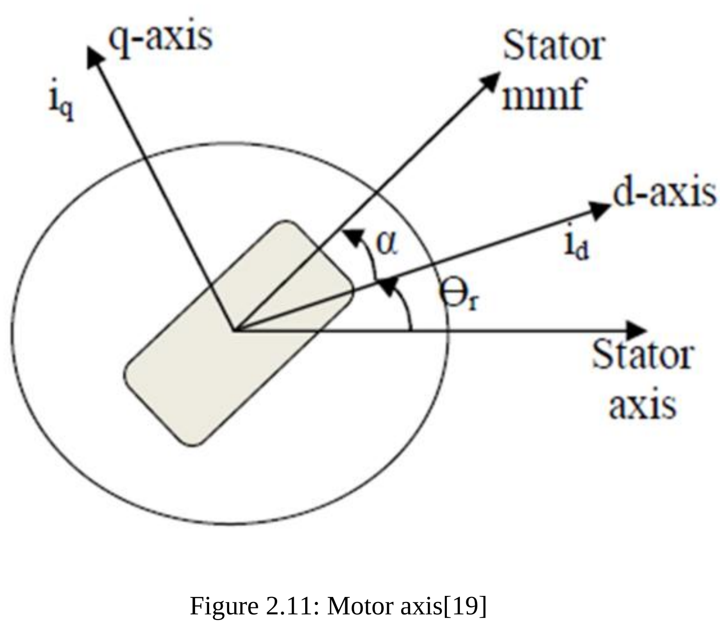

Figure 2.11: Motor axis[19]

$$i_a = I_m \sin(\omega_r t + \alpha)$$

$$i_b = I_m \sin\left(\omega_r t + \alpha - \frac{2\pi}{3}\right) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(2.28)$$

$$i_c = I_m \sin\left(\omega_r t + \alpha + \frac{2\pi}{3}\right)$$

In matrix form

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} \sin(\omega_r t + \alpha) \\ \sin\left(\omega_r t + \alpha - \frac{2\pi}{3}\right) \\ \sin \omega_r t + \alpha + \frac{2\pi}{3} \end{bmatrix} [I_m] \ldots\ldots\ldots\ldots\ldots\ldots (2.29)$$

By substituting equation (2.28) and (2.4) into equation (2.5), id and i_q is obtain in terms of I_m as follows

$$i_q = I_m \sin\alpha \ldots\ldots\ldots\ldots\ldots\ldots (2.30a)$$

$$i_d = I_m \cos\alpha \ldots\ldots\ldots\ldots\ldots\ldots (2.30b)$$

These is written as matrix form

$$\begin{bmatrix} i_q \\ i_d \end{bmatrix} = [I_m] \begin{bmatrix} \sin\alpha \\ \cos\alpha \end{bmatrix} \ldots\ldots\ldots\ldots\ldots\ldots (2.31)$$

Substituting equation (2.31) in to equation (2.27), the electromagnetic torque equation is obtained as follows as, L_d and L_q is assumed to be equal for surface permanent magnet synchronous motor.

$$T_{em} = \frac{3}{2} P [\Psi_m I_m \sin\alpha] \ldots\ldots\ldots\ldots\ldots\ldots (2.32)$$

where

α *is* termed the torque angle (the angle between the rotor field and stator mmf) as it directly influences the air gap torque. Im is the magnitude of the stator current phasor and ω_r is the rotor electrical speed.

The currents obtained are the stator currents that must be transformed to the rotor reference frame with the rotor speed ω_r using park transformation. The q and d axis currents are constants in the rotor reference frames since α is a constant for a given load torque. As these constants, they are similar to the armature and field currents in the separately excited dc machine. The q axis current is distinctly equivalent to the armature current of the dc machine; the d axis current is field current, but not in its entirety. For this reason the q axis current is called the torque producing component of the stator current and the d axis current is called the flux producing component of the stator current [19].

2.3.8. Constant torque operation

Constant torque control strategy is derived from field oriented control, where the maximum possible torque is desired at all times like the dc motor. This is performed by making the torque producing current iq equal to the supply current I_m. That results in selecting the α angle to be 90^o degrees according to equation (2.32). By making the i_d current equal to zero, the torque equation can be rewritten as:

$$T_e = \frac{3}{2}\left(\frac{p}{2}\right)\Psi_m i_q \quad (2.33)$$

Assuming that.

$$k_t = \frac{3}{2}\left(\frac{p}{2}\right)\Psi_m \quad (2.34)$$

The torque is given by

$$T_{em} = K_t i_q \quad (2.35)$$

where

P=pole pairs

K_t =torque constant and

Ψ_m = the permanent-magnet flux linkage

2.3.9. Voltage decoupling control

To achieve completely independent control of the direct-axis stator current i_d (field producing component) and the quadrature-axis stator current i_q(torque-producing component), it is necessary to cancel the effect of these coupling terms at the output of the current PI regulator [20]. The equations of the stator voltage components as shown in Equation (2.14) are coupled. The direct axis component V_d also depends on i_q, and the quadrature axis component V_q also depends on i_d The stator voltage components V_d and V_q cannot be considered as decoupled control variables for the rotor flux and electromagnetic torque. The stator currents i_d and i_q can only be independently controlled (decoupled control) if the stator voltage equations are decoupled, so these stator current components are indirectly controlled by controlling the terminal voltages of the synchronous motor.

The equations of the stator voltage components in the d, q reference frame can be reformulated and separated into two components:

- linear components and
- decoupling components.

The decoupling components are evaluated from the stator voltage equations. They eliminate cross-coupling for current control loops at a given motor operating point. Linear components $V_d^l$and V_q^l are set by the outputs of the current controllers

$$V_q = V_q^l + V_q^D \quad \ldots\ldots (2.36)$$

$$V_d = V_d^l + V_d^D \quad \ldots\ldots (2.37)$$

where

$$V_q^l = \left(R_s + L_q \frac{d}{dt}\right) i_q \quad \ldots\ldots (2.38)$$

$$V_d^l = \left(R_s + L_d \frac{d}{dt}\right) i_d \quad \ldots\ldots (2.39)$$

and

$$V_q^D = \omega_r L_d i_d + \omega_r \Psi_m \quad (2.40)$$

$$V_d^D = \omega_r L_q i_q \quad (2.41)$$

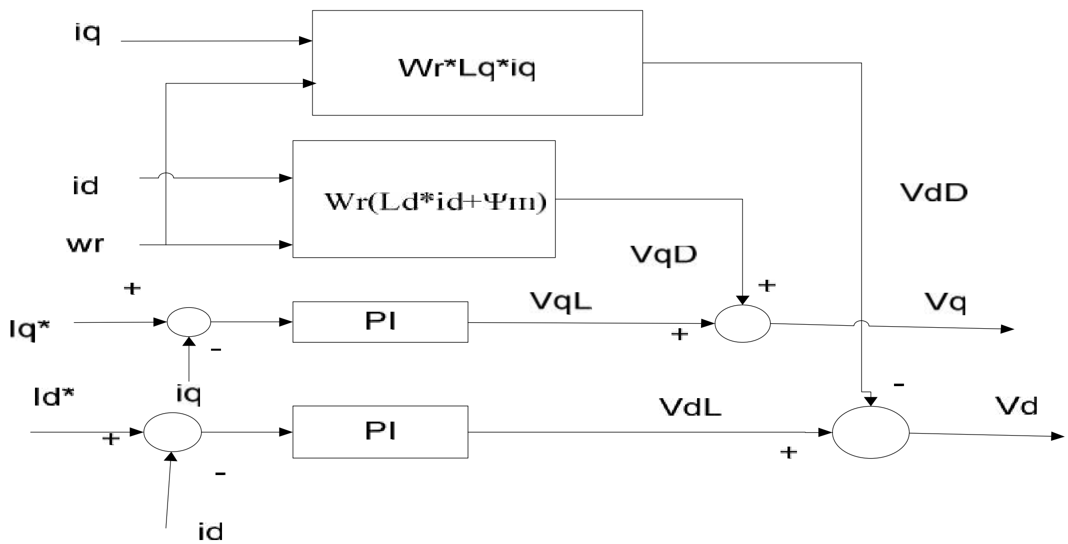

Figure 2.12: Voltage decoupling

CHAPTER THREE

Control of permanent magnet synchronous motor drives

3.1. Permanent magnet synchronous motor drive system

Rotor position information is very crucial for field oriented (vector) control. The coordinate transformation uses the value of the rotor position in order to handle the stator current vector projection in a rotating frame.

Sensor based control: In this control method, sensors are used to indicate the position of the rotor, for instance incremental encoder, absolute encoder, resolver etc. By using mechanical sensors, it is simple to measure speed and position for feedback to the controller. Moreover, controller design complexity is not that much. However, using sensor-based control increases volume, weight of the system, connection parts between motor and controller. Mechanical sensors have noise and they need space on the shaft of the motor for position measurement. It also increases the overall cost of the drive system.

Sensor-less Control: Position of the rotor is estimated using algorithms. It has got some advantages. As the first, it overcomes the drawbacks of sensor based control, it has better efficiency and lower cost without position transducer and it is reliable and faster response control method. On the contrary this control system has got some demerits. This includes sophisticated design (e.g. Kalman filter), some methods fail at standstill (e.g. Back-emf method) and expensive for low cost applications and it will take time to develop.

This thesis is totally dedicated on sensor based control of permanent magnet synchronous motor because of advantage mentioned above and the simplicity when compare to sensor less control. Whether the control scheme is sensor based or sensor-less, position of the rotor is required. Control of permanent magnet synchronous motor can be categorized as sensor based and sensor-less control [21].

The conceptual flow work of drive system is the speed, and current command is input to the drive system. The controller implements feedback control based on mechanical sensors. The speed controller outputs command for the current controller. The current controller block converts its input command into commands for the power converter using SVPWM. The power converter block imposes the desired electrical signals onto the permanent magnet synchronous motor machine with the connected load.

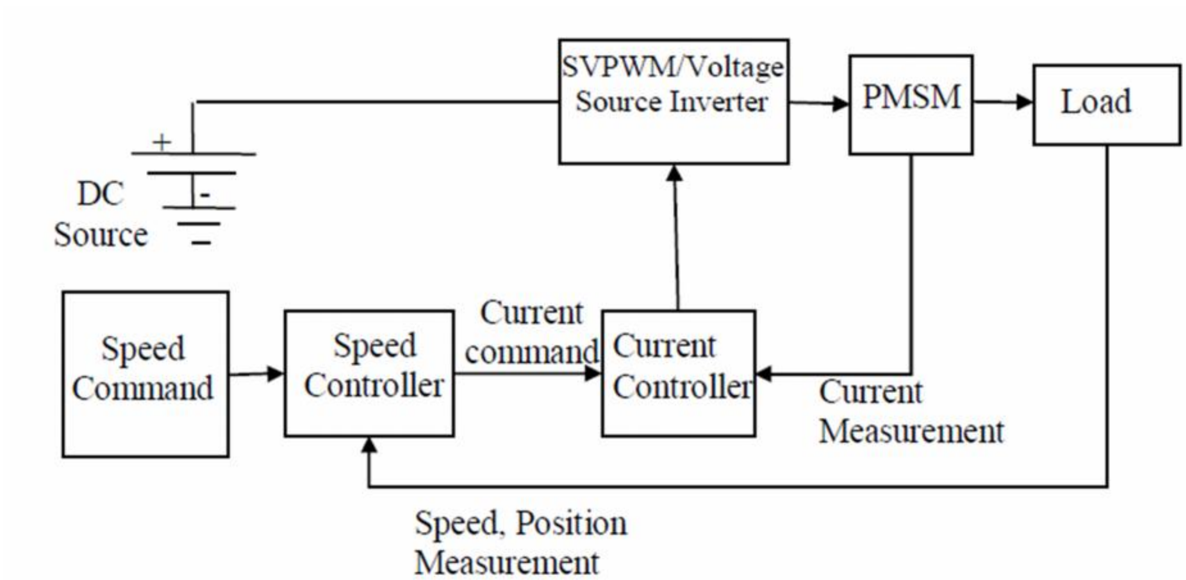

Figure 3.1: block diagram of permanent magnet synchronous motor drive system [21]

3.1.1. Inverter

Permanent magnet synchronous motor drives require variable voltage/current input at variable frequency to deliver variable speed operation. As the utility power source has constant frequency and voltage, it cannot be directly used in these machines. As a result there must be a power conversion process to obtain a variable voltage/current, variable frequency power supply from a fixed frequency, and voltage ac power source.

The power conversion process involves first a utility AC to variable or fixed DC conversion (rectification) and then a DC to variable voltage/current, variable frequency AC conversion (inversion). The rectifier may be controlled or uncontrolled type. The uncontrolled rectifier with diodes only provides a constant DC voltage. The controlled rectifier with self-commutating devices provides a variable DC voltage. In spite of its higher cost and complexity in control, this is upcoming in a few applications where its operational flexibility in providing variable voltage, AC input current shaping, and output ripple reduction are highly required. The inverter stage consists of self-commutating devices such as MOSFETs or IGBTs and only six devices are required[22][23].

3.1.2. Principle of space vector PWM

Figure 3.3 below show the circuit model of a typical three-phase voltage source PWM inverter. S_1 to S_6 are the six power switches that shape the output, which are controlled by the switching variables a, a', b, b', c and c' .When an upper transistor is switched on, i.e, when a, b or c is 1,the corresponding lower transistor is switched off, i.e., the corresponding a′, b′ or c' is 0.Therefore, the on and off states of the upper transistors S_1 and S_6 can be used to determine the output voltage.

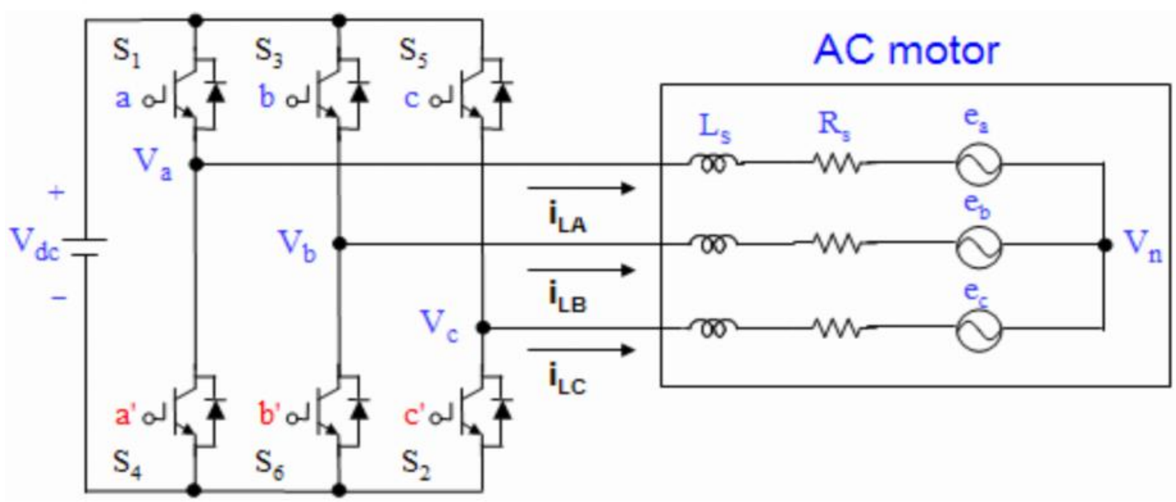

Figure 3.2: Three-phase voltage source PWM Inverter [23]

The relationship between the switching variable vector [a, b, c] and the line-to-line voltage vector [V_{ab}, V_{bc} ,V_{ca}] is given by equestion(3.1) in the following:

$$\begin{bmatrix} V_{ab} \\ V_{bc} \\ V_{ca} \end{bmatrix} = V_{dc} \begin{bmatrix} 1 & -1 & 0 \\ 0 & 1 & -1 \\ -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad (3.1)$$

where

[a b c] are a vector representing the upper switches of the inverter the switching variable vector [a, b, c] and the phase voltage vector $[V_a\ V_b\ V_c]$ can be related as expressed below.

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad \ldots (3.2)$$

There are eight possible combinations of on and off patterns for the three upper power switches. The on and off states of the lower power devices are opposite to the upper one and so are easily determined once the states of the upper power transistors are determined. According to equations (3.1) and (3.2), the eight switching vectors, output line to neutral voltage (phase voltage), and output line-to-line voltages in terms of DC-link V_{dc}, are given below in Table 3.1

Table 3.1. Switching vector, phase voltage and line to line output voltage.

Voltage vectors	Switching vector			Line to neutral voltage			Line to line voltage				
	A	B	C	V_{an}	V_{bn}	V_{cn}	V_{ab}	V_{bc}	V_{ca}	V_{α}	V_{β}
V_0	0	0	0	0	0	0	0	0	0	0	0
V_1	1	0	0	2/3	-1/3	-1/3	1	0	-1	2/3	0
V_2	1	1	0	1/3	-1/3	-1/3	1	0	-1	1/3	$1/\sqrt{3}$
V_3	0	1	0	-1/3	2/3	-1/3	-1	1	0	-1/3	$1/\sqrt{3}$
V_4	0	1	1	-2/3	1/3	1/3	-1	0	1	-2/3	0
V_5	0	0	1	-1/3	-1/3	2/3	0	1	-1	-1/3	$-1/\sqrt{3}$
V_6	1	0	1	1/3	-2/3	1/3	1	-1	0	1/3	$1/\sqrt{3}$
V_7	1	1	1	0	0	0	0	0	0	0	0

The respective voltage should be multiplied by DC voltage

SVPWM treats the sinusoidal voltage (reference voltage) as a constant amplitude vector rotating at constant frequency. This PWM technique approximates the reference voltage V_{ref} by a combination of the eight switching patterns (V_0 to V_7). Coordinate transformation (abc reference frame to the stationary $\alpha - \beta$ frame). That is a three-phase voltage vector is transformed into a vector in the stationary α-β coordinate frame represents the spatial vector sum of the three phase voltage. The vectors (V_1 to V_6) divide the plane into six sectors (each sector: 60 degrees). V_{ref} is generated by two adjacent non-zero vectors and two zero vectors.

Implementation of space vector PWM

For 180° mode of operation, there exist six switching states and additionally two more states, which make all three switches of either upper arms or lower arms ON. To code these eight states in binary (one-zero representation), it is required to have three bits ($2^3 = 8$). Discussion, status of the upper bridge switches will be represented and the lower switches will it's complementary. Let (1) denote the switch is ON and (0) denote the switch is OFF. Table-3.1 gives the details of different phase and line voltages for the eight states

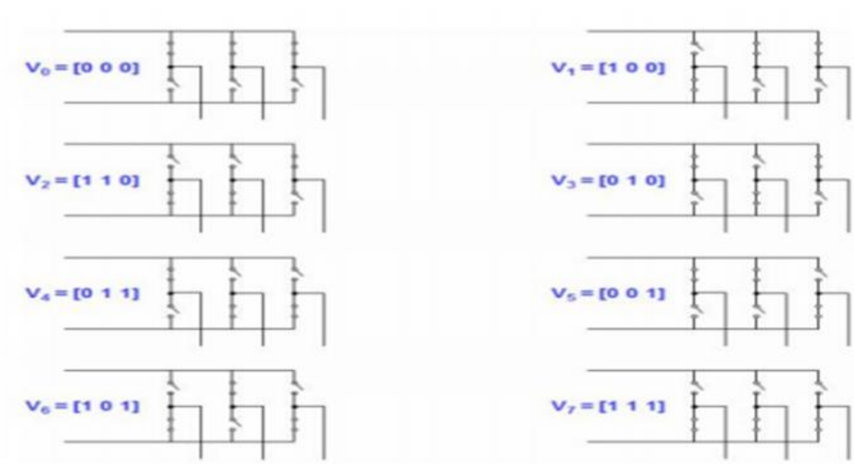

Figure 3.3: The eight inverter voltage vectors V_0 to V_7)[23]

To implement the space vector PWM, the voltage equations in the *abc* reference frame must be transformed into the stationary $(\alpha - \beta)$ reference frame that consists of the horizontal (α) and vertical (β) axes, as a result, six non-zero vectors and two zero vectors are possible. Six nonzero vectors (V_1 to V_6) shape the axes of a hexagonal as depicted in Figure 3.4, and feed electric power to the load or DC link voltage is supplied to the load.

The objective of space vector PWM technique is to approximate the reference voltage vector V_{ref} using the eight switching patterns. One simple method of approximation is to generate the average output of the inverter in a small period, T_z to be the same as that of V_{ref} in the same period. Consider that voltage phasor V_{ref} is commanded. Its position is in between two switching voltage vectors, say V_1 and V_2, and has a relative phase of α from V_1, as shown in Figure 3.4. The commanded voltage phasor can only be realized with the use of the neighboring switching voltage vectors and in this case V_1 and V_2. Taking these switching vectors for a fraction of time as it is not possible to take the fraction of them, and then combining them through the load gives the desired command space voltage phasor.

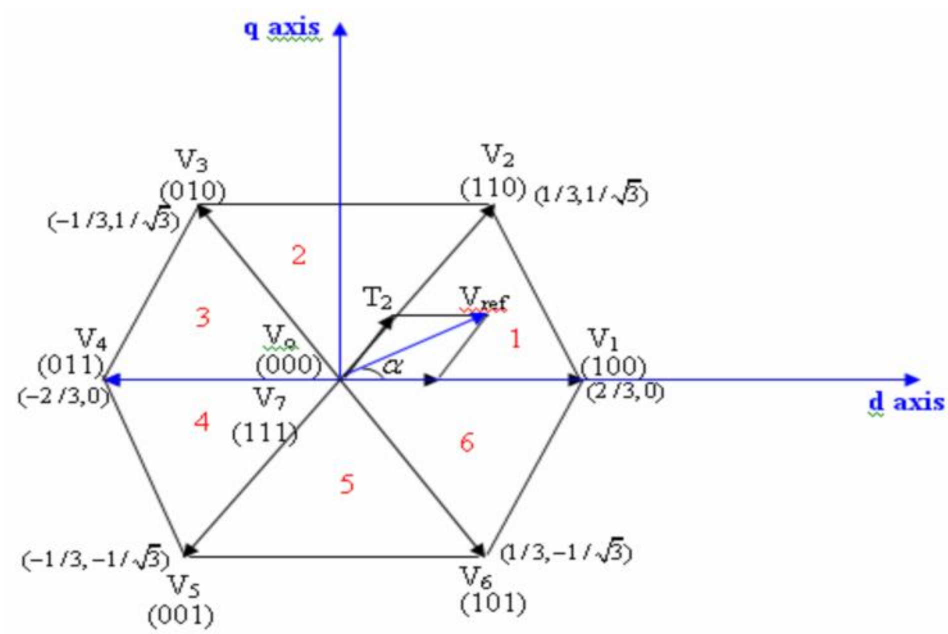

Figure 3.4: Basic switching vectors, sectors and a reference vector[23]

The following steps are followed to implement space vector PWM.

Step 1. Determine V_α, V_β, V_{ref}, and angle (α)

Step 2. Determine time duration T_1, T_2, T_0 for the specific sector where T_1, T_2 are the respective time for which the basic space vectors V_1 and V_2 should be applied within the time period T_z and T_0 is the course of time for which the null vectors V_0 and V_7 are applied.

Step 3. Determine the switching time of each transistor (S_1 to S_2)

Step 1. Determine V_α, V_β, V_{ref}, and angle (α)

The V_d, Vq, V_{ref}, and angle (α) can be determined by using the co-ordinate transformation to 2-phase stationary reference frame as shown in Figure 3.5,can be determined as follows:

$$V_\alpha = V_{an} - V_{bn}\cos 60 - V_{cn}\cos 60 \quad \text{……………………………………………(3.3)}$$

$$= V_{an} - \frac{1}{2}V_{bn} - \frac{1}{2}V_{cn}$$

$$V_\beta = 0 + V_{bn}\cos(30) - V_{cn}\cos(30) \quad \text{………………………...…..………………..(3.4)}$$

$$= \frac{\sqrt{3}}{2}V_{bn} - \frac{\sqrt{3}}{2}V_{cn}$$

The above equations can be written in matrix form as follows

$$\begin{bmatrix} V_\alpha \\ V_\beta \end{bmatrix} = \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{(3.5)}$$

V_{ref} can be determined from the magnitude.

$$|V_{ref}| = \sqrt{V_\alpha^2 + V_\beta^2} \quad \text{………………………………………………………………..(3.6)}$$

where

V_{ref} =The reference space vector voltage crossing every sector

The current sector in which the reference voltage vector found is determined by equation (3.7)

$$\alpha = \tan^{-1}\left(\frac{V_\beta}{V_\alpha}\right) = \omega t = 2\pi f t \dots\dots\dots\dots\dots\dots (3.7)$$

where

f= the fundamental frequency at which the reference voltage rotates.

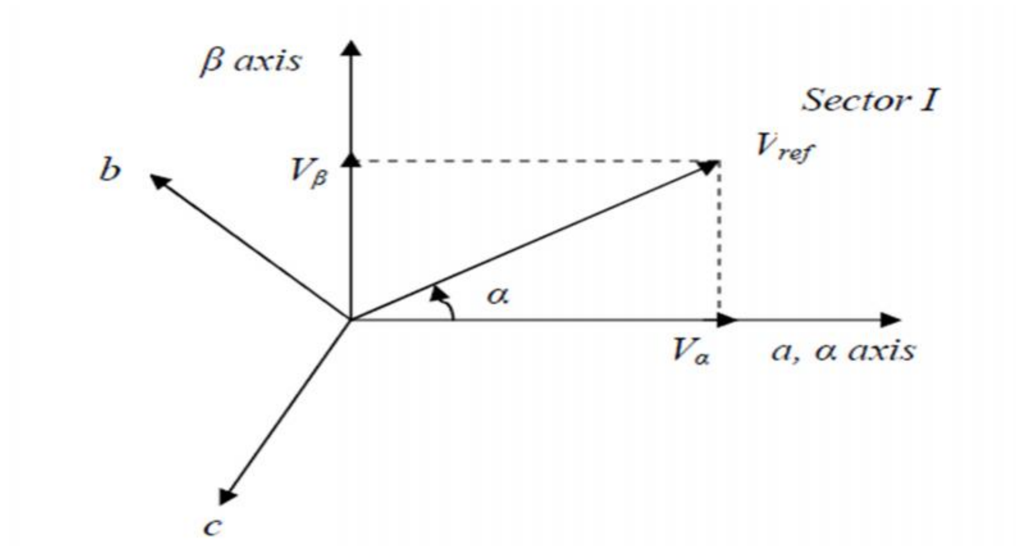

Figure 3.5: Voltage space vector and its components in (α and β) [24]

Step 2: Determine time duration T1, T2, T0

From Figure 3.5, the switching time duration can be calculated as follows:

Switching time duration at Sector 1:

$$\int_0^{T_z} \bar{V}_{ref} dt = \int_0^{T_1} \bar{V}_1 dt + \int_{T_1}^{T_1+T_2} \bar{V}_2 dt + \int_{T_1+T_2}^{T_z} \bar{V}_0 dt \dots\dots\dots\dots\dots\dots (3.8)$$

There for:

$$T_z.\bar{V}_{ref} = T_1\bar{V}_1 + T_2\bar{V}_2 \dots\dots\dots\dots\dots\dots (3.9)$$

For the first sector, the average voltage which is made by vectors V_0, V_1, V_2, and V_7 is given by equation (3.10). (Where, $0 \leq \alpha \leq 60$)

$$T_z.\bar{V}_{ref}\begin{bmatrix}\cos\alpha\\ \sin\alpha\end{bmatrix} = T_1\frac{2}{3}V_{dc}.\begin{bmatrix}1\\0\end{bmatrix} + T_2\frac{2}{3}V_{dc}\begin{bmatrix}\cos\left(\frac{\pi}{3}\right)\\ \sin\left(\frac{\pi}{3}\right)\end{bmatrix} \quad (3.10)$$

There for: $T_1 = T_z.Y.\frac{\sin\left(\frac{\pi}{3}-\alpha\right)}{\sin\left(\frac{\pi}{3}\right)}$

$$T_2 = T_z.Y.\frac{\sin(\alpha)}{\sin\left(\frac{\pi}{3}\right)} \quad (3.11)$$

$$T_0 = T_z - (T_1 + T_2)$$

Where

$$T_z = \frac{1}{f_z} \quad and \quad Y = \frac{|V_{ref}|}{\frac{2}{3}V_{dc}} \quad (3.12)$$

T_1, T_2 are the switching time durations of vectors V_1 and V_2 respectively.

T_0 =the time duration of the zero vector.

Tz=the time period for which one sector is applied.

f_z = the fundamental frequency

Switching time duration at any sector is given by the following equations:

$$T_n = \frac{\sqrt{3}.T_z.|V_{ref}|}{V_{dc}}\left(\sin\left(\frac{\pi}{3} - \alpha + \frac{n-1}{3}\pi\right)\right)$$

$$= \frac{\sqrt{3}.T_z.|V_{ref}|}{V_{dc}}.\sin\left(\frac{n\pi}{3} - \alpha\right)$$

$$= \frac{\sqrt{3}.T_z.|V_{ref}|}{V_{dc}}.\left(\sin\left(\frac{n}{3}\pi\right)\cos\alpha - \cos\left(\frac{n}{3}\pi\right).\sin\alpha\right) \quad (3.13)$$

$$T_{n+1} = \frac{\sqrt{3}.T_z.|V_{ref}|}{V_{dc}}\left(\sin\left(\alpha - \frac{n-1}{3}\pi\right)\right)$$

$$\simeq \frac{\sqrt{3}.T_z.|V_{ref}|}{V_{dc}}.\left(-\cos\alpha.\sin\frac{n-1}{3}\pi + \sin\alpha.\cos\frac{n-1}{3}\pi\right) \quad (3.14)$$

$$T_0 = T_z - (T_n + T_{n+1}) \quad (3.15)$$

where

n = 1 through 6 (sector I to VI) and $0 \leq \alpha \leq 60$

The method used to approximate the desired stator reference voltage with only eight possible states of switches is to combine adjacent vectors of the reference voltage and to determine the time of application of each adjacent vector as shown in Figure 3.5 for the first sector

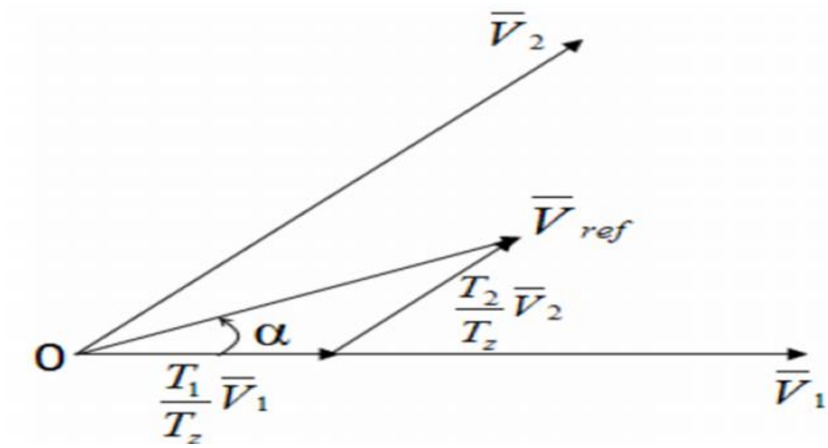

Figure 3.6: Reference vector as a combination of adjacent vectors at sector[23]

Step 3: Determine the switching time of each transistor (S_1 to S_6)

Figure 3.7 shows space vector PWM switching patterns at each sector

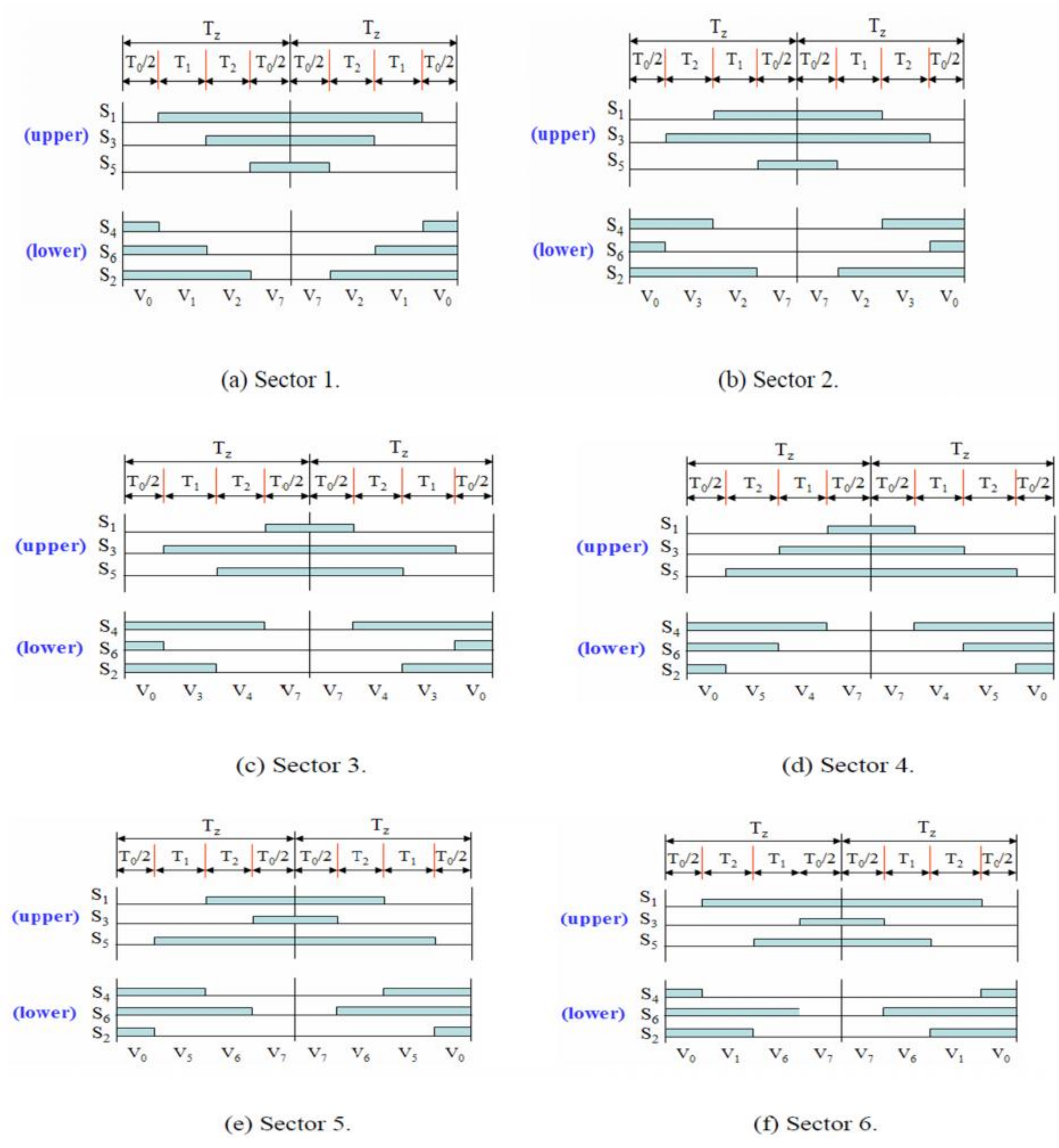

Figure 3.7: Space vector PWM switching patterns at each sector[23]

Based on Figure 3.7, the switching time at each sector is summarized in Table 3.2.

Table 3.2: Switching time calculation at each sectors

Sector	Upper switches (S_1, S_3, S_5)	Lower switches (S_2 S_4 S_6)
1	$S_1=T_1+T_2+T_0/2$	$S_4=T_0/2$
	$S_3=T_2+T_0/2$	$S_6=T_1+T_0/2$
	$S_5=T_0/2$	$S_2=T_1+T_2+T_0/2$
2	$S_1=T_1+T_0/2$	$S_4=T_2+T_0/2$
	$S_3=T_1+T_2+T_0/2$	$S_1=T_0/2$
	$S_5=T_0/2$	$S_2=T_1+T_2+T_0/2$
3	$S_1=T_0/2$	$S_4=T_1+T_2+T_0/2$
	$S_3=T_1+T_2+T_0/2$	$S_1=T_0/2$
	$S_5=T_2+T_0/2$	$S_1=T_2+T_0/2$
4	$S_1=T_0/2$	$S_4=T_1+T_2+T_0/2$
	$S_3=T_1+T_0/2$	$S_6=T_2+T_0/2$
	$S_5=T_1+T_2+T_0/2$	$S_2=T_0/2$
5	$S_1=T_2+T_0/2$	$S_4=T_1+T_0/2$
	$S_1=T_0/2$	$S_6=T_1+T_2+T_0/2$
	$S_5=T_1+T_2+T_0/2$	$S_2=T_0/2$
6	$S_1=T_1+T_2+T_0/2$	$S_4=T_0/2$
	$S_3=T_0/2$	$S_6=T_1+T_2+T_0/2$
	$S_5=T_1+T_0/2$	$S_2=T_2+T_0/2$

Steps to perform vector control

1. Measure the motor quantities (phase voltages and currents).

2. Transform them to the 2-phase system (α ,β) using a clarke transformation.

3. Calculate the rotor flux space vector magnitude and position angle.

4. Transform stator currents to the d-q coordinate system using a park transformation.

5. The stator current torque-(i_q) and flux-(i_d) producing components are separately controlled.

6. The output stator voltage space vector is calculated using the decoupling block.

7. An inverse park transformation transforms the stator voltage space vector back from the d-q coordinate system to the 2-phase system fixed with the stator.

8. Using the space vector modulation, the output 3-phase voltage is generated

3.2. Design of fuzzy logic controller

Fuzzy logic is extensively used in processes where system dynamics is either very complex or exhibit a highly nonlinear character. The first fuzzy logic control algorithm implemented by mamdani was constructed to synthesize the linguistic control protocol of a skilled human operator.

The basic idea behind fuzzy logic control is to incorporate the experience of a human operator in the design of a controller in controlling a process whose input-output relationship is described by a collection of fuzzy control rules (if-then rules) involving linguistic variables. The design procedure of fuzzy logic controller (FLC) is based upon knowledge derived from imprecise heuristic knowledge of experienced operator, does not require precise knowledge of the system model[25][26]. The rule base can be constructed either from human expert knowledge or designer intuition[27]

A. Fuzzy Sets

Define a universe of discourse, X, as a collection of objects all having the same characteristics. The individual elements in the universe X will be denoted as *x*. In classical, or crisp sets the transition for an element in the universe between membership and non-membership in a given set is abrupt and well-defined (said to be ‘‘crisp’’). For an element in a universe that contains fuzzy sets, this transition can be gradual. This transition among various degrees of membership can be lead to the fact that the boundaries of the fuzzy sets are vague and ambiguous Hence, membership of an element from the universe in this set is measured by a function that attempts to describe vagueness and ambiguity. Elements of a fuzzy set are mapped to a universe of membership values (degree to which a quality is possessed) using a function-theoretic form in the range of [0, 1]. If an element in the universe, say x, is a member of fuzzy set A, then this mapping is given by $\mu(x)$ □ [0, 1][28].

A fuzzy set is also a set containing elements that have varying degrees of membership in the set. This idea is in contrast with classical or crisp, set because members of a crisp set would not be members unless their membership was full or complete, in that set (i.e. their membership is assigned a value of 1). Elements in a fuzzy set, because their membership need not be complete, can also be members of other fuzzy set on the same universe. Fuzzy set is mapped to a real numbered value in the interval 0 to 1. If an element of universe, say *x*, is a member of fuzzy set A, then the mapping is given by $\chi_A(x)$ v[0,1][29]. This is the membership mapping and is shown in Figure 3.9

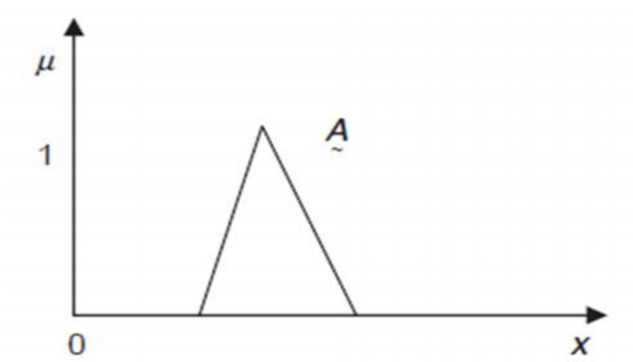

Figure 3.8: Membership function of fuzzy set A[29]

B. Fuzzy Set Operations

Define three fuzzy sets A, B, and C on the universe X. For a given element *x* of the universe, the following function-theoretic operations for the set-theoretic operations of union, intersection, and complement are defined for A, B, and C on X.

Union

$x_{AuB}(x) = x_A(x) \vee x_B(x) = \max(x_A(x), x_B(x))$……………………………………...(3.16)

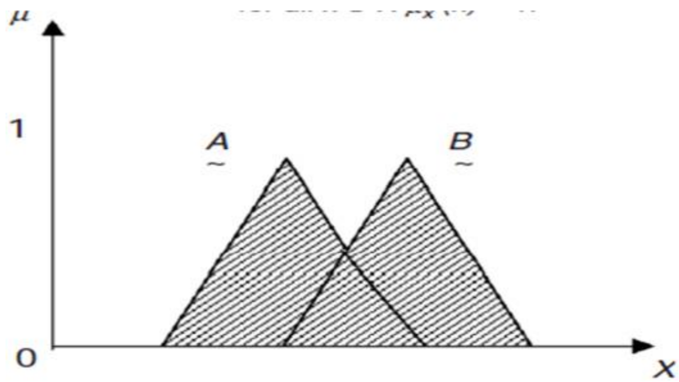

Figure 3.9: Union of fuzzy set[29].

Intersection

$x_{A \cap B}(x) = x_A(x) \wedge x_B(x) = \min(x_A(x), x_B(x))$…………………………………….(3.17)

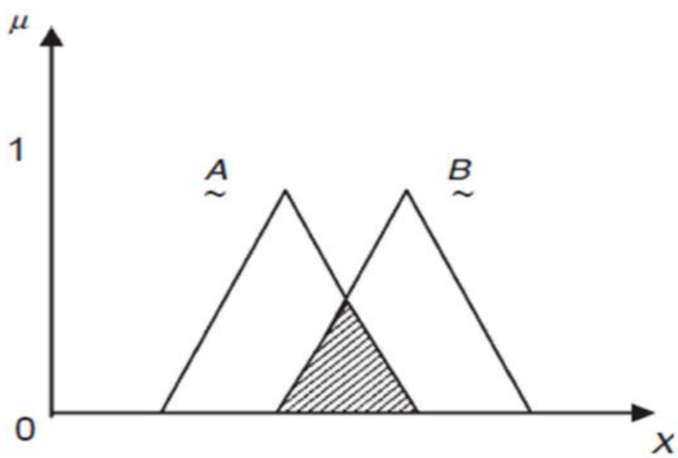

Figure 3.10: Intersection of fuzzy set[29]

Complement

$\chi_{\bar{A}}(x) = 1- \chi_{A}(x)$……………………………………………………….…(3.18)

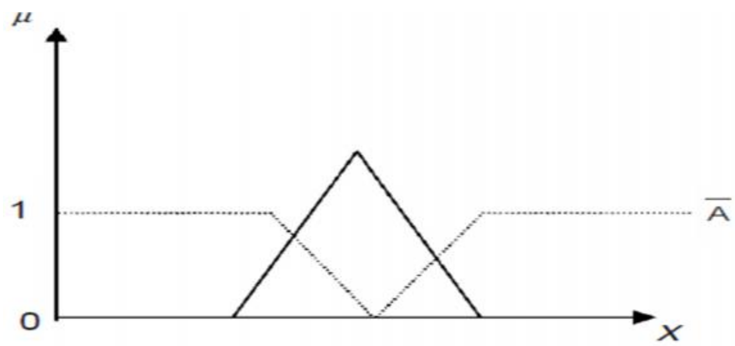

Figure 3.11: Complement of fuzzy set[29]

All operations on classical sets also hold for the fuzzy set except for the excluded middle laws. These two laws does not hold good for fuzzy sets. Since fuzzy sets can overlap, a set and its complement also can overlap[29].

Excluded middle axioms, extended for fuzzy sets, are expressed by $A \cup A \neq X$ (axiom of excluded middle) and $A \cap A \neq \emptyset$ (axiom of contradiction).

C. Linguistic variables

The humans base their thinking primarily on conceptual patterns and mental images rather than on any numerical quantities. Furthermore, humans communicate with their own natural language by referring to previous mental images with rather vague but simple terms. A linguistic variable associates words or sentences with a measure of belief functions, also called membership function. The set of values that it can take is called term set. Each value in the set is a fuzzy variable defined over a base variable. The base variable defines the universe of discourse for all the fuzzy variables in the term set

D. Membership function (MF)

Membership function is a function that changes the crisp variables (classical sets) in to fuzzy sets. All information contained in a fuzzy set is described by its membership function. MFs describe the degree of confidence of all elements in the universe of discourse to each fuzzy set. The shape of membership functions depends on the system that is to be controlled. Trapezoidal (or triangular) shapes are the most frequently used membership functions. Every element in the universe of discourse is a member of the fuzzy set to some grade, may be even zero. The function that ties a number to each element x of the universe is called the membership function μ(x)[21][27][29].

E. components of fuzzy logic controller

Fuzzy logic controller consists of four main parts as shown in Figure 3.9. Fuzzification, rule base, inference engine and defuzzification.

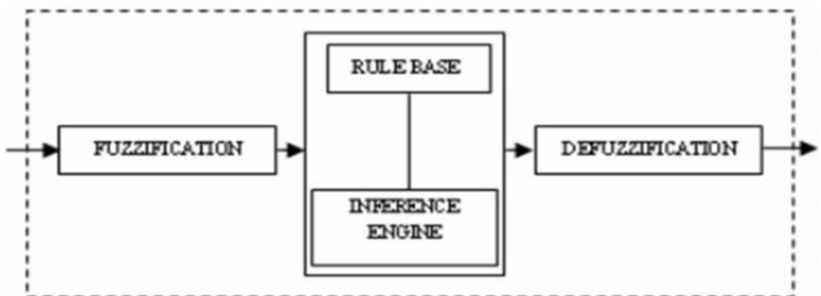

Figure 3.12: Fuzzy Logic Controller block[27]

3.2.1. Fuzzification of crisp data

The process of converting classical data or crisp data into fuzzy data or membership functions is called fuzzification. The fuzzification sub process turns the measurement into a degree of membership [30]. It transforms the physical values of the error signal, rate of change of error which is input to the fuzzy logic controller, into a fuzzy set consisting of an interval for the range of the input values and an associate membership function describing the degrees of the confidence of the input belonging to this range. The conversion process is performed by a membership function. The purpose of this fuzzification step is to make the input physical signal compatible with the fuzzy control rule base in the core of the controller.

In the other word fuzzification module converts the crisp values of the control inputs into fuzzy values, so that they are compatible with the fuzzy set representation in the rule base. The output of fuzzification module will be the input to the next module, the fuzzy logic IF-THEN rule base for the control which requires fuzzy-subset inputs in order to be compatible with the fuzzy logic rules.

3.2.2. Rule base for control strategy of the system

The rule base is essentially the control strategy of the system. It is usually obtained from expert knowledge or heuristics and expressed as a set of IF-THEN rules. The rules are based on the fuzzy inference concept and the antecedents and consequents are associated with linguistic variables [31]. Basically a linguistic controller contains rules in the if-then format, but they can be presented in different formats [32]. Generally, the fuzzy rules are depended on the plant to be controlled, the type of the controller and from practical experience. The performance specifications of the systems such as rise time overshoot, settling time and error steady state can be improved by tuning value of parameters K_p and K_i of the PI controller, because each component has its own special purposes. Formulation of the fuzzy rules is based on the simulation analysis of the PMSM drive system. The rule base is expressed as IF (antecedent)-THEN (consequent) rules. In our case, the rules are presented similar to the one below.

1. If error is NB and change in error is NB then K_{p1} is VB and K_{i1} is VB
2. If error is NB and change in error is NM then $K_{p1}$is VB and K_{i1} is VB
3. If error is NB and change in error is NS then $K_{p1}$is MB and K_{i1} is MB
4. If error is NB and change in error is ZE then $K_{p1}$is MB and K_{i1} is MB
5. If error is NB and change in error is PS then $K_{p1}$is B and K_{i1} is B
6. If error is NB and change in error is PM then $K_{p1}$is B and K_{i1} is B
7. If error is NB and change in error is PB then $K_{p1}$is M and K_{i1} is M

where

NB,NM,NS ZE,PS,PM PB is the linguistic label of inputs and ZE M,B,MB and VB is the linguistic label of output.

3.2.3. Inference engine for combining input MF with control rules

The inference engine is used to combine membership functions with the control rules to derive the fuzzy output. For each rule, the inference engine looks up the membership values in the condition of the rule. It has three main operations. Aggregation, activation and accumulation

Aggregation: The aggregation operation is used when calculating the degree of fulfillment or a firing strength of the condition of a rule. The main function of aggregation is to combine the generated membership function values that come from error and change in error measurement. Aggregation is sometimes also called fulfillment of the rule or firing strength.

Activations: The activation of a rule is the deduction of the conclusion, possibly reduced by its firing strength. Min or product (*) is used as the activation operator. The multiplication scales the membership curves thus preserving the initial shape rather than clipping them as the min operation does. Both methods work well in general. Although the multiplication results in a slightly smoother control signal [32][33].

Accumulations: All activated conclusions are accumulated, using either sum or max operations. In order to draw conclusions from a rule base, we need a mechanism that can produce an output from a collection of IF-THEN rules. This is done using the computational rule of inference. The inference mechanism provides the mechanism for referring to the rule base such that the appropriate rules are fired. The two most commonly used inference procedures in FLC are madman's max-min and max-algebraic product (or Max-Dot) composition. In this thesis a Madman's max-min composition inference method is used.

Figure 3.13 shows graphical construction of control signal when single rule is fired for both error and change-in error is NB with the value of -0.988, and then Kp1 & Ki1 is VB with the value of 0.947.

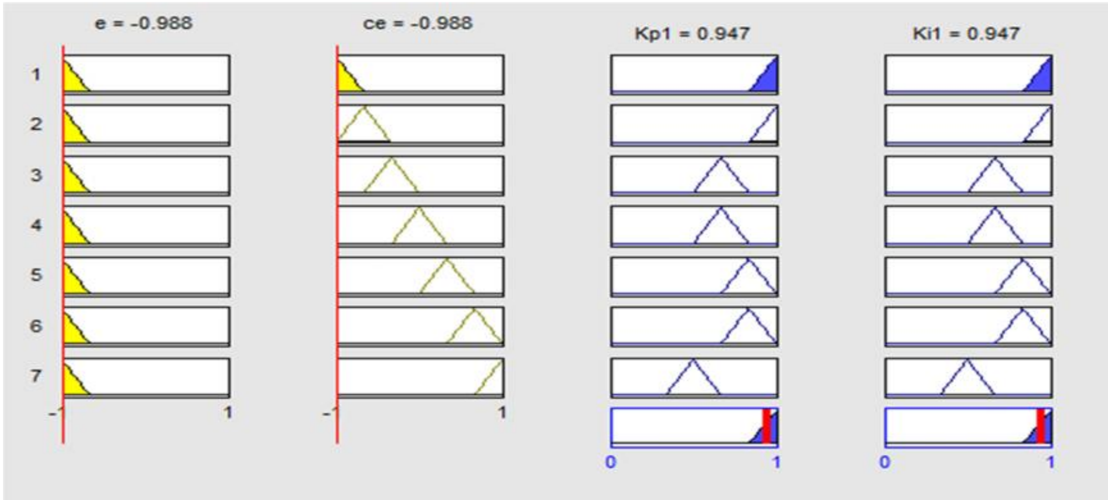

Figure 3.13: Construction of control signal when single rule is fired for e & ce is NB

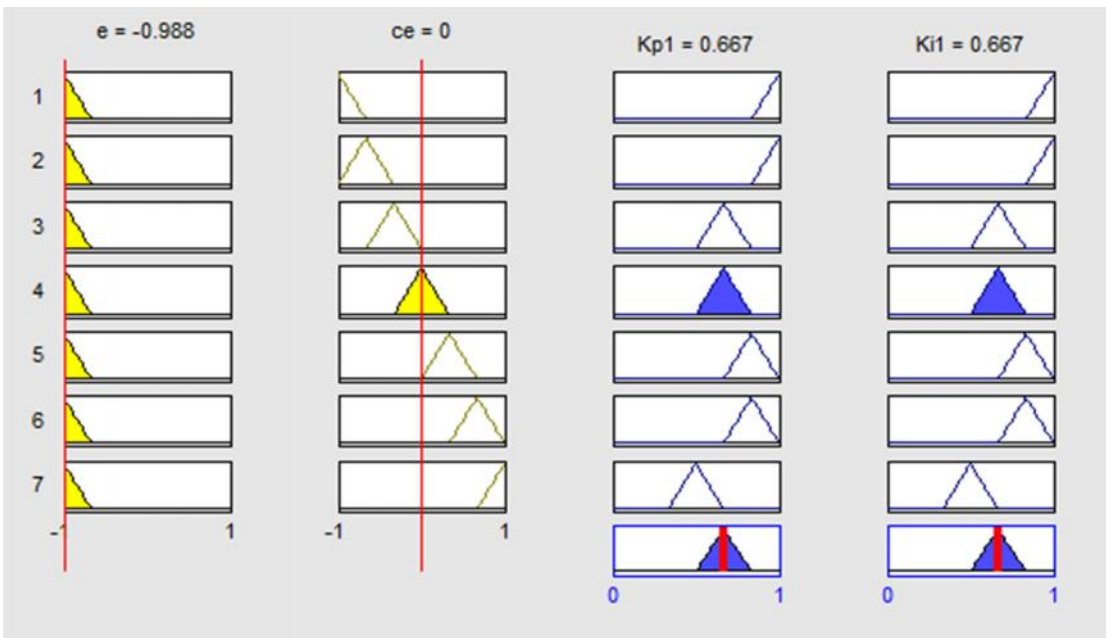

Figure 3.14: Construction of control signal when single rule is fired for e is NB & ce is ZE

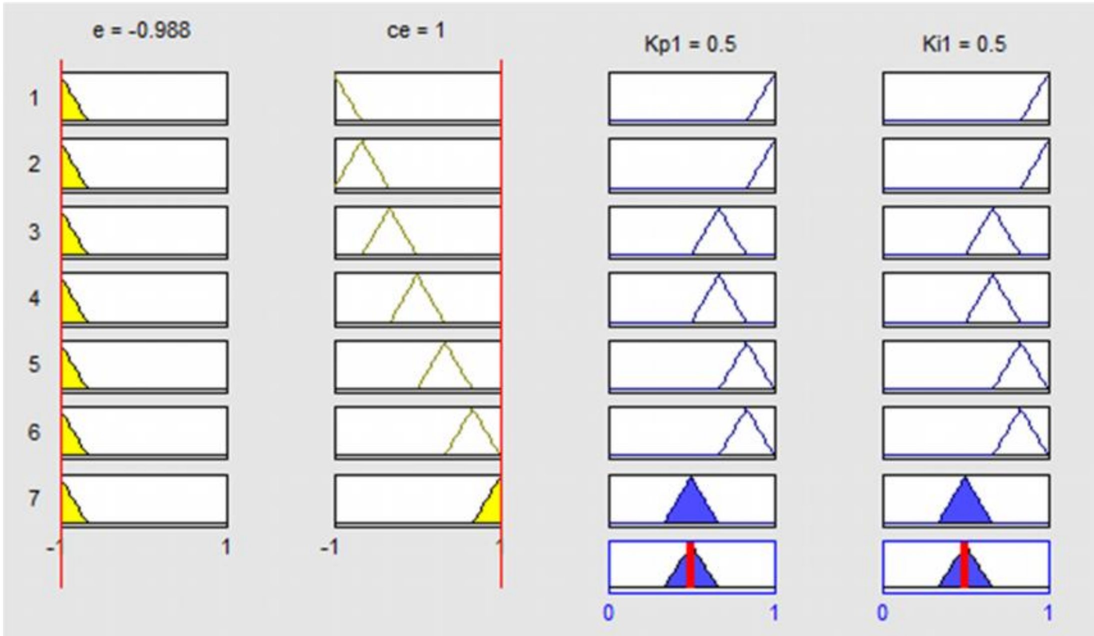

Figure 3.15: Construction of control signal when single rule is fired for e is NB & ce is PB

3.2.4. Defuzzifications of fuzzy data

Defuzzification unit in FLC is the reverse of the fuzzification process. It converts the fuzzy controller output fuzzy variables in to a crisp real signal. It uses different methods to calculate each associated output and put them into a table. The defuzzification uses methods such as center of gravity, center of sum and weighted mean to convert from the inference mechanism into the crisp values applied to the actual system. There are several commonly used, logically meaningful, and practically effective defuzzification formulas available, which are by nature weighted average formulas in various forms. Unfortunately, there is no systematic procedure for choosing a defuzzification strategy. In this thesis a center of gravity method is adopted for difuzzification as given below.

$$x^* = \frac{\sum \mu_x(x).x}{\sum \mu_x(x)} \quad \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots \ldots (3.19)$$

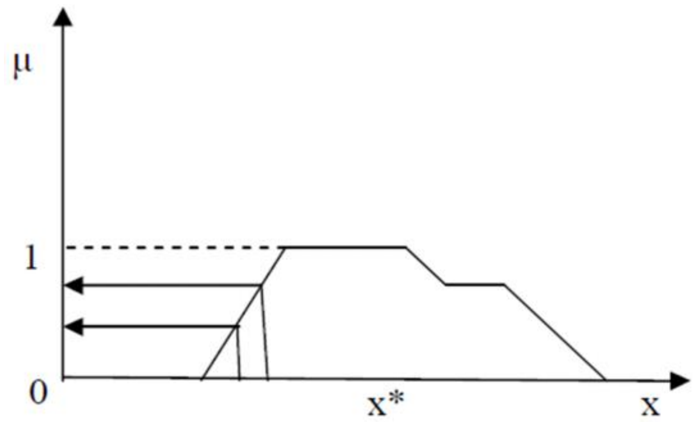

Figure 3.16: Center of graphite defuzzification method[21].

3.3. Design of self-tuning PI controller using fuzzy logic approach

To achieve a high performance speed control of permanent magnet synchronous motor, self-tuning PI controller based on fuzzy logic approach is proposed in speed loop of the control system. Fuzzy logic controller is used to aid conventional method to enhance the output performance by retuning the gains of PI controller at different operating conditions. Self-tuning fuzzy PI controller means that the two parameters K_p and K_i of PI controller are tuned by using fuzzy tuner [34][35]. The coefficients of the conventional PI controller are not often properly tuned for the nonlinear plant with unpredictable parameter variations. Hence, it is necessary to automatically tune the PI parameters but as permanent magnet synchronous motor is nonlinear plant, fuzzy logic idea is included to tune PI parameters. The FLC has two inputs. Speed error $e(k)$and change in speed error $ce(k)$ and one output $u(k)$ which represents the change in quadrature reference current $I_q^*(k)$. $e(k)$ and $ce(k)$ are calculated as

$$e(k) = \omega^*(k) - \omega(k) \quad \ldots \ldots \ldots \quad (3.20)$$

$$ce(k) = \frac{d}{dt}\big(e(k)\big) \quad \ldots \ldots \ldots \quad (3.21)$$

where $\omega^*(k)$ is reference speed and $\omega(k)$ is actual speed value.

The control action of the PI controller after self-tuning can be describing as:

$$U(k) = K_{p2} * e(k) + K_{i2} \sum_{i=0}^{k-1} e(i) \quad (3.22)$$

$$K_{p2} = K_{p1} * K_p \quad (3.23)$$

$$K_{i2} = K_{i1} * K_i \quad (3.24)$$

where

K_{p2} and K_{i2} are the new gains of the PI controller.

K_{p1} and K_{i1} are the gains outputs of fuzzy control that are varying on line with the output of the system under control and K_p,K_i are the optimal values of the PI controller before tuning. The structure of the overall controller to control the speed of sensored feedback permanent magnet synchronous motor is shown in the figure 3.18

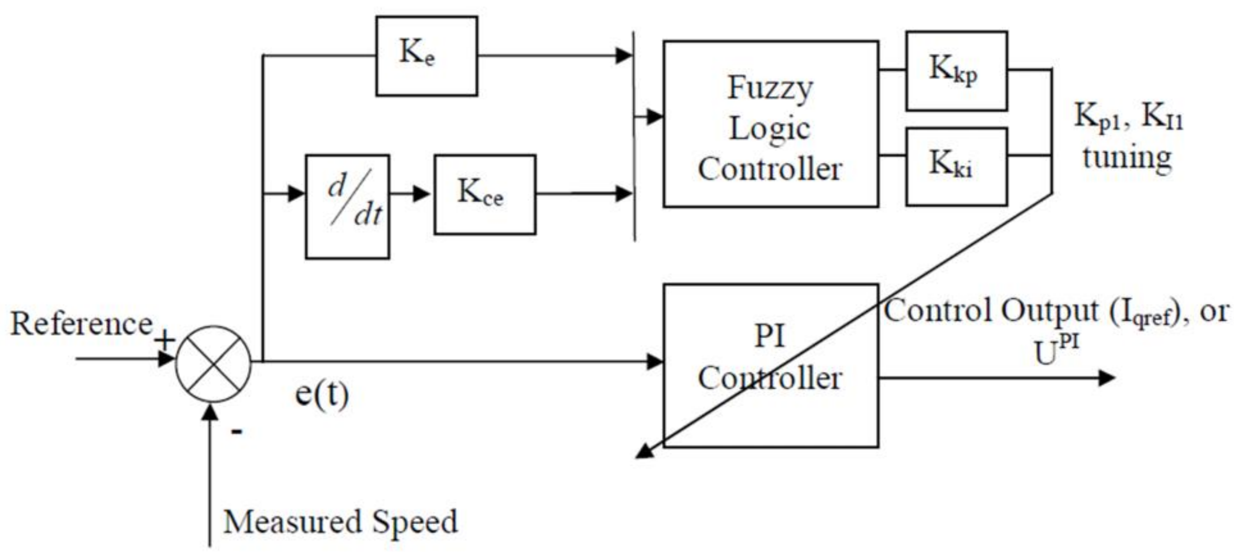

Figure 3.17: Structure of self-tuning fuzzy PI controller[21]

3.3.1. Fuzzy language of input and output variables

The gains K_e, K_{ce} K_{kp} and K_{ki} in figure 3.18 are called scalling factors used to obtain normalized inputs and outputs for the fuzzy logic controller. The scalling factor should fulfill the following requirements. Thus are:

$K_{ce} * K_{ki} = K_p$ ……………………………………………………………… (3.25)

$K_e * K_{kp} = K_i$ ……………………………………………………………… (3.26)

$K_e * max|e| \leq 1$ ……………………………………………………………… (3.27)

$K_{ce} * max|\Delta e| \leq 1$ ……………………………………………………………… (3.28)

Most of the time scaling factors is determined by trial and error, but in this thesis, all factors are approximately one and saturation block is used before fuzzy inputs. For the system under study the universe of discourse for both inputs e(t) and ce(t) is normalized to the range -1 to 1 as the range of the universe of discourse for the membership functions is selected to be from -1 to 1 and the linguistic labels(fuzzy sets) are defined as{Negative big, Negative medium, Negative small, Zero, Positive small, Positive medium and Positive big} and are referred to in the rules bases as {NB,NM,NS,ZE,PS,PM,PB }as it is shown in Figure 3.18. The linguistic labels of the outputs K_{p1} and K_{i1} in the range 0 to 1 are {Zero, Medium small, Small, Medium, Big, Medium big, very big} and referred in the rule bases as {Z, MS, S,M, B, MB, VB}. The membership functions of the inputs fuzzy sets and output fuzzy sets are shown in Figure 3.18 and 3.19. These membership functions are chosen based on intuition, simply derived from the capacity of humans to develop membership functions through their own understanding. The number of membership functions used, and the overlapping characteristics are the most important[6][36].

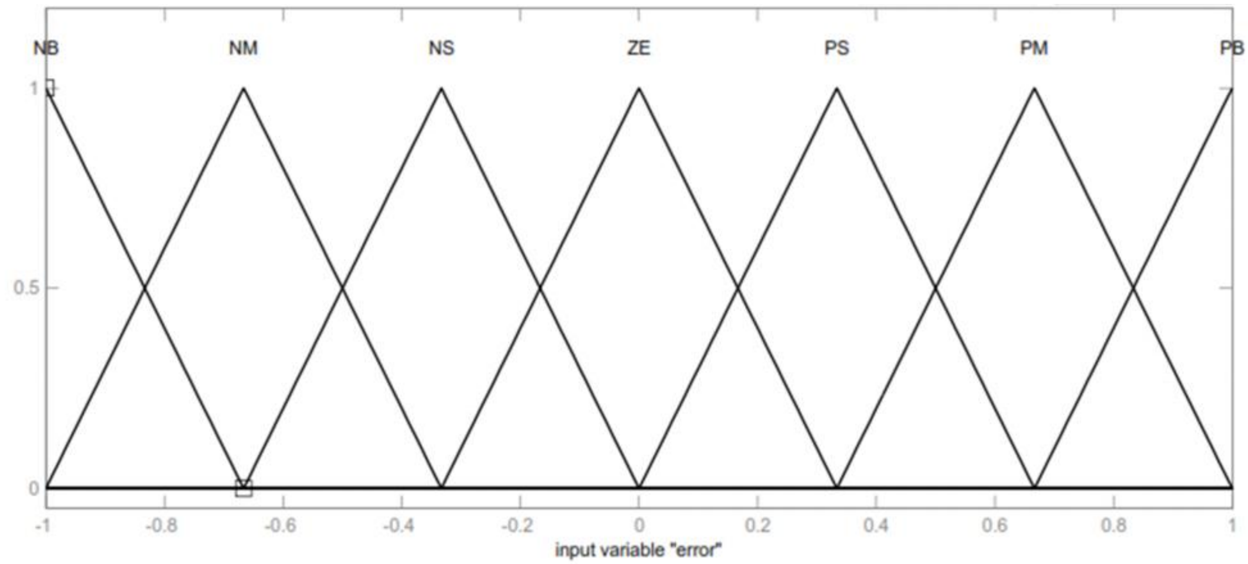

Figure 3.18: Member ship function of inputs (e & ce)

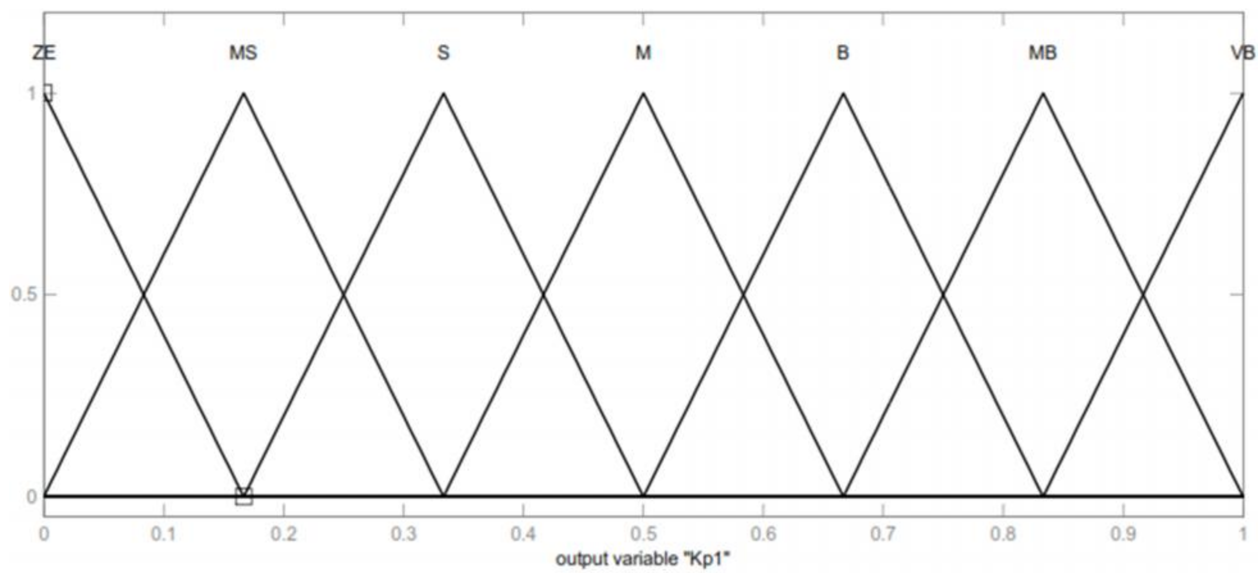

Figure 3.19: Member ship function of outputs (Kp1 and Ki1)

The PI parameters are tuned by using fuzzy inference, which provide a nonlinear mapping from the error and derivation of error to PI parameters. The accurate mathematical model is not necessary while tuning PI controller based on fuzzy logic approach. The rules designed are based on the characteristic of the permanent magnet synchronous motor and properties of the PI controller. Therefore, the fuzzy reasoning of fuzzy sets of outputs is gained by aggregation operation of fuzzy sets inputs and the designed fuzzy rules. The aggregation and defuzzification method are used respectively max-min and centroid method. Regarding to the fuzzy structure, there are two inputs to fuzzy inference: error e(t) and derivative of error de(t) and two outputs for each PI controller parameters respectively K_{p1} and K_{i1}. Mamdani model is applied as structure of fuzzy inference with some modification to obtain the best value for K_p and K_i as shown in Figure 3.20

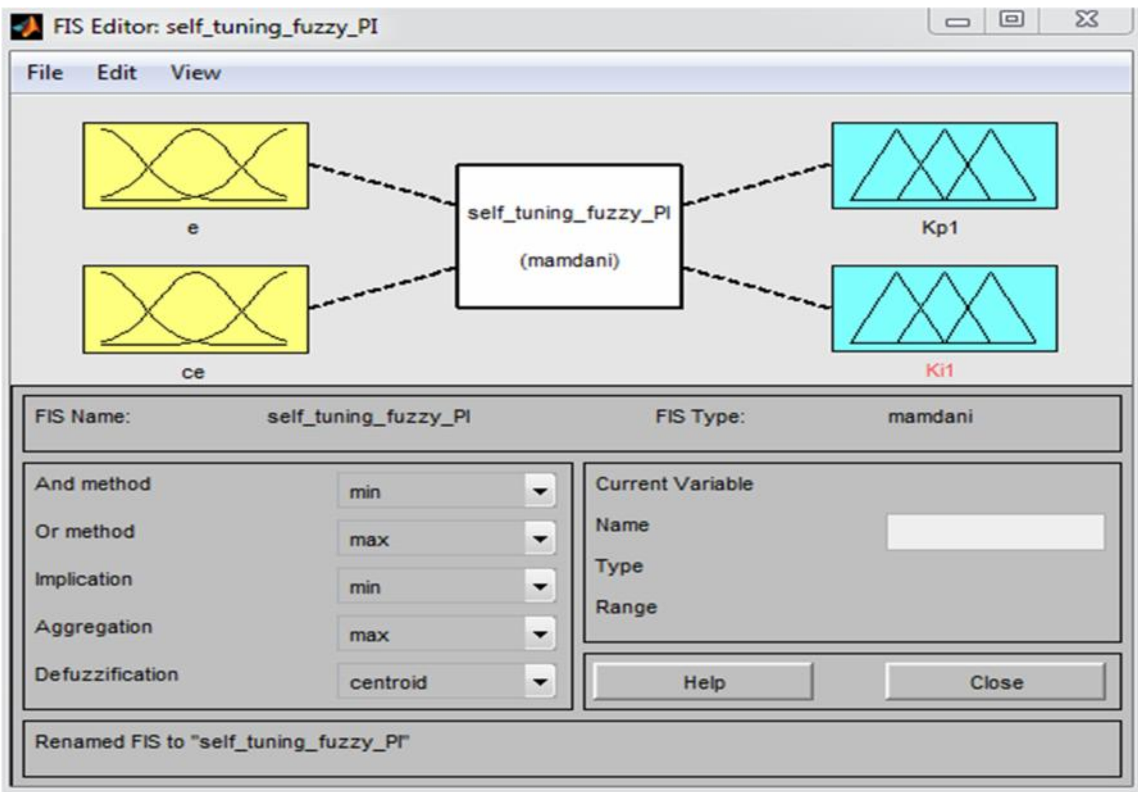

Figure 3.20: Fuzzy inference block

Table 3.3. Rule base for determining the gains Kp1 and Ki1.

Ce / E	NB	NM	NS	ZE	PS	PM	PB
NB	VB	VB	MB	MB	B	B	M
NM	PB	MB	MB	B	B	M	S
NS	MB	MB	B	B	M	S	S
ZE	MB	B	B	M	S	S	MS
PS	B	B	M	S	S	MS	MS
PM	B	M	S	S	MS	MS	ZE
PB	M	S	S	MS	MS	ZE	ZE

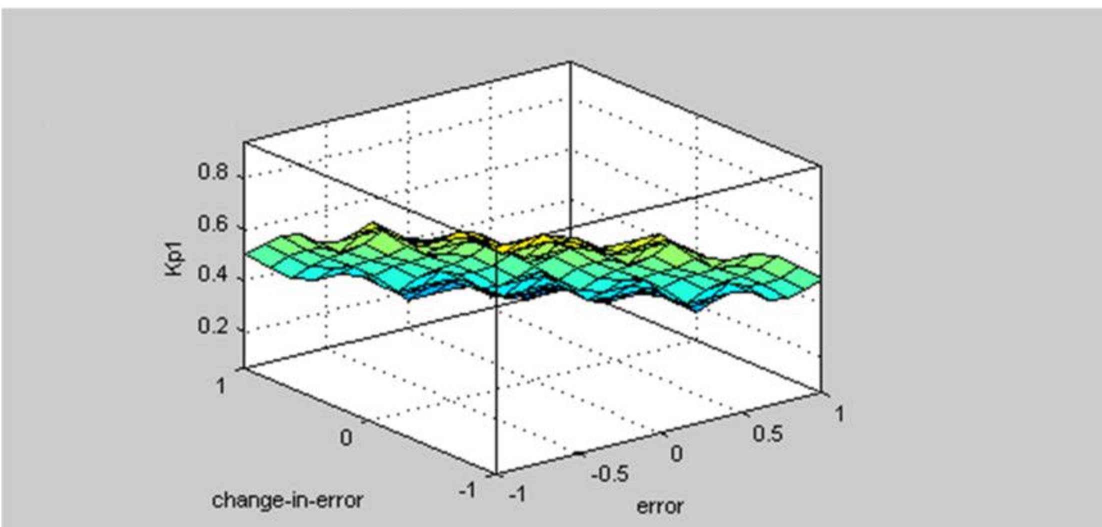

Figure 3.21: Variation of Kp1 and Ki1 with respect to error and change in error of speed.

3.4. PI Controller

Today in process control in large industries, more than 95% of the control loops are of PID type, most loops are actually PI control. In the control of dynamic systems, no controller has enjoyed both the success and the failure of the PID control. There is actually a great variety of types and design methods for the PID controller [37].

The PID controller is a combination of PD and PI controllers. It used in closed loop (feedback) control system, generally with Single Input-Single Output (SISO). A portion of the signal being fed back is:

- Proportional to the signal (P)
- Proportional to integral of the signal (I)
- Proportional to the derivative of the signal (D)

PID control works well on SISO systems of 2^{nd} Order, where a desired Set Point can be supplied to the system control input. PID control also handles step changes to the set point especially well:

- Fast Rise Times
- Little or No Overshoot
- Fast settling Times
- Zero Steady State Error

The PID controller can be used to improve both the system transient response and steady state errors. PD controller is used to improve the system transient response. PI control is used in cascade compensation (integral term in the feedback path is equivalent to a differentiator in the forward path). PI design for a plant transfer function G(s) is the same as PD design of G(s)/s. A better design is often possible by "almost canceling" the controller zero and the controller pole (negligible effect on time response). PI control is used only if P-control meets the transient response but not the steady-state error specifications. Otherwise, another control is selected.

In practical applications the PI controller zero is placed very close to its pole located at the origin so that the angular contribution of this “dipole” to the root locus is almost zero. A PI controller is used to improve the system response steady state errors.

In general the effects of increasing parameters are:

Parameter:	Rise Time	Overshoot	Settling Time	S.S.Error
Kp	Decrease	Increase	Small Change	Decrease
Ki	Decrease	Increase	Increase	Eliminate
Kd	Small Change	Decrease	Decrease	None

In general, in the practical applications pure derivative action is never used due to the derivative kick produced in control signal for step input and undesired noise amplifications.

The inner loop system uses PI controller to control current and the outer loop system uses self-tuning PI controller based on fuzzy logic approach to control the speed of permanent magnet synchronous motor.

3.4.1. Design of current controller

For decoupled torque and flux control the rotor flux and torque current have to be properly controlled, which makes current regulator very important part of the field oriented based control system. In order to design the regulator the existing delays in the system need to be taken into account. These delays in the case of a motor drive are due to the digital Implementation of the control (which implies the sampling of signals), the use of filters, the processing of the control algorithm and the use of pulse width modulators (PWM).

3.4.2. PI controller design for torque producing component (i_q)

PID controller consists of the three terms: proportional (P), integral (I), and derivative (D). Its behavior can be roughly interpreted as the sum of the three term actions: The P term gives a rapid control response and a possible steady state error; The I term eliminates the steady state error; and

The D term improves the behavior of the control system during transients. In this thesis derivative part is set to zero. The objectives which should be achieved by the application of the control system are associated with the following control system features:

- Regulating performance
- Robustness
- Noise attenuation.

The objectives can be stated in many ways such as through: specifications within the time domain; specifications within the frequency domain; robustness specifications and other specifications. The specifications within the time domain give some values related to the shape of control system signals in the time domain.

In order to design PI-controllers it is necessary to know the closed loop transfer function by relating input and output. Using the q-axis voltage equation, torque equation, and mechanical equation, the block diagram of permanent magnet synchronous motor (PMSM) in *q*-axis is shown in Figure 3.22

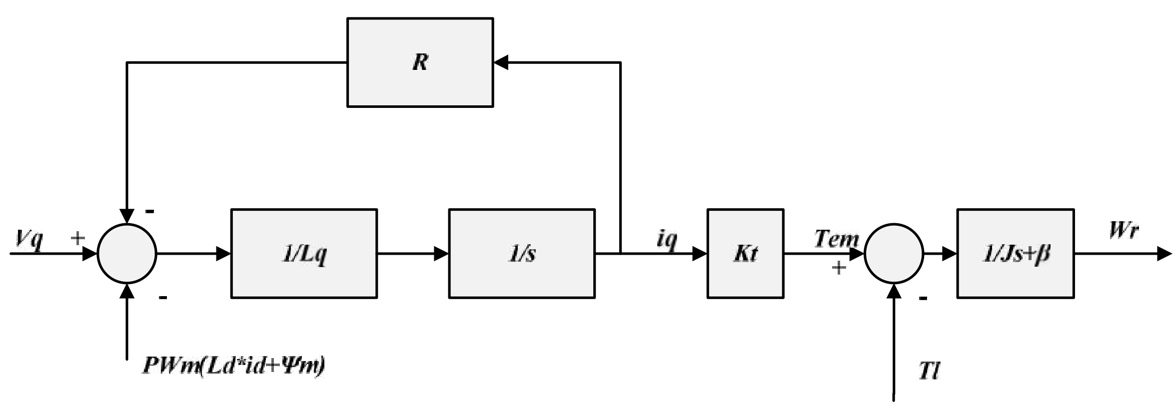

Figure 3.22: Block diagram of permanent magnet synchronous motor in q-axis.

Here, the design objective is to determine the PI's gains k_p and k_i or (T_i) so as to achieve a good closed loop response. The transfer function between torque producing component, i_q and input voltage V_q in q-axis is:

$$\frac{i_q}{V_q} = \frac{1}{L_q s + R_s} = \frac{1/R_s}{T_s s + 1} \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots \dots . (3.29)$$

Equation (3.29) shows transfer function of torque producing linear component of decoupled system of V_q and i_q. Once the input output relation is determined V_q as input and i_q as output PI-controller is designed for the purpose of regulating current error in q-axis. For the design of PI-controller of current loop in q-axis, the block diagram in Figure 3.23 is used.

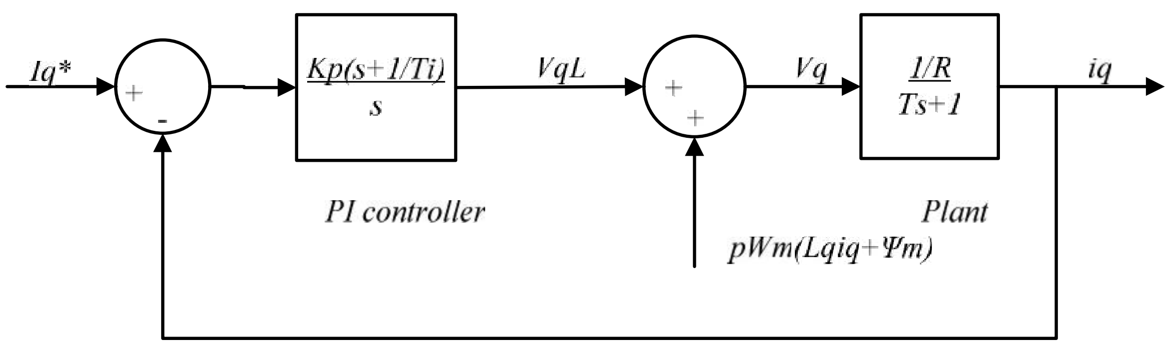

Figure 3.23: PI controller design for loop iq's(torque producing) component.

Design specification [21]

PI controller is designed to regulate error or deviations from the set point. The proposed controller has to meet the following specification.

Peak overshoot (MP)=20%,

Rise time(tr) =0.0012sec =10Ts

Where,T_s is sampling time of the current (inner)loop = 12μs.

From the design specification the damping coefficient (ζ) and undammed natural frequency (ω_n) are assigned as in equation (3.30).

$\zeta >= 0.48$, $w_n >= 1500$ rad/sec ………………………………………………..…… (3.30)

Using equation (3.30) the damping coefficient and natural frequency are chosen to be:

$\zeta = 0.7$ and $w_n = 1500$ rad/sec ………………………….……………………… (3.31)

From Figure 3.24 the overall 2nd order transfer function T(s) obtained between the torque component i_q and its reference i_q^* is given as (3.32)

$$T(s) = \frac{i_q}{i_q^*} = \frac{k_p * s + k_i}{L_q s^2 + (R_s + K_p)s + K_i} \quad \text{(3.32)}$$

The equation of 2nd order closed loop system is given as:

$$T(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad \text{(3.33)}$$

The values of K_p & K_i are determined by comparing the denominator of equation (3.32 & 3.33) for inner loop(current loop).

For the specific target, motor used in simulation with parameters:

Stator resistance: R_s= 2.875Ω Stator inductance: L_q=L_d= 0.0085H

Rotor flux linkage: Ψ_m = 0.175 V.s Rotor inertia: J=0.0008Kgm²

Friction coefficient: B= 0.00005396Nms Number of pole: P=4

$$K_p = 2\zeta\omega_n L_q - R_s \quad \text{(3.34)}$$

$$k_i = \omega_n^2 * L_q \quad \text{(3.35)}$$

In this thesis, damping ratio (ζ) and natural frequency (ω_n) to be selected as 0.7 and 1500 red/sec respectively. After calculation, the values of K_p and K_i are adjusted to get optimal values which is 110 and 20 respectively.

3.4.3. PI controller design for flux producing component

From the d-axis stator voltage equation the block diagram of permanent magnet synchronous motor (PMSM) in *d*-axis is shown as follow in Figure 3.24.

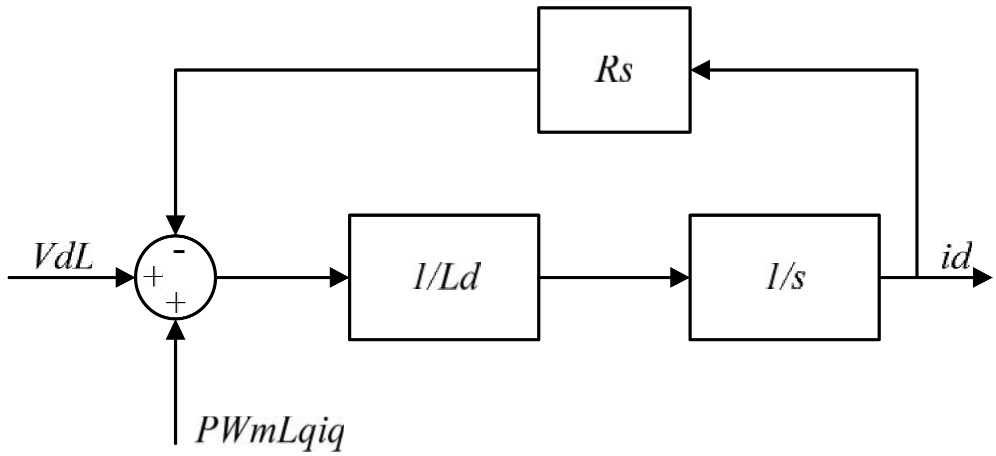

Figure 3.24: Block diagram of permanent magnet synchronous motor in d-axis.

The transfer function between flux component i_d and input voltage, V_d in *d*-axis stator current is given by equation (3.35).

$$\frac{i_d}{V_d} = \frac{1/R_s}{T_e s + 1} \quad \text{(3.36)}$$

where

In the constant torque operation region of permanent magnet synchronous motor due to the presence of the constant flux of the permanent magnet, there is no need to generate flux by means of the i_d current, i.e. i_d=0, which decreases the stator current and increases the efficiency of the drive. In order to design the PI current regulator in *d* axis it is followed the same procedure as q axis and k_p and k_i are the same as equation (3.34) and (3.35) since L_d=L_q in surface permanent magnet synchronous motor (SPMSM). In the overall block diagram shown in figure 3.25, the two PI controllers are designed to control the current, whereas self-tuning PI controller based on fuzzy logic approach is designed for speed loop only.

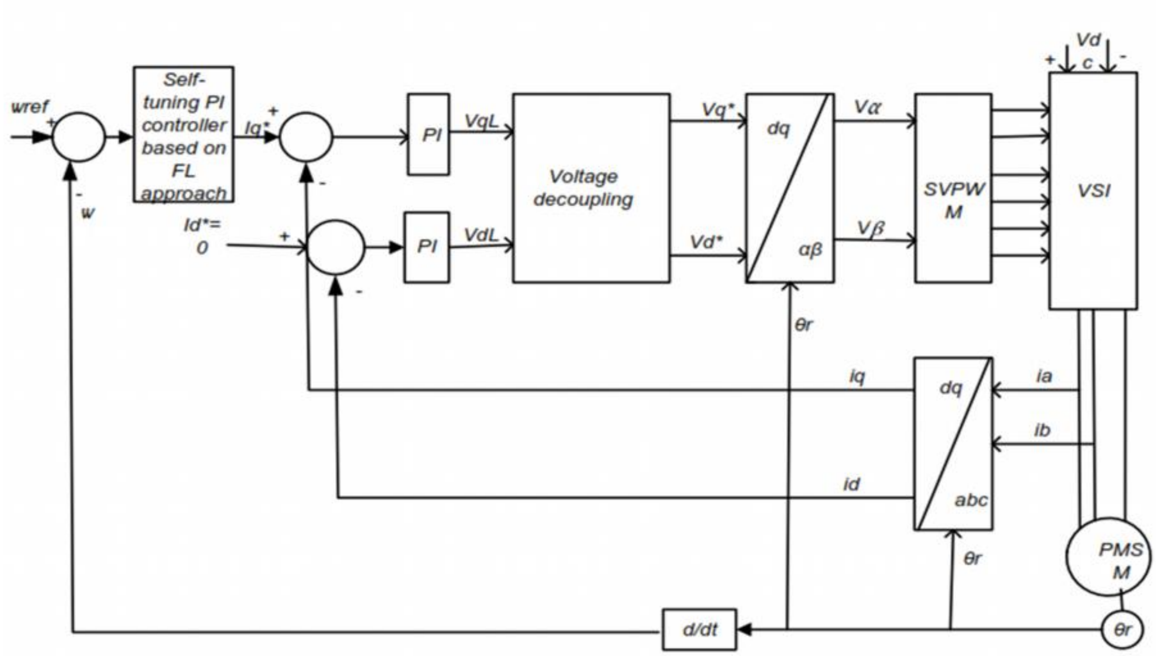

Figure 3.25: The overall control structure of permanent magnet synchronous motor

CHAPTER FOUR

Simulation studies and results discussion

4.1. Results discussion

This chapter is concerned with the simulation of the closed loop control of permanent magnet synchronous motor drive system using PI controller for inner loop and self-tuning PI controller based on fuzzy logic approach for outer loop with matlab/Simulink. matlab/Simulink R2013a is used for simulation of permanent magnet synchronous motor drive. SIMULINK® is a toolbox extension of the matlab program. It is a program for simulating dynamic systems. Simulink has the advantages of being capable of complex dynamic system simulations. Simulink simulates analogue systems and discrete digital systems.

The overall system simulation block includes different sub-functional blocks, such as: PI controller for vector control, coordinate transformation block which used to transform stationary reference frame to rotor reference frame. SVPWM block, inverter (VSI) block and motor block are taken from simulink library. SVPWM block: To generate/approximat e reference voltage (V_{ref}), Inverter (VSI) block: used to convert dc-voltage to ac-voltage. The inverter consists of “universal bridge” block with parameter of IGBT. Fuzzy logic develop 7x7 rules for speed loop controller. When each rule is fired, controller parameters (K_{p1} and K_{i1}) are varied with equal magnitude as shown in table 4.1. Because same rule is applied for each PI parameters (K_{p1} and K_{i1}).

Table 4.1: Variation of Kp1 and Ki1 when each rule is fired

e \ ce	NB	NM	NS	ZE	PS	PM	PB
NB	0.947	0.948	0.833	0.817	0.667	0.661	0.5
NM	0.931	0.822	0.819	0.655	0.651	0.479	0.333
NS	0.833	0.822	0.67	0.65	0.484	0.336	0.333
ZE	0.803	0.655	0.617	0.469	0.33	0.303	0.167
PS	0.667	0.631	0.481	0.314	0.314	0.166	0.166
PN	0.649	0.489	0.336	0.317	0.167	0.165	0.0525
PB	0.5	0.333	0.319	0.167	0.166	0.0531	0.0523

When error and change in error is NB, the values of Kp1 and K_{i1} is equal to 0.947. Then the new PI controller parameters are calculated as:

$K_{p2} = K_{p1} * K_p = 0.947*110 = 104.17$

$K_{i2} = K_{i1} * K_i = 0.947*20 = 18.94$

In this thesis, same rule is developed for K_{p1} and K_{i1}.

The motor parameters used for simulations are given in table 4.2.

Table 4.2: Permanent magnet synchronous motor parameters specifications.

Parameters	Values
PWM inverter frequency	2.5KHz
dc-link voltage	320V
Pole pairs(p)	2x4
Magnetic flux linkage(Ψ_m)	0.175ωb
Rotor inertia(J)	0.0008Kg.m^2
Stator resistance	2.875W
Inductance(L_q=L_d)	0.0085H
Friction coefficient(B)	0.00005396N.m/(rad/sec)
Rated power	0.7KW

The overall Matlab/Simulink model of the drive system is shown in figure 4.1

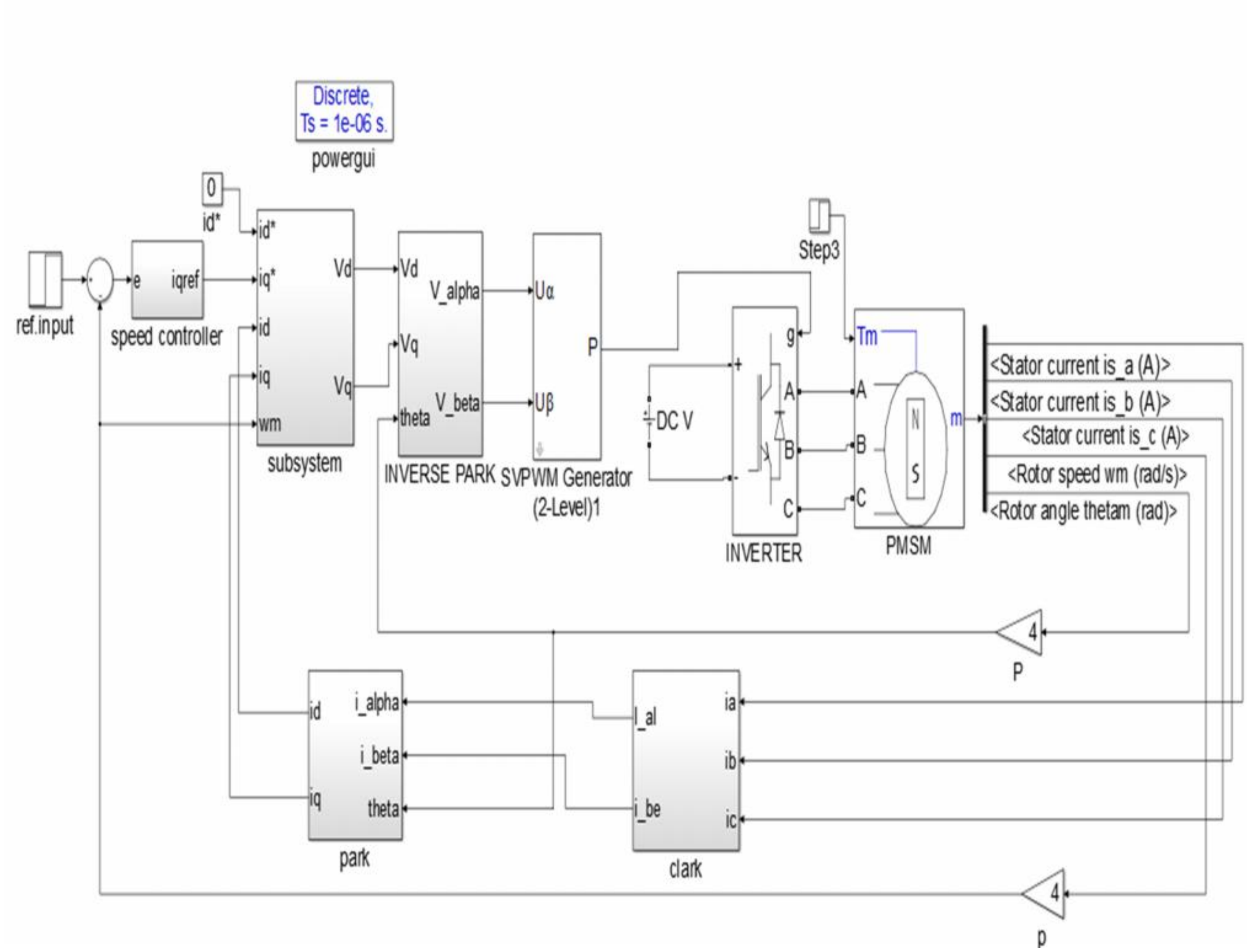

Figure 4.1: Overall system simulation block diagram

Table 4.3 summarizes the comparison between self-tuning PI controller with fuzzy and without fuzzy for different conditions such as with load and without load as well as with parameter variations.

Table 4.3: Comparison of conventional (PI) and FL-PI controller performance.

No	Performance of the system with different controller .		Conventional PI controller	FL based self-tuning PI-controller
1	With no-load torque at 100rad/sec	Overshoot(M_p)	34.85 %	3.38 %
		Settling time(t_s)	0.0182sec	0.0175 sec
2	When 0.25Nm load torque is applied 100rad/sec	Overshoot(M_p)	32.9 %	3.56 %
		Settling time(t_s)	0.0169sec	0.0155 sec
3	When Rs is increased by 50%	Overshoot(M_p)	31.97%	3.84%
		Settling time(t_s)	0.0465sec	0.0599 sec
4	When Rs is increased by 25%	Overshoot(M_p)	32.4%	0.99%
		Settling time(t_s)	0.0249sec	0.0394 sec
5	When inductance(L) increased by 50%	Overshoot(M_p)	34.64%	0.99%
		Settling time(t_s)	0.069sec	0.029 sec
6	When inductance(L) increased by 25%	Overshoot(M_p)	34.64%	5.66%
		Settling time(t_s)	0.0592sec	0.0459 sec
7	When both Rs & L increased by 50%	Overshoot(M_p)	31.97%	1.96%
		Settling time(t_s)	0.0590sec	0.0285 sec
8	When both Rs & L increased by 25%	Overshoot(M_p)	32.9%	0.99%
		Settling time(t_s)	0.0369sec	0.0209 sec

The PI controllers are designed for controlling stator current/indirectly torque of the system and the controller parameters proportional and integral gains are calculated as stated in chapter three and adjusted to get the optimal value. The output of PI controllers are added with nonlinear part of the system as shown in figure 4.2 and input to inverse park transform. The rotating reference frame, V_d and V_q are transformed to stationery reference frame V_α and V_β which is supplied to SVPWM. SVPWM, inverter permanent magnet synchronous motor blocks are taken from matlab simulink library.

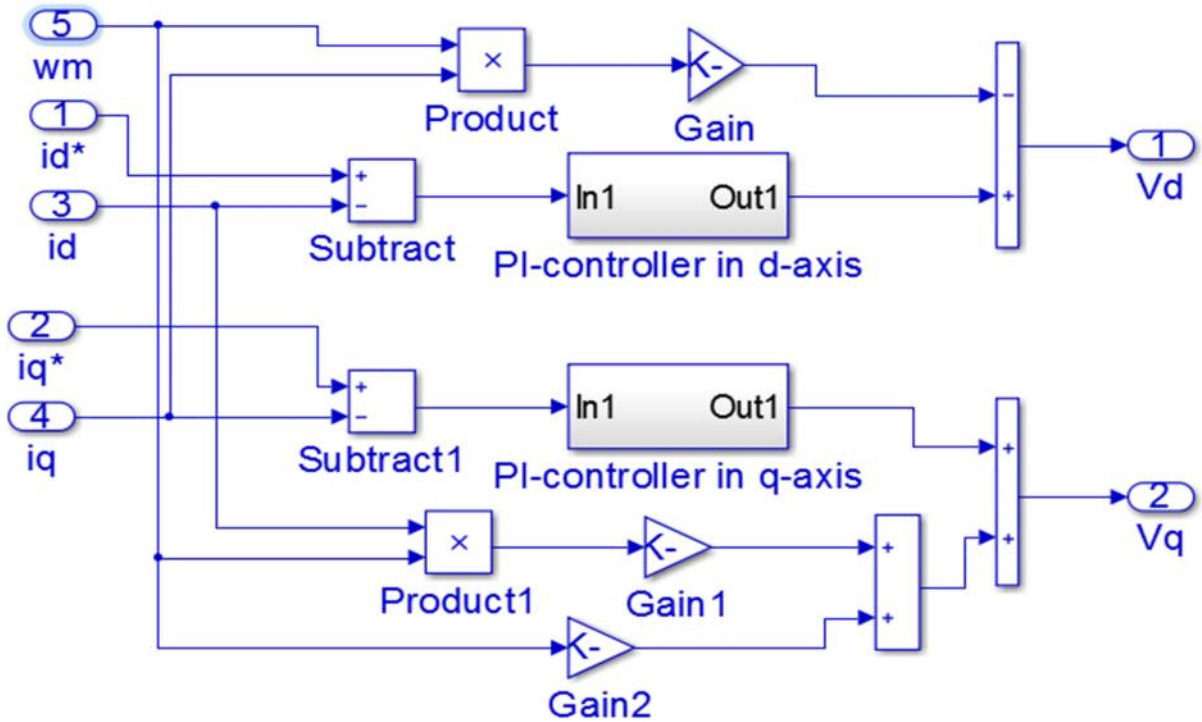

Figure 4.2: Current controller and non-linear part in matlab/Simulink.

Figure 4.3 shows the block diagram of self-tuning PI controller which used to control the speed of permanent magnet synchronous motor based on fuzzy logic approaches.

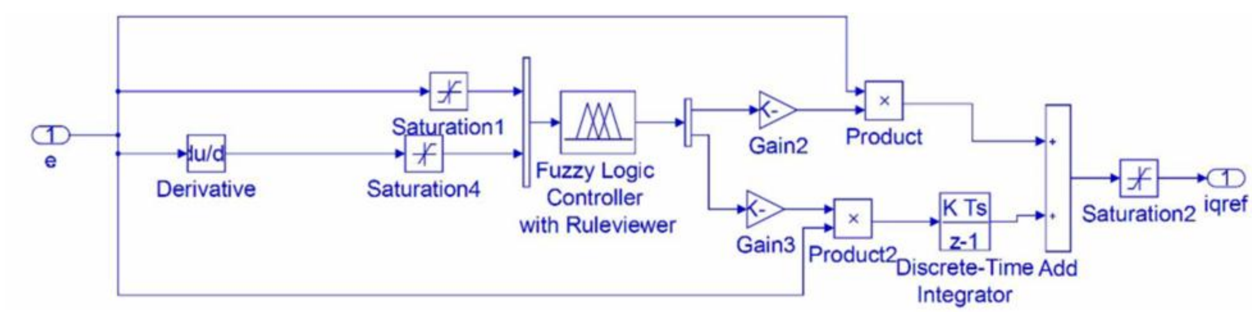

Figure 4.3: Block diagram of speed controller in matlab/Simulink.

4.2. Simulation results and analysis

Overall matlab/Simulink model of the PMSM drive allows to analyses the behavior of the machine using sensored feedback for position and self-tuning fuzzy-PI controller algorithm for speed control at different operating conditions/modes. The first simulation is performed for step speed input without load torque (T_l=0) as shown with Figure 4.4 which shows the speed response of the motor when the motor is commanded at reference speed of 100rad/sec. At the begging speed response with PI controller has shoot up to 153rad/sec for 100rad/sec reference speed and set to reference speed with 1.12 steady state error.

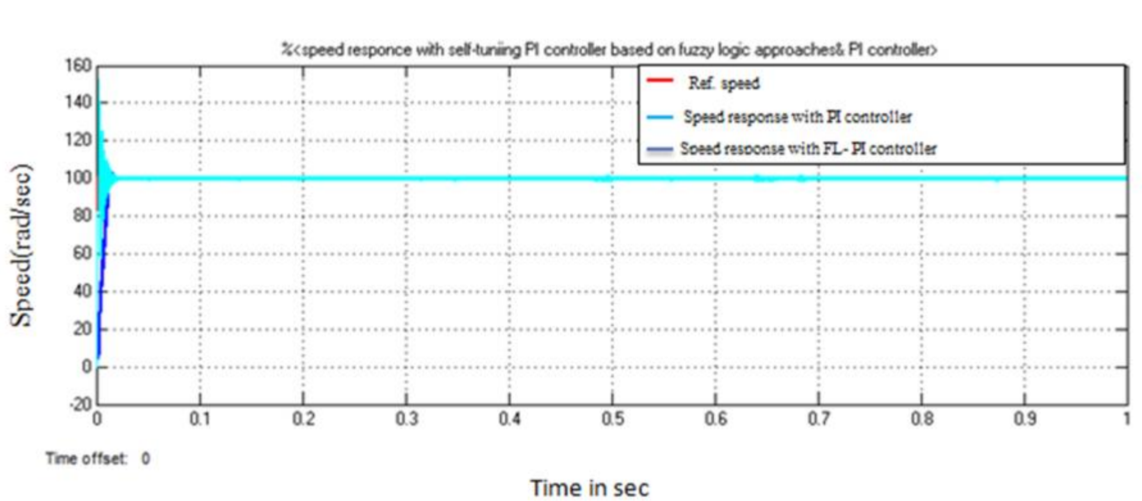

Figure 4.4: Step speed response of the permanent magnet synchronous motor for 100rad/sec with fuzzy for no-load torque.

Figure 4.5 show the speed response of the motor when the motor is commanded at reference speed of 100rad/sec for no load torque by using fuzzy logic based self-tuning PI controller. At the begging speed response shoot 107rad/sec for 100rad/sec reference speed and set to reference speed with zero steady state error.

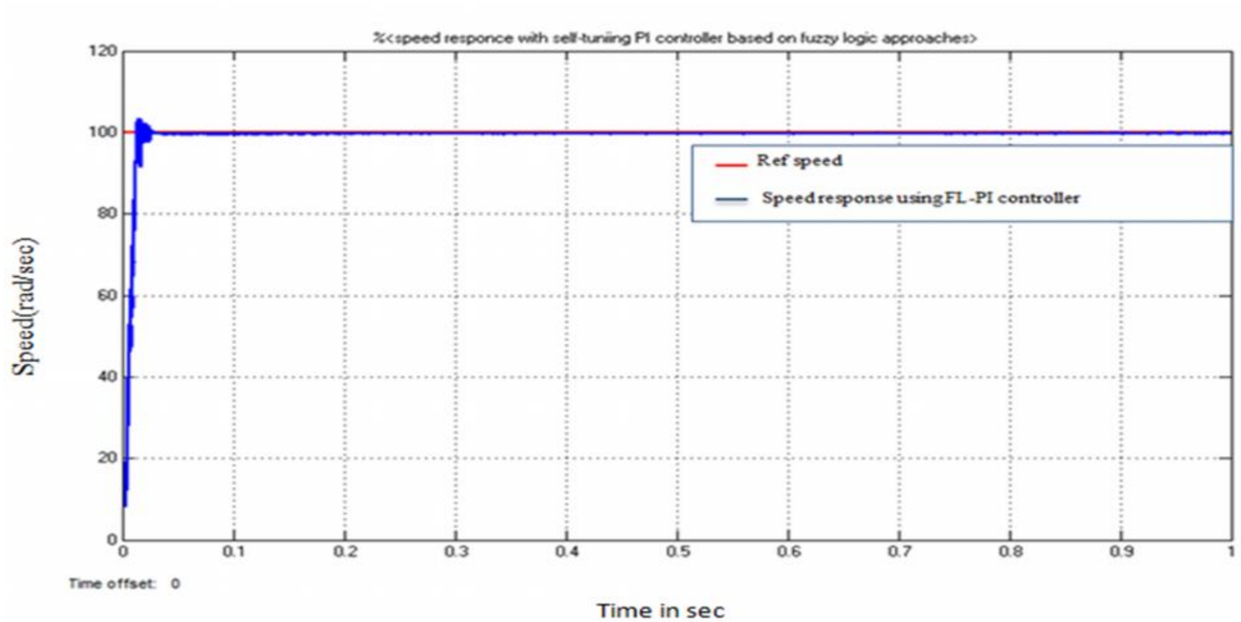

Figure 4.5: Speed response of the permanent magnet synchronous motor for 100rad/sec reference speed with only fuzzy based self-tuning PI controller for no-load torque is applied.

Figure 4.6: Shows the simulation result of three phase stator currents which are generated by three phase inverter. This three phase inverter is controlled by SVPWM for appropriate stator current generation. These three phase currents should be equal magnitude and 120^o phase shift with each other for appropriate rotating flux generation as shown below

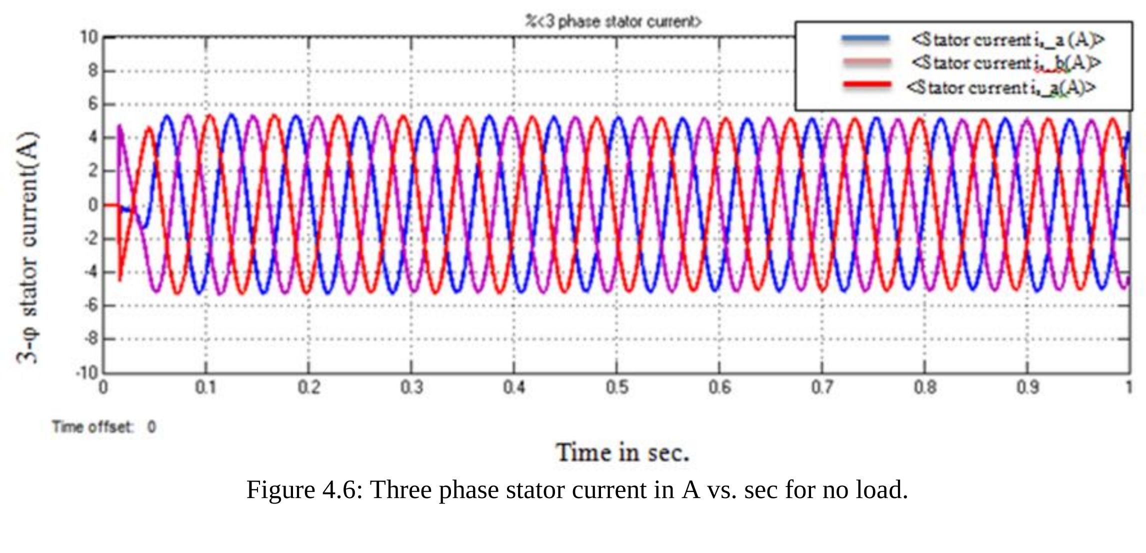

Figure 4.6: Three phase stator current in A vs. sec for no load.

As we see from the above Figure, the appropriate stator phase current value is generated with good accuracy. Therefore, the system can feed the appropriate stator voltage to the motor. If voltage applied to the motor is applied with appropriate magnitude and frequency, the speed of the motor is respected as set the reference values.

Figure 4.7: Shows the generated electromagnetic torque. The starting torque until 0.02 sec is around 3.8Nm. This is due to primarily the acceleration of the rotor to reach/arrive the steady state speed of 100rad/sec as well as to overcome the friction retard without load torque ($T_l = 0$) as shown blow.

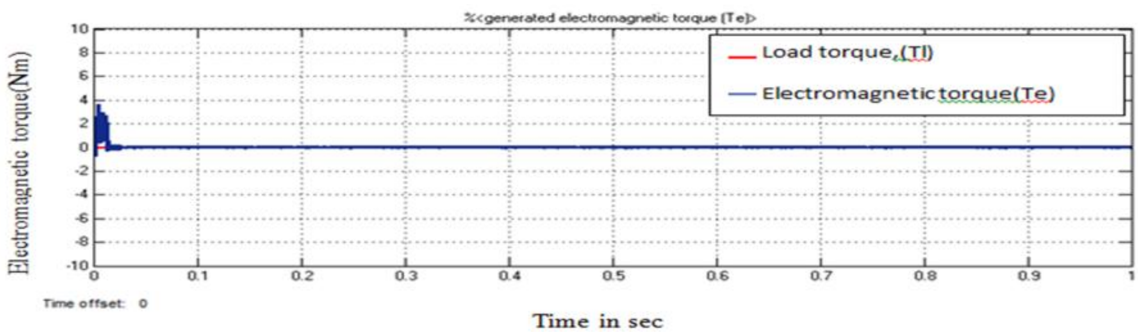

Figure 4.7: Developed electromagnetic torque for step speed input.

However; after 0.02sec almost the generated torque is reduced to 0Nm to support approximately zero retard friction. Figure 4.8 shows the speed response of the machine with fuzzy for 0.25Nm load torque is applied at 0.02sec.

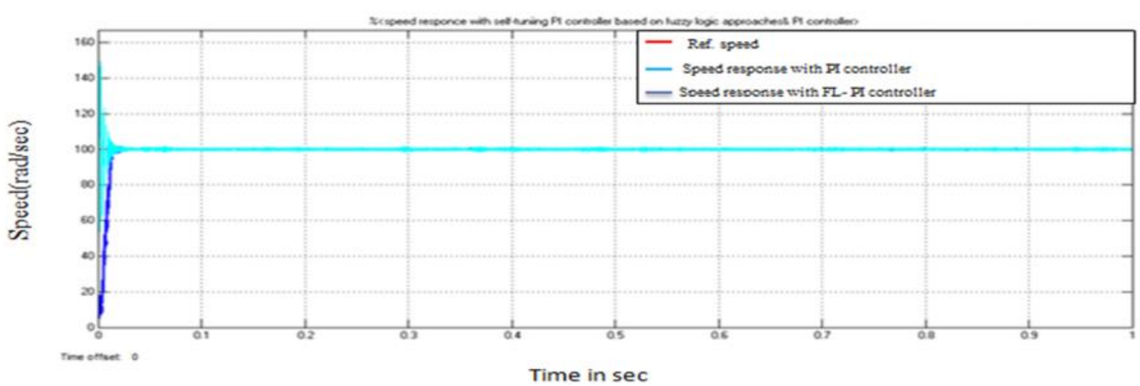

Figure 4.8: Step speed response of permanent magnet synchronous motor with fuzzy for 0.25Nm load torque is applied.

The speed response of permanent magnet synchronous motor with fuzzy does not change with small change in load torque. The settling time is almost the same as it read from the graph. There is zero steady state error with fuzzy based self-tuning PI controller and 1.47% steady state error with conventional PI controller. Figure 4.9 shows 3-ϕ stator current response of the permanent magnet synchronous motor with fuzzy for load torque is applied. The magnitude of current does not changed with the load change.

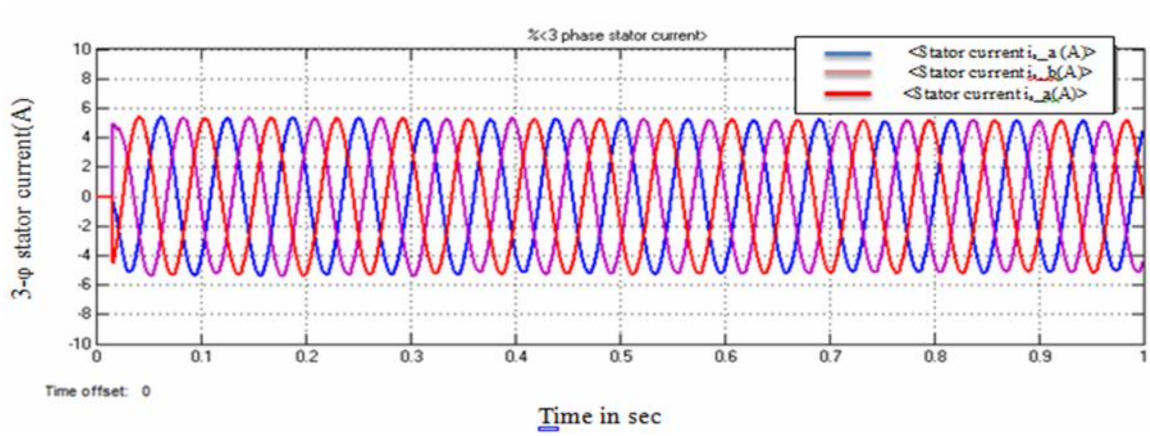

Figure 4.9: 3-ϕ stator current response of the permanent magnet synchronous motor with fuzzy when 0.25Nm load is applied.

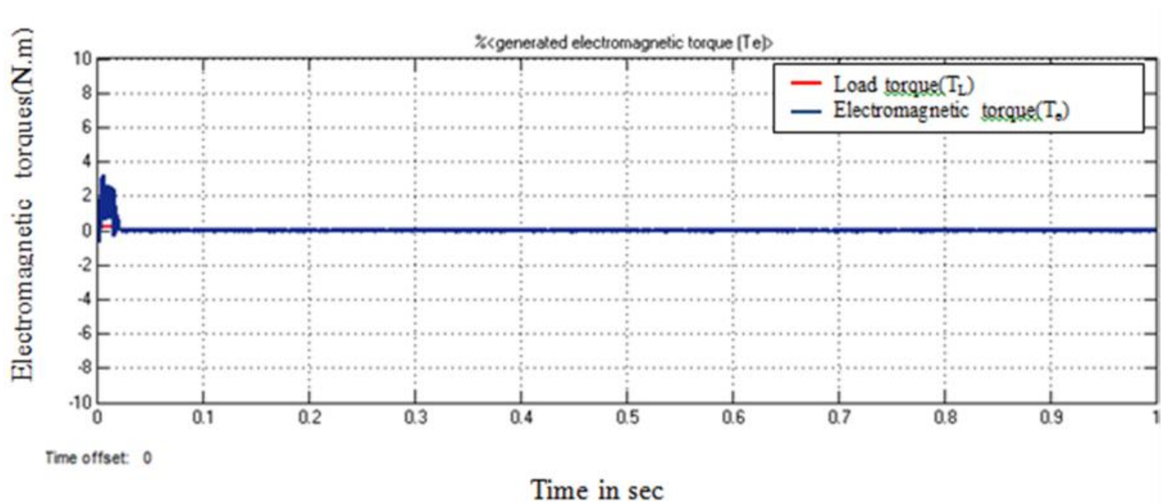

Figure 4.10: Developed electromagnetic torque for step speed input.

Figure 4.11 shows the speed response of the machine for two step speed level desired. First the desired speed is set to be 100rad/sec. Then the speed change instantly at

T=0.4sec and attain its steady speed of 120rad/sec. Figure 4.12: Shows the current drawn by the machine for the two step speed level command. At the speed change instantly from 100rad/sec and attain 120rad/sec, the phase sequence is changed.

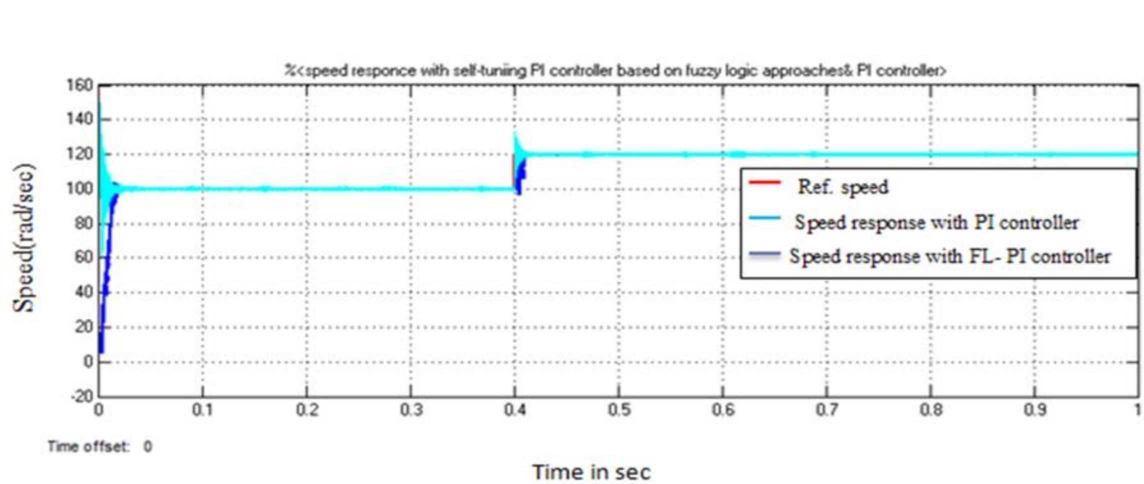

Figure 4.11: Speed response at two step speed levels with fuzzy for 0.25Nm load torque.

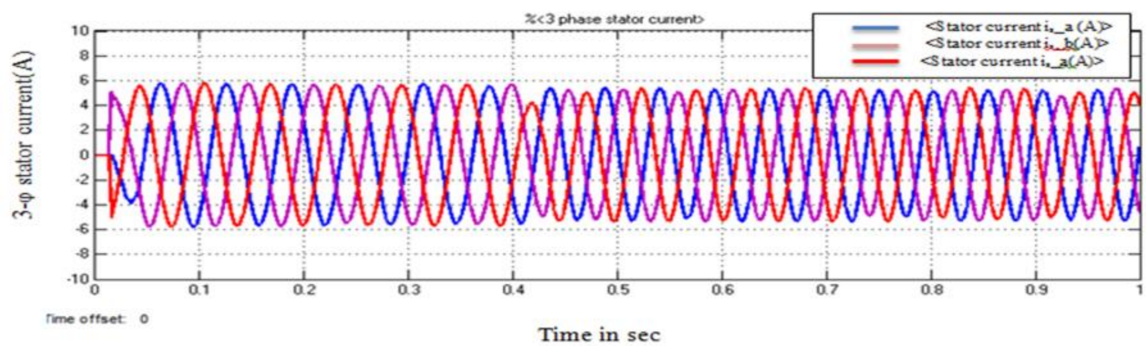

Figure 4.12: Response of 3-ϕ stator current for two ramped speed levels with 0.25Nm load is applied.

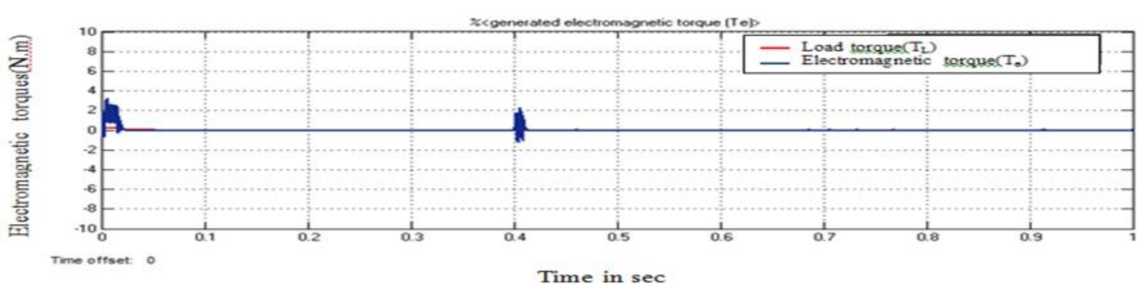

Figure 4.13: Developed electromagnetic torque for two ramped speed levels.

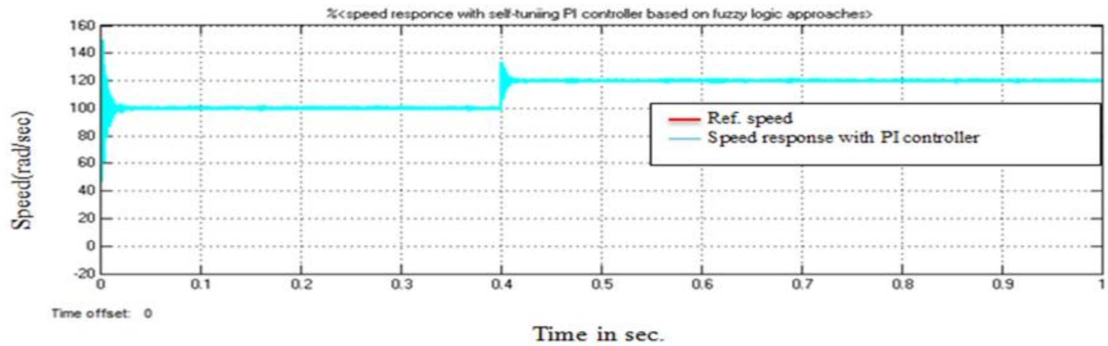

Figure 4.14: Speed response at two step speed levels without fuzzy for 0.25Nm load torque is applied

Figure 4.15 Shows the speed response at two step speed levels with fuzzy for no-load torque. The speed response does not changed with the change of reference speed trajectory for no load torque.

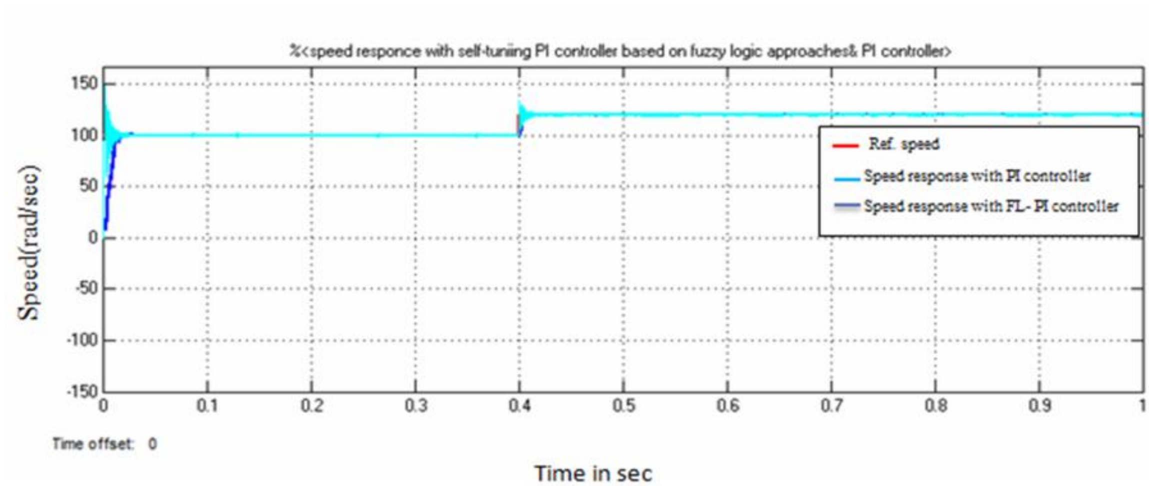

Figure 4.15: Speed response at two step speed levels with fuzzy for no-load torque.

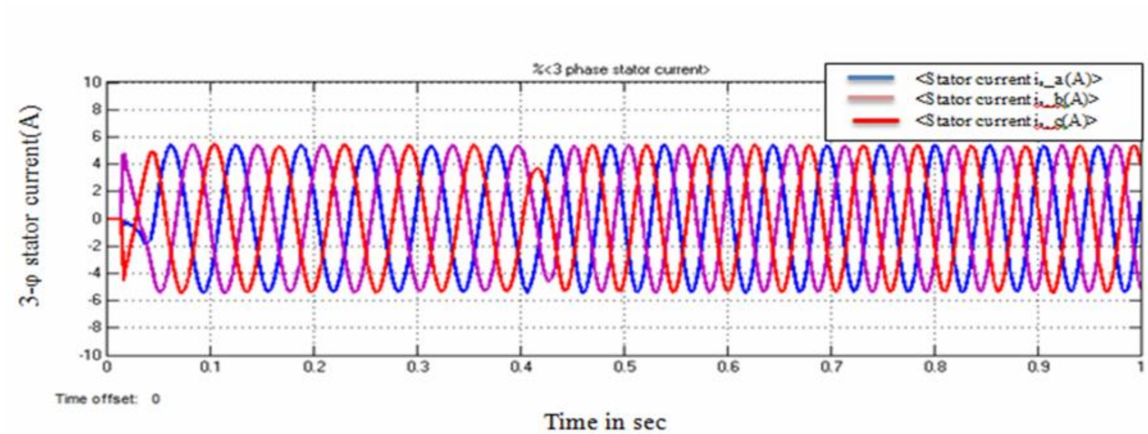

Figure 4.16: Response of 3-ϕ stator current for two steps speed levels with no- load torque.

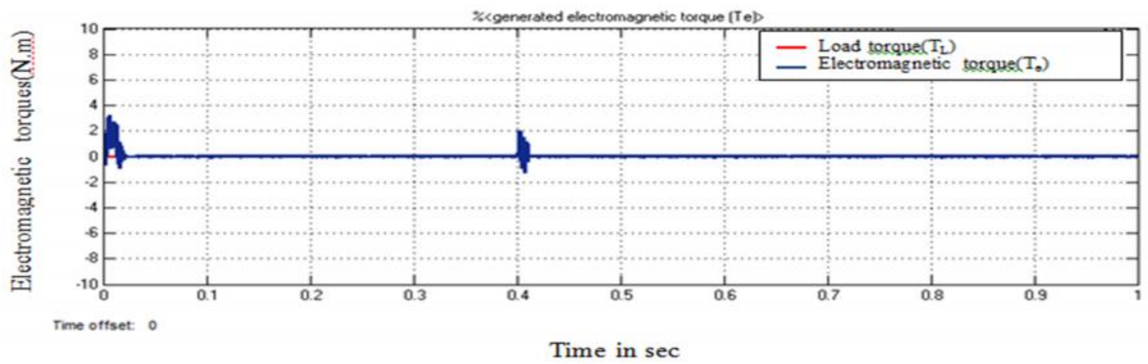

Figure 4.17: Developed electromagnetic torque for two ramped speed levels.

Figure 4.18 shows the speed response for different level of reference speeds(100rad/sec, 0rad/sec and -100rad/sec) respectively as seen from the graph. For 100rad/sec, the motor run in counter clock wise direction. 0 rad/sec means the motor is at rest or no rotation at all and for -100rad/sec the motor run in clock wise direction.

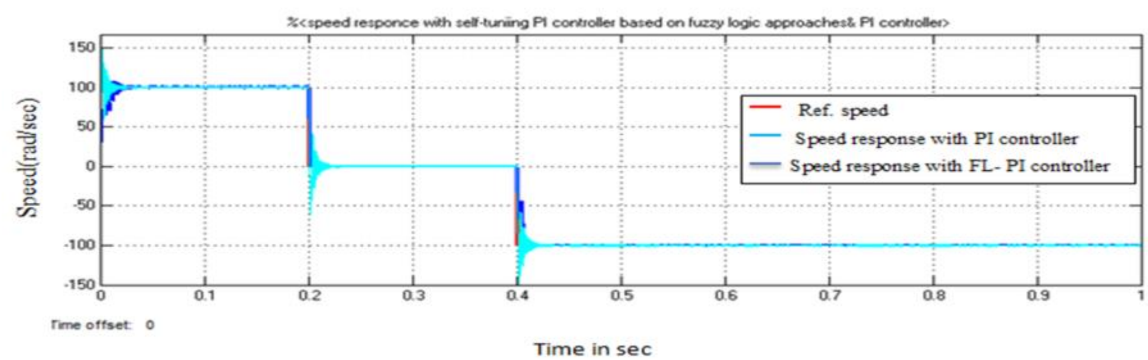

Figure 4.18: Three step level speed response of machine for 100 rad/sec and -100 rad/sec with fuzzy for 0.25Nm torque is applied.

Figure 4.19 shows the generated electromagnetic torque. At the starting point produced torque is about 3.8Nm due to the acceleration of rotor to attain steady speed of 100rad/sec. At the instant of 0.2sec, torque is overshoot to the negative to force the speed from 100rad/sec to zero and also at the instant of 0.4sec, torque is shoot to the negative to force the speed from zero to attain steady speed of -100rad/sec .

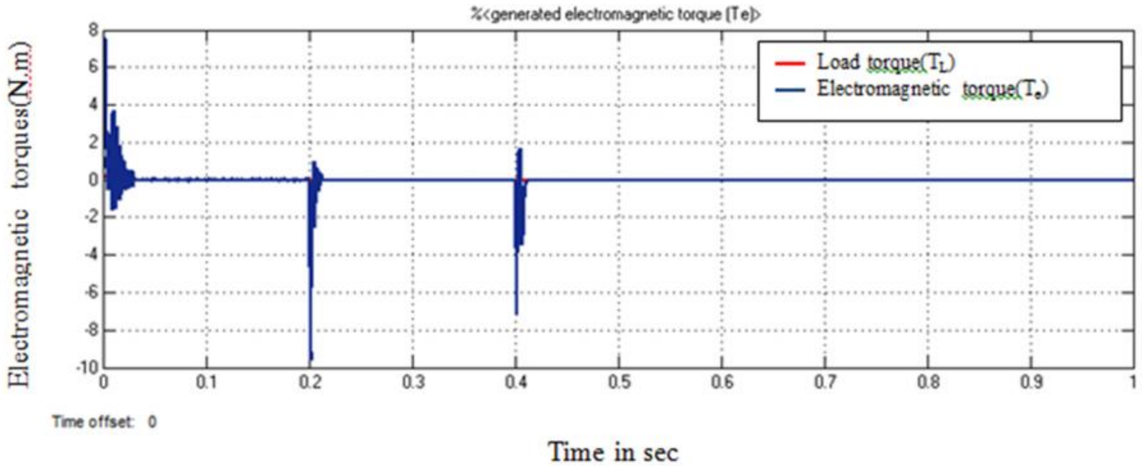

Figure 4.19: Electromagnetic torque when the machine is running at100 rad/sec, and -100 rad/sec respectively

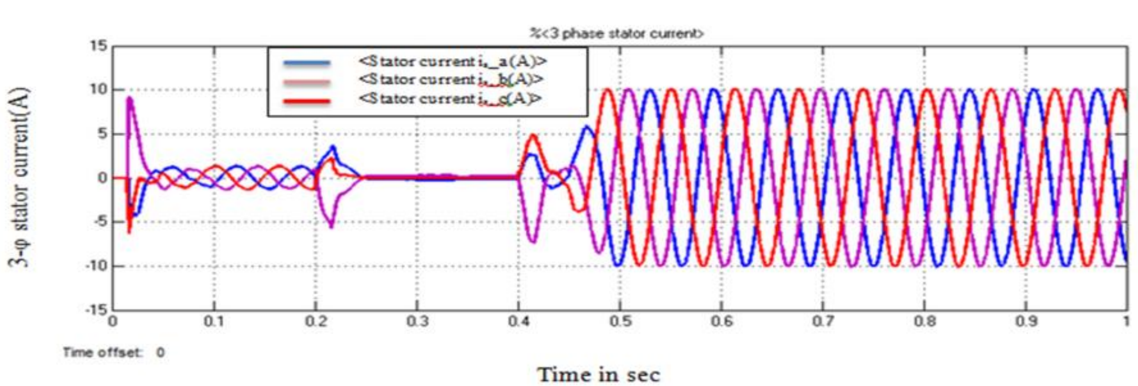

Figure 4.20: 3-phase stator current when the machine is running at100 rad/sec and -100 rad/sec

Figure 4.21 shows the speed response of permanent magnet synchronous motor for different level of reference speeds without fuzzy controller which is with 32.9% overshoots for 0.25Nm load torque is applied.

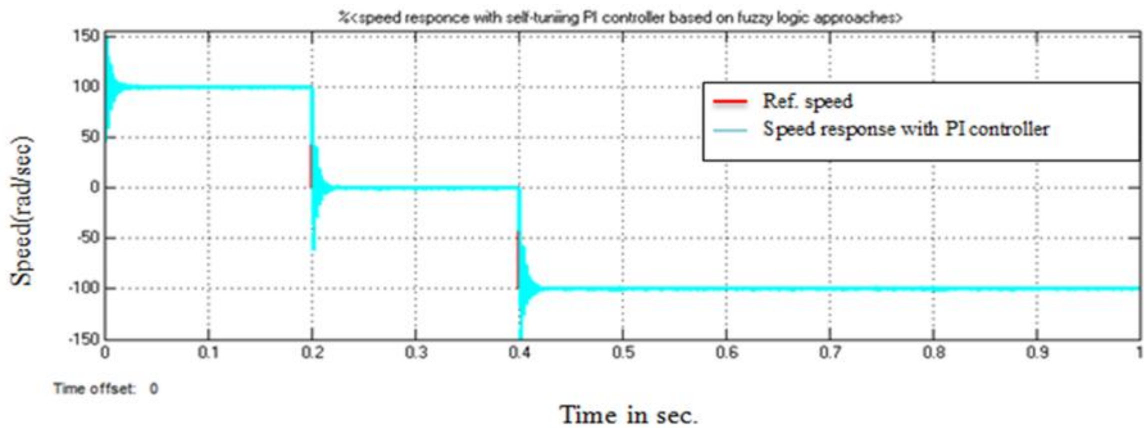

Figure 4.21: Three step level speed response of permanent magnet synchronous motor without fuzzy for 0.25Nm load torque is applied.

Figure 4.22 shows the speed response of permanent magnet synchronous motor for data reversal. Data reversal refers to the decelerations of motor from attained steady speed after certain time and attained another steady state. Figure 4.23 shows 3-ϕ stator current response of the permanent magnet synchronous motor. When the speed is de-accelerated from 100rad/sec and attained steady speed of -100rad/sec, the directions of current is changed or the three phases are reversed. Figure 4.24 shows the generated electromagnetic torque. When the speed is decelerated, the torque values are negative. at this stage the motor act as a generator.

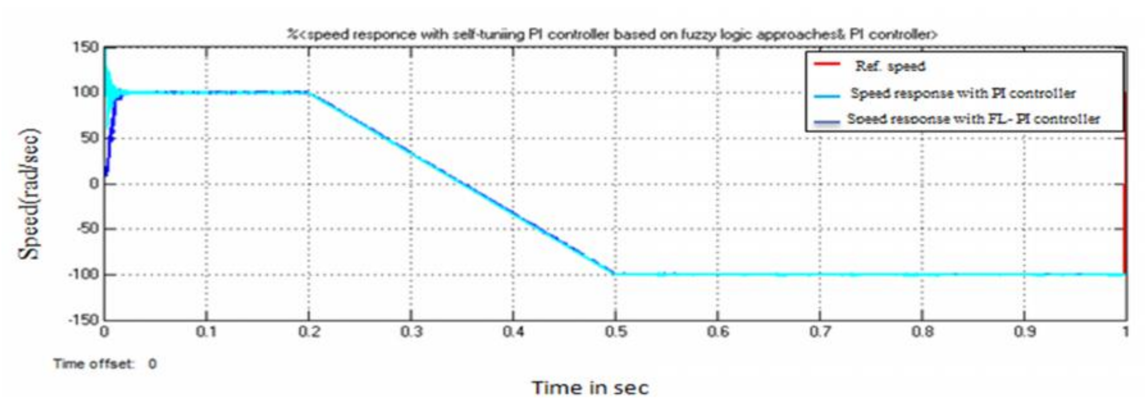

Figure 4.22: Speed response of permanent magnet synchronous motor for data reversal with fuzzy when 0.25Nm load is applied.

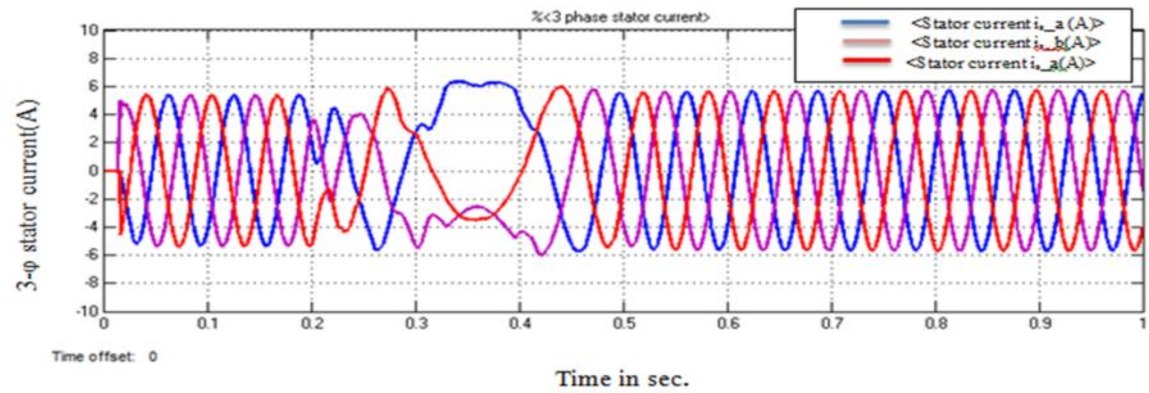

Figure 4.23: 3-ϕ stator current response of the permanent magnet synchronous motor for data reversal with fuzzy when 0.25Nm load is applied.

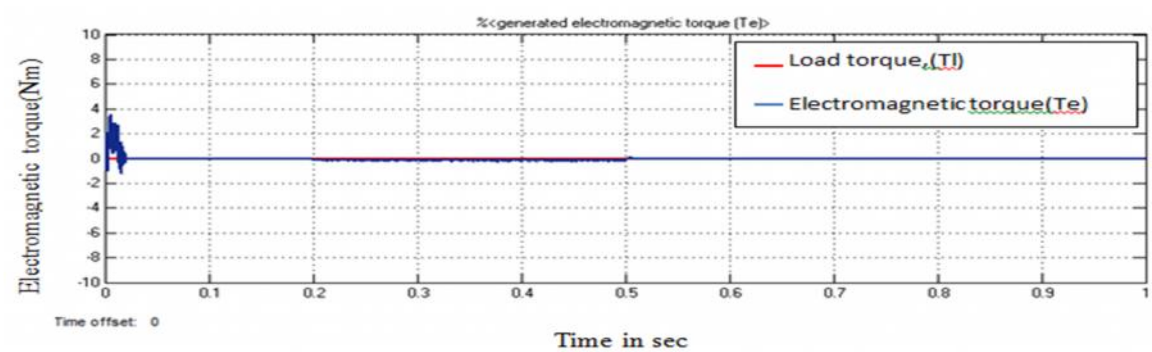

Figure 4.24: Electromagnetic torque of permanent magnet synchronous motor when data is reversed.

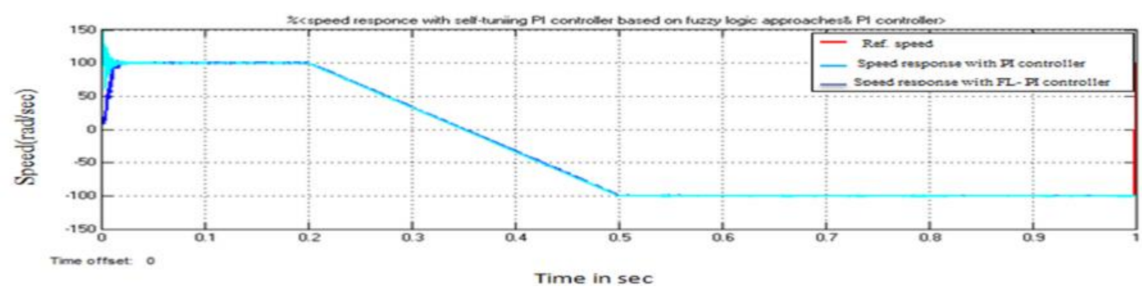

Figure 4.25: Speed response of permanent magnet synchronous motor for data reversal (when the motor is decelerate after certain time) with fuzzy when no-load is applied.

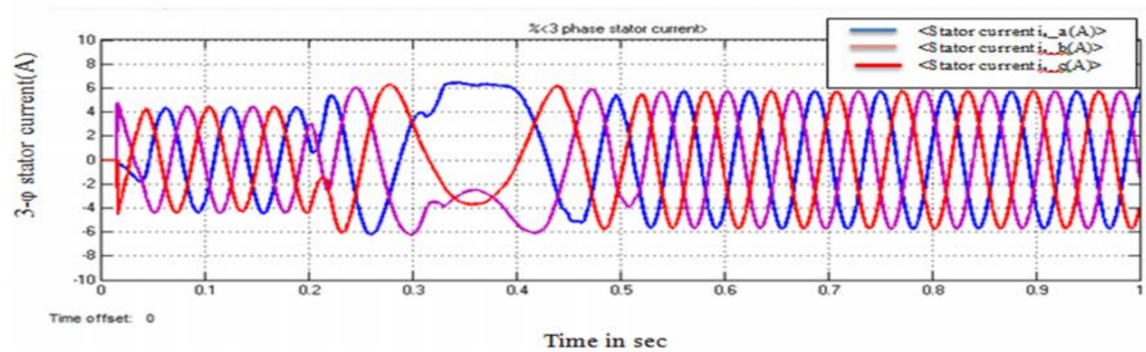

Figure 4.26: 3-ϕ stator current response of the permanent magnet synchronous motor for data reversal with fuzzy when no-load is applied**.**

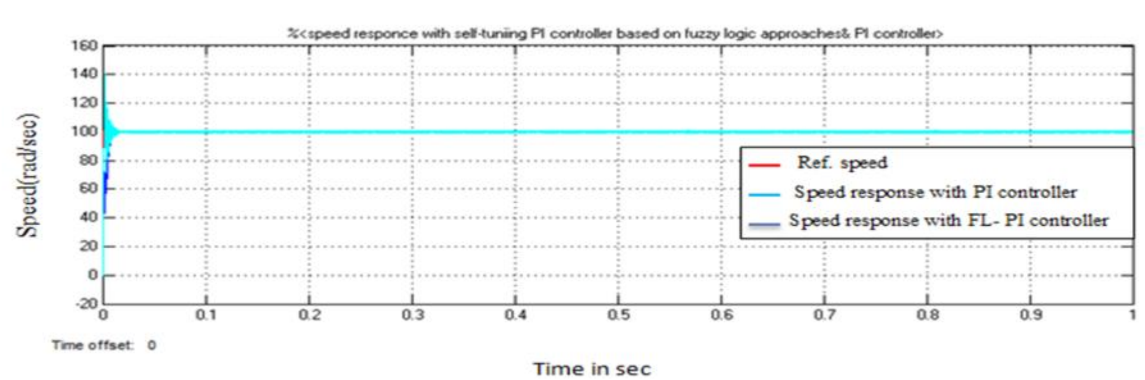

Figure 4.27: Speed response with fuzzy when Rs is increased by 50% for no-load torque.

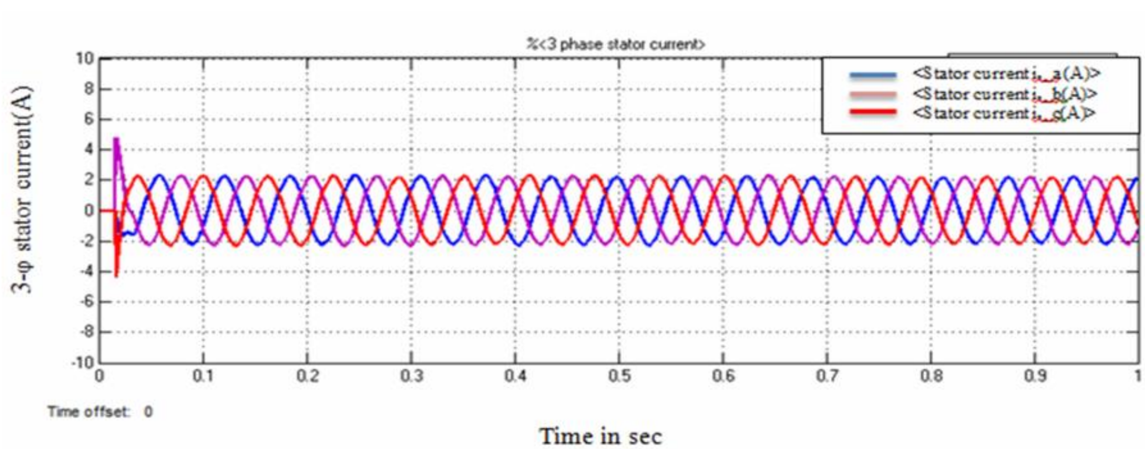

Figure 4.28: 3-ϕ stator current response of the permanent magnet synchronous motor with fuzzy for no-load is applied.

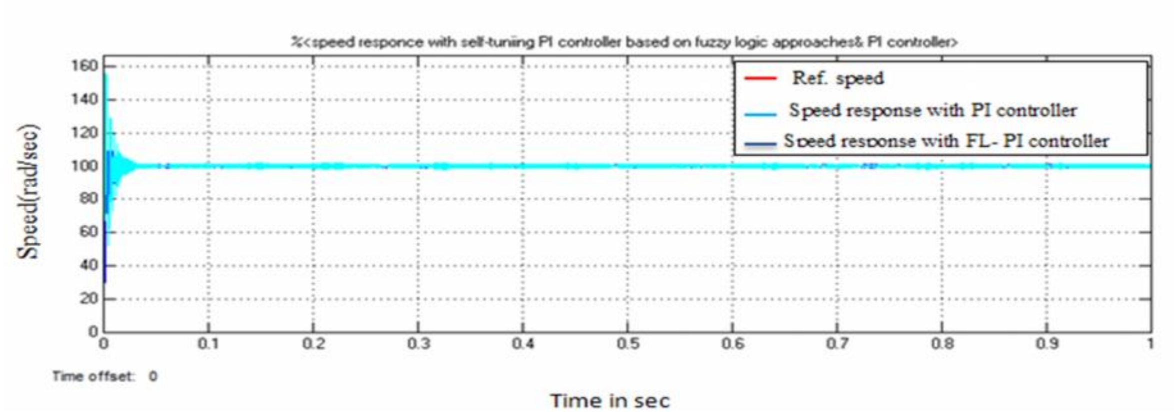

Figure 4.29: Speed response with fuzzy when inductance (L) is increased by 50%

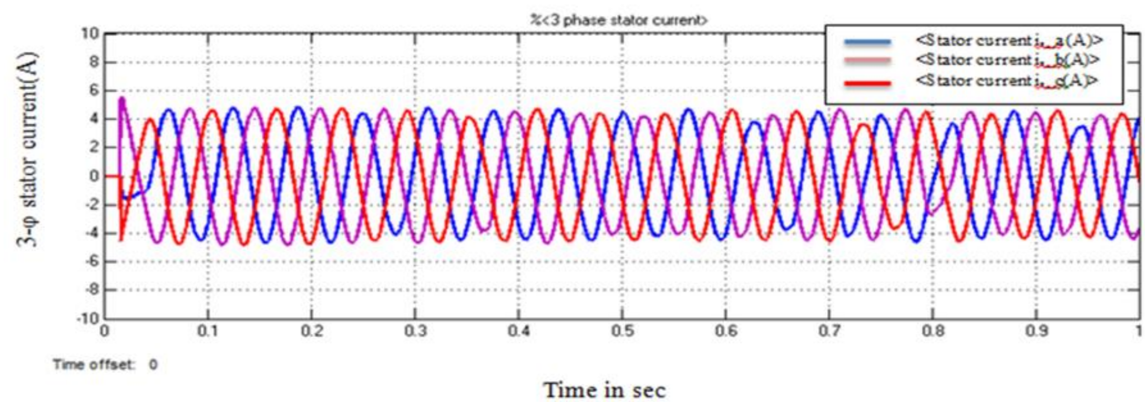

Figure 4.30: 3-ϕ stator current response of the permanent magnet synchronous motor when inductance (L) is increased by 50%.

In general, speed response is not altered with parameter variations for fuzzy logic based self-tuning PI controller. The magnitude and phase of three phase current is changed with the change of inductance values.

CHAPTER FIVE

Conclusions, recommendations and future works

5.1. Conclusions

In this thesis a complete review of permanent magnet synchronous motor speed control is presented. As the permanent magnet synchronous motor servo system is non-linear and time-varying system, the result of traditional PI controller as speed control of permanent magnet synchronous motor is not satisfactory to the higher degree of accuracy conditions. The fuzzy control not only has prominent advantage in complex, time varying and non-linear system control but also not need the mathematical model of the controlled object.

Fuzzy logic controller is proposed to tune the parameters of PI controller by developing 49-rules(7x7)member ship functions)based on the characteristics of permanent magnet synchronous motor and PI controller for speed control of sensored feedback permanent magnet synchronous motor drive system. By increasing the member ship functions, accuracy of the system is increased.

The proposed controller plays the great roles as speed controller in permanent magnet synchronous motor control system by adjusting PI-controller parameters on-line according to the speed error and change in speed error.

The complete system has been simulated using matlab/Simulink software package. The simulation is performed under different constraints, such as load torque variation and speed changes. From the simulation result we observe that, for no-load torque, over shoot is 34.85 % and settling time is around 0.018sec for PI-controller whereas over shoot is 3.38 % and settling time is around 0.0167sec with zero steady state error for proposed controller. When 0.25Nm load is applied at 0.02sec, over shoot is 32.9 % and settling time is around 0.019sec for PI-controller whereas over shoot is 3.56 % and settling time is around 0.018sec for self-tuning PI controller based on fuzzy logic approach. When the values of stator resistance and inductance is increased by 50%, speed response of the system shows 31.97% and 34.64% overshoot with conventional PI controller, where as

3.84% and 0.99% overshoot with proposed controller respectively. The proposed controller does also not altered with the complexity of reference speed trajectories.

The simulation result show that, self-tuning PI-controller based on fuzzy logic approach have improvements in terms of zero steady state error, small settling time, low over shoot, fast recovery from load changes and parameter variations compared to the traditional PI controller with fixed gains.

5.2. Recommendations

The recommendations for this thesis are,

1. Implementing FL-based self-tuning PI controller for real time processing which is capable of the DSP allows for highly reliable drives which is able to operate efficiently under wide range of speeds and which also offers the potentiality of implementing more advanced on complex control schemes high performance variable speed drives.
2. The proposed fuzzy logic based self-tuning PI speed controller may be implemented for sensor based PMSM control using the DSP board.

5.3. Suggestions for future works

The suggestions for future work are listed as follows:

1. To include neural network and genetic algorithm in FL based self-tuning PI controller.
2. For the same range of input membership functions, applying different rules for output membership functions.
3. For the different range of input membership functions, applying same rules for output membership functions.
4. For the different range of input membership functions, applying different rules for output membership functions.
5. To design fuzzy based self-tuning PD controller for speed control of permanent magnet syschronous motor.

References

[1] R. Krishnan, "*Permanent Magnet Synchronous And Brushless Dc Motor Drives*", Electrical And Computer Engineering Department, Virginia Tech Blacksburg, Virginia, U.S.A., Crc Press Taylor & Francis Group, 2010.

[2] R. Krishnan," *Electric Motor Drives Modeling, Analysis, And Control*", Virginia Tech Blacksburg, Virginia, U.S.A., 2001.

[3] Dereje Shibeshi, "*DSP Based Field Weakening Control Of PMSM*" ,Aait, School Of Graduate Studies, Department Of Electrical And Computer Eng., 2007.

[4] Rajesh Kumar, R. A. Gupta, Bhim Singh, "*Intelligent Tuned Pid Controllers For PMSM Drive - A Critical Analysis*", Department Of Electrical Engineering, Malaviya National Institute Of Technology, Jaipur-3020 17, India, Department Of Electrical Engineering, Indian Institute Of Technology, Delhi-110016, India, 2006, PP 2055-2060.

[5] Bimal K. Bose, "*Modern Power Electronic And Ac Drives*", The University Of Tennessee, Knoxville, Upper Saddle River, Nj07458, 2001.

[6] Mutasim Nour (Corresponding Author), Omrane Bouketir & Ch'ng Eng Yong, "*Self-Tuning Of PI Speed Controller Gains Using Fuzzy Logic Controller*", Modern Applied Science, November 2008, PP 55-65.

[7] Yanpeng Dou, Zhang Ze, "*Design And Realization Of Fuzzy Self-Tuning PID Speed Controller Based On Tms320f2812 DSPS*", International Conference On Mechatronics AndAutomation, August , 2007, PP 3316-3320.

[8] Limei Wang, Mingxiu Tian And Yanping Gao, "*Fuzzy Self-Adapting PID Control Of PMSM Servo System*", School Of Electrical Engineering, Shenyang University Of Technology, Shengyang, China, 2007, PP 860-863.

[9] Joe Sung Yu, Sun Mo Hwang, Chung-Yuen Won, "*Performance Of Fuzzy-Logic-Based Vector Control For Permanent Magnet Synchronous Motor Used In Elevator Drive System",* The 30th Annual Conference Of The IEEE Industrial Electronics Society, 2004, PP 2679-2683.

[10] Chunhua Zang, "*Vector Controlled PMSM Drive Based On Fuzzy Speed Controller*" ,2nd International Conference On Industrial Mechatronics And Automation, 2010, PP 199-202.

[11] Y.S. Kung And P.G. Huang, "*High Performance Position Controller For PMSM Drives Based On Tms320f2812 DSP*", 2004 IEEE CCA/ISIC/CACSD Joint Conferences, Cca Proceedings, Vol.I, September , 2004, PP. 290-295.

[12] Enrique L. Carrillo Arroyo, "*Modelling And Simulation Of Permanent Magnet Synchronous Motor System*", Universiy Of Puerto Rico Mayaguez Campus, 2006.

[13] HENOK BEREHANU "*DSP Based Implementation Of Field-Weakening On Synchronous Motor For High Speed Operation*", AAIT, School Of Graduate Studies, Department Of Electrical And Computer Engineering, January, 2011.

[14] R. Krishnan, Electric Motor Drives Modeling, Analysis and Control, Prentice Hall 2001.

[15] International Journal Of Advanced Research In Electrical, Electronics And Instrumentation Engineering. Vol. 3, Issue 10, October ,2014.

[16] R. Krishnan, "*Electric Motor Drives Modeling, Analysis, And Control*" Virginia Tech Blacksburg, Virginia, U.S.A., 2001.

[17] Ludovic A. Chretien, "*Position Sensor less Control Of Non-Salient Permanent Magnet Synchronous Machines*", The Graduate Faculty Of The University Of Akron, May, 2006.

[18] P. Pillay And R. Krishnan, "*Modelling Of Permanent Magnet Motor Drives*" ,Industrial Electronics, IEEE Transactions On, Vol. 35, 1988, PP. 537-541.

[19] By Enrique L. Carrillo Arroyo, "*Modeling And Simulation Of Permanent Magnet Synchronous Motor Drive System*", Electrical Engineering University Of Puerto Rico Mayagüez Campus, 2006.

[20] International Journal Of Advanced Research In Electrical, Electronics And Instrumentations Engineering Vol.2.Issued,January, 2013.

[21] By Mahlet Legesse "*Speed Control Of Vector Controlled Pmsm Drive Using Fuzyy Logic-PI Controller*", AAIT, School Of Graduate Studies, Department Of Electrical And Computer Engineering, Augest ,2011.

[22] Jin-Woo Jung, "*Space Vector PWM Inverter*", Ph.D., Thesis, Department Of Electrical And Computer Engineering, The Ohio State University, Feb.20, 2005.

[23] By Jin-Woo Jung,Ph.D Student "*Project#2space Vector PWM Inverter*" ,Mechatronics System Laboratory, Department Of Electrical And Computer Engineering,The Ohio State University, February 20, 2005.

[24] Erwan Simon, "*Implementation Of A Speed Field Oriented Control Of 3-Phase PMSM Motor Using Tms320f240*", Texas Instruments Application Report Spra588.

[25] Leonid Reznik, "*Fuzzy Controllers*" Victoria University Of Technology, Melbourne, Australia, Reed Elsevier PLC Group, 1997.

[26] Limei Wang, Mingxiu Tian And Yanping Gao, "*Fuzzy Self-Adapting PID Control Of PMSM Servo System*", School Of Electrical Engineering, Shenyang University Of Technology, Shengyang, China, 2007, PP 860-863.

[27] Manoj Kumar Pandey, "Analysis Of Self Tuning Fuzzy PID Internal Model Control", Msc Thesis, Thapar University, Patiala, July, 2010.

[28] Education For Engineers, "*Fundamentals Of PID Control*" ,Pdhengineer.Com, 2003.

[29] Introduction to Fuzzy Logic using MATLAB, Library of Congress Control Number: 2006930099, ISBN-10 3-540-35780-7 Springer Berlin Heidelberg New York.

[30] Leonid Reznik, "*Fuzzy Controllers*", Linacre House, Jordan Hill, Oxford, 1997.

[31]Timothy J. Ross, "*Fuzzy Logic With Engineering Applications*", John Wiley & Sons Ltd, England, 2004.

[32] Jan Jantzen, "*Design Of Fuzzy Controllers*", Technical University Of Denmark, Department Of Automation, Denmark, Aug, 1998.

[33] Kyoung K. A., Bao K. N., Yoon H. S. "*Self-Tuning Fuzzy PID Control For Hydraulic Load Simulator*", Int. Conference On Control, Automation, And Systems , Ceox, Seoul, Korea., 2007, PP 345-349.

[34] Kyoung K. A. Bao K. N. "*Position Control Of Shape Memory Alloy Actuators Using Self Tuning Fuzzy PID Controller*",. Int. J. Of Control, Automation, And Systems,Vol.4 ,2006, PP 756-762.

[35] Timthy J.Ross, "*Fuzzy Logic With Engineering Applications*" University Of New Mexico, John Wiley & Sons, Ltd, 2004.

[36] International Journal of Advanced Research in Electrical, Electronics and Instrumentation Engineering on "*Field Oriented Control of a Permanent Magnet Synchronous Motor using a DSP*", *(An ISO 3297: 2007 Certified Organization)* Vol. 3, Issue 10, October ,2014.

[37] M. G.uzelkaya, I.Eksin, E. Yes, "*Self-tuning of PID-type fuzzy logic controller coefficients via relative rate observer*", Science Direct Engineering Applications of Artificial Intelligence 16 ,2003 ,PP 227–236.