



Addis Ababa University

Addis Ababa Institute of Technology (AAiT)

School of Electrical and Computer Engineering

**Speed Control of Permanent Magnet Synchronous Motor  
Using Higher Order Sliding Mode Controller**

A thesis submitted to Addis Ababa Institute of Technology, School of  
Graduate Studies, Addis Ababa University

In partial fulfillment of the requirement for the Degree of Master of Science in Electrical  
Engineering (**Electrical Control Engineering**).

**By: Derege Eskezia**

**Advisor: Dr. Mengesha Mamo**

**Co-Advisor: Mr. Andinet Negash**

May, 2016

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By

Derege Eskezia

**Approval by Board of Examiners**

\_\_\_\_\_  
Chairman, Dept. Graduate Committee

\_\_\_\_\_  
Signature


Dr. Ing. Mengesha Mamo

Advisor's Name

\_\_\_\_\_  
Advisor Signature

Mr. Andinet Negash

Co-advisor's Name

4/30/2016 

Co-advisor Signature

Dr. Dereje Shiferaw

Internal Examiner

\_\_\_\_\_  
Signature

Mr. Getu Gabissa

External Examiner

\_\_\_\_\_  
Signature

## ACKNOWLEDGMENT

First and foremost, I would like to express my deepest thanks and gratitude to my advisor Dr. Mengesha Mamo and co-advisor Ato Andinet Negash for their invaluable advices, comments, encouragement, continuous guidance and supervision, their friendship, remarkable patience and caring support during my graduate studies.

I do want to thank Ato Shimelis Demssie for guiding on the design and implementation of space vector pulse width modulation (SVPWM). In addition I would like to say thanks to different authors who I mentioned in the references, and companies like Texas Instruments.

My thanks are extended to my fellow colleagues in Electrical Engineering, especially in the control engineering stream, Lijaddis Getnet, Fahmi Mohammed, Asalifew Belachew, and Abere Getachew who built an academic and very friendly study environment that made my study at the university most enjoyable and skillful.

Last but not least, I am always indebted to my girlfriend, Bezawit Mekonnen, my father, Eskezia Lakew and my family members for their endless support and love throughout these years. They gave me additional motivation and determination during my graduate study.

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Derege Eskezia

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## LIST OF ACRONYMS

AAiT	Addis Ababa Institute of Technology
AC	Alternating current
d, q	synchronous reference frame quantities (d-axis, q-axis)
DC	Direct current
DSPs	digital signal processors
DTC	Direct Torque Control
EMF	Electro Motive Force
ESO	extended state observer
FOC	field oriented control
HOSM	higher order sliding mode
HOSMC	high -order sliding mode control
IM	induction motor
IPark	inverse Park
ISMC	Integral sliding mode controller
MIMO	multiple input multiple output
mmf	magneto-motive force
MRAS	Model Reference Adaptive Systems
PI	Proportional Integral
PM	permanent-magnet
PMSM	permanent magnet synchronous motor
PWM	Pulse Width Modulation
QEP	Quadrature Encoder Pulse
RPM	revolution per minute
SMC	sliding mode control
SRM	switched reluctance machines
SVPWM	Space vector pulse width modulation
THD	total harmonic Distortion
V/F	volt per frequency
WECSs	wind energy conversion systems

### ABSTRACT

Permanent magnet synchronous motor (PMSM) is the most used drive in machine tool servos and other modern high performance applications. In this thesis, a quasi-continuous higher order sliding mode controller is designed for PMSM in order to regulate the desired reference speed with robustness and minimized chattering. Proportional, Integral and derivative (PID) controllers are also designed for direct and quadrature components of the current in order to control the flux and torque.

For the investigation of the problem, the controller is tested through a reference speeds as well as with load variations through simulation using MATLAB/SIMULINK. In the simulation, the SVPWM algorithm was written in M-file and embedded on simulation block. To make system control simple, field oriented control (FOC) is used for the operation of the motor to control the torque and flux of the motor directly as DC motor.

Simulation results demonstrate that the speed response of the proposed drive is unaffected by load torque and parameters variation. Furthermore, the developed controller has good performance when the reference speed is changed, i.e., the developed controller speed regulation is unaffected by the nature of the reference speed. From the simulation result, the speed error in steady state response is zero, the speed response has a maximum over shoot of 1.08% and the chattering effect on the system is reduced.

**Keywords:** PMSM, HOSM controller, FOC, quasi-continuous HOSM controller, chattering effect.

# CHAPTER 1

## INTRODUCTION

### 1.1 BACKGROUND

Due to the convenience of torque and speed control, DC electric machine systems had been adopted in a variety of industrial applications for more than 100 years. During the past 30 years, with the development of power electronics, digital signal processors (DSPs), and computer-aided design technologies, AC motors have replaced DC motor and have become dominant in variable-frequency motors applications. Currently, various types of AC motors like induction motor (IM), PMSM, switched reluctance machines (SRM), etc., are widely used in industrial applications [1]-[3].

Among the AC motors, PMSM systems have been used more and more in many industrial applications, e.g., home appliances [4], electric-drive vehicle systems [5], and wind energy conversion systems (WECSs) [6], due to their distinctive advantages of high efficiency, high power density, and wide constant power region [25]. With the continuous reduction in the cost of permanent-magnet (PM) materials and the development of control techniques, PMSMs have become more attractive and competitive [7]. Moreover, due to worldwide concerns over environmental problems and a possible energy crisis, much effort from both academia and industry has gone into the development of renewable energy conversion systems and electric-drive vehicles, creating a large market for PMSM technologies.

To get fast four-quadrant operation, good acceleration and smooth starting, the field-oriented control or vector control is used in the design of PMSM drives [8]. Nevertheless, the high performance control necessitates accurate rotor position information. That can be realized by using position sensors "encoders, tachometers and resolvers". Even if, this mechanical sensor increases the size of the machine, it measures accurate rotor speed and position of the PMSM.

However, the PMSM is a typical high nonlinear, multivariable coupled system, and its performance is sensitive to external load disturbances, parameter variations in the plant, unmodelled and nonlinear dynamics [8].

To achieve good dynamic response, some robust control strategies such as nonlinear control, adaptive control, and sliding mode control (SMC) have been developed [8].

The SMC is a powerful nonlinear control technique and has been widely used for speed and position control of PMSM because it provides a fast dynamic response, and is not sensitive to external load disturbance and parameter variations. In addition, the appropriate sliding surface can be selected to reduce order for control system. Usually, the design of sliding surface for a PMSM is limited to integer-order, which means the sliding surface is constructed by the integer order integration or differentiation of the state variables. However, due to the chattering phenomena of sliding mode control, the high frequency oscillation of control system brings challenge for the application of sliding mode control. On the other hand, the choice of sliding surface strictly requires system relative degree to equal to 1, which limits the choice of sliding surface. In order to solve the above problems, a new type of sliding mode control, that is, higher order sliding mode control is introduced. The technology not only retains advantage of strong robustness in the traditional sliding mode control, but also enables discontinuous item variables transmit into the first order or higher order sliding mode derivative to eliminate the chattering. Besides, the design of the controller no longer must require relative degree to be 1. Therefore, it is greatly simplified to design parameters of sliding mode surface.

### **1.2 STATEMENT OF THE PROBLEM**

Recently, permanent magnet synchronous motors (PMSMs) are increasingly used in high performance variable speed drives of many industrial applications. This is because the PMSM has many features, like high efficiency, compactness, high torque to inertia ratio, rapid dynamic response, simple modeling and control, and maintenance-free operation [1], [3], [17].

But the dynamic model of a PMSM is highly nonlinear because of the coupling between the motor speed and the electrical quantities, i.e. the d, q currents. Previously, many different control algorithms have been used to improve the performance of the PMSM. For example, as the dynamic model of the machine is nonlinear, a natural approach is the exact feedback linearization control method, by which the original nonlinear model can be transformed into a linear model through proper coordinate transformation. However, in general, the dynamics of the synchronous motors may not be fully known, since some of parameters appearing in the

equations will vary. For instance, the resistance and inductance will be changed when the temperature alters. As a consequence, nonlinearities can only be partially cancelled by the feedback linearization technique, and parameters uncertainties act on the equations of the motion. Then an important aim of the control design is to develop a robust controller which ensures good dynamic performances in spite of parameters uncertainties and perturbation [22].

To obtain high dynamic performance on PMSMs and actual position of the rotor, variable voltage and variable frequency operation has to be performed by measuring speed of rotation and stator currents through mechanical sensors and feedback them to the controllers. The cost of the additional sensors and timing hardware to produce a sensor-based PMSMs is minimal compared to the tremendous benefits they provide for the performance and efficiency of the system. Since inaccurate rotor position information will not only degrade the control performance but also cause instability in the control system. In addition to this, the sliding mode control is known to be a robust approach to solve the control problems of nonlinear systems. Robustness properties against various kinds of uncertainties such as parameter perturbations and load torque disturbances can be guaranteed.

But, conventional sliding mode controller has some disadvantages such as chattering phenomenon. Therefore to solve the problem of a SMC, this thesis focuses on applying higher order sliding mode controller.

This thesis presents a comprehensive study of the higher order sliding mode methods of speed control for PMSM, which can assure robustness against a load torque disturbance and parameter variation with chattering reduction.

### **1.3 OBJECTIVES OF THE STUDY**

The primary objective of this thesis is to model, design, simulate and implement a speed control of PMSM using higher order sliding mode controller. Specifically,

- To study the mathematical model of permanent magnet synchronous motor.
- To model and design a robust speed controller of the PMSM using higher order sliding mode controller with chattering reduction.
- To simulate the system using MATLAB/Simulink.

## 1.4 CONTRIBUTION OF THE THESIS WORK

The contribution of this work is to model and design a robust sensor based control scheme such that the motor speed tracks a desired reference by using mechanical position /speed sensor and the measurement of the motor currents. The controller is a quasi-continuous high order sliding mode controller.

Sliding mode control approaches is recognized as one of the efficient tools to design robust controller for complex higher order nonlinear dynamic plant operating under uncertain condition [15], [22], [27].

The major advantage of sliding mode control is low sensitivity to plant parameter variation and disturbance which eliminates the necessity of exact modeling [4], [27]. Sliding mode control enables the decoupling of overall system motion into independent partial components of a lower dimension and, as a result, reduces complexity of feedback design. In addition, high -order sliding mode control (HOSMC), which inherits advantages of the standard SMC, also reduce and (or) remove the chattering effect [9], [11], [15]. Moreover, better accuracy can be achieved by using HOSMC [12], [22].

Higher order sliding mode control based speed control of the PMSM advantages of the system under study has improved static and dynamic performance, high speed control accuracy, bounded and assures robustness against a continuous disturbance with chattering elimination.

Specifically the contribution of this thesis is:

- Designing a speed control of a PMSM using HOSM controller to reduce the chattering effect.

## 1.5 METHODOLOGY

For the accomplishment of speed controlling of PMSM using HOSM controller, the following methodologies have been followed:

- Gathering & survey related works from various literature and publication.
- Studying the mathematical model of the system.

- Designing a HOSM Controller for the system.
- Analyzing the system with Matlab/Simulink simulation.

### **1.6 SCOPE AND LIMITATION**

The scope of this thesis is studying, simulating and designing of the speed control of the permanent magnet synchronous motor using higher order sliding mode control scheme and the mechanical position/speed sensor which is used for measuring instantaneous mechanical position/speed of the rotor.

In our Laboratory, the Motor has no parameter specification labeled and I am taking the values from the Hamdihun's paper [24] that was done on the motor which is found on the lab currently. The other constraint is the inaccessibility of the mechanical position/speed sensor.

### **1.7 OUTLINE OF THE THESIS**

This thesis includes five chapters. The first chapter generally presents the introduction of the speed control of PMSM using HOSM controller which is the background of the PMSM speed control, statement of the problem, relevance, objectives of the study, methodology, scope and limitation of the study and different literatures, related to speed control of PMSMs, are reviewed.

In chapter two, it presents the different control theory techniques of the PMSM. System modeling, designing, analysis and implementation are presented in chapter three. Simulation results are presented and discussed in chapter four. The contributions of the thesis work are also discussed in the same chapter. Finally, chapter five presents conclusions and recommendations.

## CHAPTER 2

### COMMON CONTROL SCHEMES FOR PMSM

#### 2.1 INTRODUCTION

A Permanent Magnet Synchronous Motor (PMSM) is a synchronous motor that uses permanent magnets for the rotor to produce the air gap magnetic field rather than using electromagnets. It has a multiphase stator and the stator electrical frequency is directly proportional to the rotor speed in the steady state. However, it differs from a traditional synchronous machine in that it has permanent magnets in place of the field winding and otherwise has no rotor conductors. The use of permanent magnets in the rotor enhances efficiency, eliminates the need for slip rings, and eliminates the electrical rotor dynamics that complicate control (particularly vector control). The combination of an inner permanent magnet rotor and outer windings offers the advantages of low rotor inertia, efficient heat dissipation, reduces the motor size, high power density, compact structure, high air-gap flux, no conversion spark, high torque inertia ratio and high efficiency over other kinds of motors [1] [17].

There are different techniques of speed control of permanent magnet synchronous motors to regulate these motor drives in high-performance applications. Depending on the requirements of each application, PMSM control techniques can be divided into scalar and vector control. Scalar control is based on relationships valid in steady state. Amplitude and frequency of the controlled variables are considered. In vector control amplitude and position of a controlled space vector is considered. These relationships are valid even during transients which is essential for precise torque and speed control [16] [19]. Some of the control techniques are shown in the figure 2.1 below.

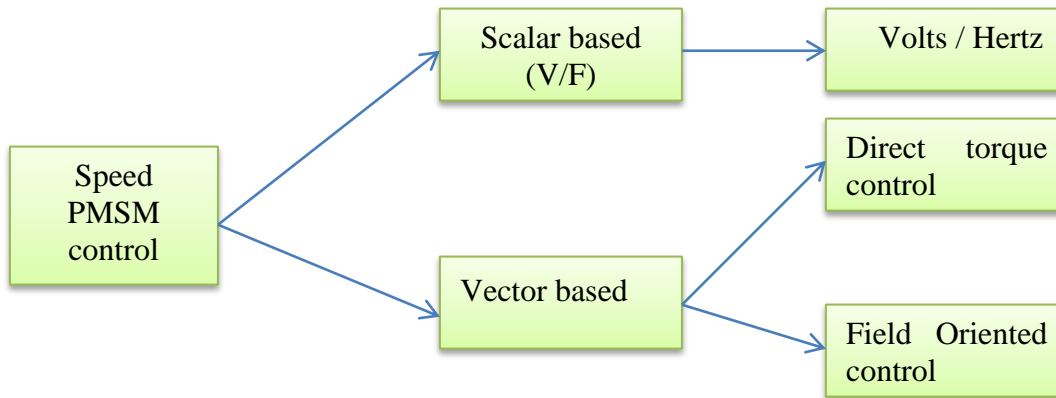


Figure 2.1: Different techniques of PMSM

## 2.2 SCALAR CONTROL

Scalar control is based on relationships valid in steady state. Only magnitude and frequency of voltage and current are controlled. Scalar control is used e.g. where several motors are driven in parallel by the same inverter [19], [24].

The main idea of this method is variation of the supply voltage frequency abstractedly from the shaft response (position, angular speed). The magnitude of the supply voltage is changed according to the frequency in a constant ratio. Then the motor is in the condition, where the magnetic flux represents the nominal value and the motor is neither overexcited nor under excited. The biggest advantage of this simple method is running in a sensorless mode because the control algorithm does not need information about the angular speed or actual rotor position and which is suitable for low-cost drive systems. On the contrary, the big disadvantages are the speed dependence on the external load torque, mainly for IM, and reduced dynamic performances [18].

### 2.2.1 Volts/Hertz control

Volts/Hertz control is among the simplest control schemes for motor control. The control is an open loop scheme and does not use any feedback loops. Using this technique for synchronous motors with permanent magnets offers a big advantage of sensorless control. Information about the angular speed can be estimated indirectly from the frequency of the supply voltage. The angular speed calculated from the supply voltage frequency according to equation 2.1 can be

considered as the value of the rotor angular speed if the external load torque is not higher than the breakdown torque [18].

The mechanical synchronous angular speed  $\omega_s$  is proportional to the frequency  $f_s$  of the supply voltage [18]

$$\omega_s = \frac{2\pi f_s}{p} \quad (2.1)$$

where  $p$  is the number of pole pairs.

### 2.3 VECTOR CONTROL

The problem with scalar control is that motor flux and torque in general are coupled. This inherent coupling affects the response and makes the system prone to instability if it is not considered. In vector control, not only the magnitudes of the stator and rotor fluxes are considered but also their mutual angle. Generally, the vector control of PMSM allows separate closed loop control of both the flux and torque, hence, achieving a similar control structure to that of a separately excited DC machine [10], [17], [19].

#### 2.3.1 Direct Torque Control

The principle of Direct Torque Control (DTC) is to directly select voltage vectors according to the difference between reference and actual value of torque and flux linkage. Torque and flux errors are compared in hysteresis comparators.

Advantages of DTC are low complexity and that it only needs to use of one motor parameter, the stator resistance. The system possess good dynamic performance but shows quite poor performance in steady state since the crude voltage selection criteria give rise to high ripple levels in stator current, flux linkage and torque. Its simplicity makes it possible to execute every computational cycle in a short time period and use a high sampling frequency [16] [19].

#### 2.3.2 Field Oriented Control (FOC)

In order to achieve better dynamic performance, a more complex control scheme needs to be applied to control the PM motor. With the mathematical processing power offered by the

microcontrollers, advanced control strategies can be implemented, which uses mathematical transformations in order to decouple the torque generation and the magnetization functions in the PM motors. Such decoupled torque and magnetization control is commonly called rotor flux oriented control, or simply FOC.

The vector control of currents and voltages results in control of the spatial orientation of the electromagnetic fields in the machine and has led to the name field orientation. FOC usually refers to controllers which maintain a  $90^\circ$  electrical angle between rotor and stator field components [16] [19] [20].

The FOC consists of controlling the stator currents represented by a vector. This control is based on projections that transform a three phase time and speed dependent model into a two coordinate (d and q coordinates) time invariant model. These projections lead to a structure similar to that of a DC machine control. FOC machines need two constants as input references: the torque component (aligned with the q co-ordinate) and the flux component (aligned with d co-ordinate). As FOC is simply based on projections, the control structure handles instantaneous electrical quantities. This makes the control accurate in every working operation (steady state and transient) and independent of the limited bandwidth mathematical model.

Thus, the advantages of FOC:

- Transformation of a complex and coupled AC model into a simple linear model
- Independent control of torque and flux, similar to a DC motor
- Fast dynamic response and good transient and steady state performance
- High torque and low current at startup
- High Efficiency
- Wide speed range through field weakening

Because of the above advantages and others, this thesis is done based on the FOC techniques of speed control of PMSM. FOC technique which involves the three reference frames and needs transformations from one to the other are [21]:

1. Stator reference frame (a, b, c) in which the a, b, c are coplanar, at 120 degrees to each other.

2. An orthogonal reference frame ( $\alpha, \beta$ ) in the same plane as the stator reference frame in which the angle between the two axes is 90 degrees instead of 120 degrees. The alpha - axis is aligned with a-axis in the second frame.
3. Rotor reference frame (d, q), in which the d axis is along the N and S poles or along the flux vector of the rotor and the q-axis is at 90 degrees to the d-axis.

Figure 2.2 shows the transformations done for decoupling the stator currents into the torque producing ( $I_q$ ) and flux producing ( $I_d$ ) component.

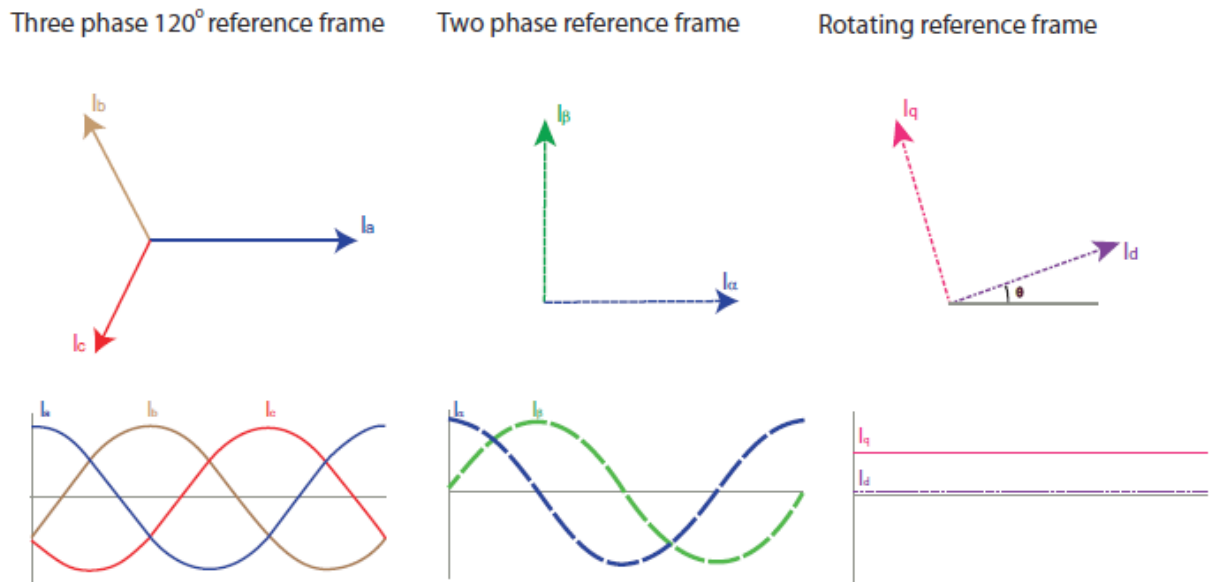


Figure 2.2: Transformations and Reference [21]

Once the torque and flux producing components are controlled with the PID controller then the controlled outputs, which are the voltages, are transformed back (inverse transformation) to the stator reference frame.

### Transformations

Transforming from a stator reference frame of a PMSM into a rotary reference frame is used to get the time invariance values of the motor parameters. In FOC, the components  $I_q$  and  $I_d$  are measured in the rotating reference frame. Hence the measured stator currents have to be

transformed from the three phase time variant stator reference frame to the two axis rotating d-q rotor reference frame. This can be done in two steps as shown in Figure 2.3.

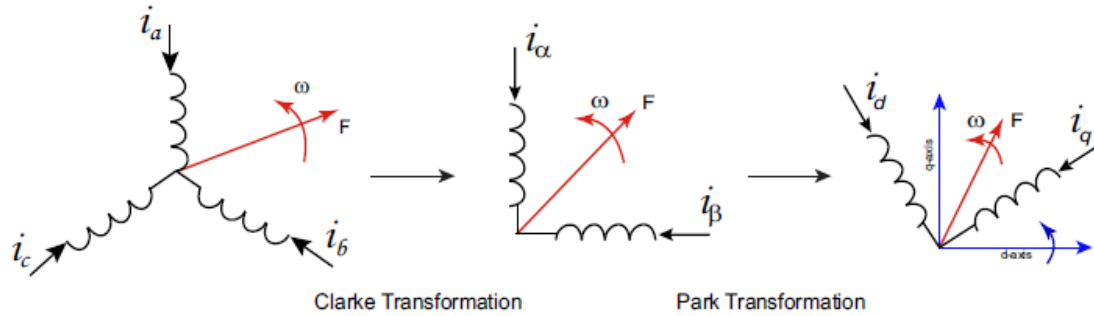


Figure 2.3: Forward Transformations [21]

- a, b, c to  $\alpha$ - $\beta$  (Clarke transformation): The measured motor currents are first translated from the 3-phase reference frame to the two axis orthogonal reference frame. The transform is expressed by the following equations [1] [21].

$$\begin{aligned} i_{\alpha} &= i_a \\ i_{\beta} &= (i_a + 2i_b) / \sqrt{3} \end{aligned} \quad (2.2)$$

where  $i_a + i_b + i_c = 0$

- $\alpha$ - $\beta$  to d-q (Park transformation): The two axis orthogonal stationary reference frame quantities are then transformed to rotating reference frame quantities. The transform is expressed by the following equations [1] [21]:

$$\begin{aligned} i_d &= i_{\alpha} \cos \theta + i_{\beta} \sin \theta \\ i_q &= i_{\beta} \cos \theta - i_{\alpha} \sin \theta \end{aligned} \quad (2.3)$$

- d-q to  $\alpha$ - $\beta$  (Inverse Park transformation): the quantities in rotating reference frame are transformed to two axis orthogonal stationary reference frame using Inverse Park transformation. The Inverse Park transformation is expressed by the following equations [1], [21]:

$$V_{\alpha} = V_d \cos \theta - V_q \sin \theta$$

$$V_{\beta} = V_q \cos \theta + V_d \sin \theta \quad (2.4)$$

- $\alpha$ - $\beta$  to a, b, c (Inverse Clarke transformation): The transformation from two axis orthogonal stationary reference frame to the three phase stator stationary reference frame, is accomplished using the inverse Clarke transformation. The inverse Clarke transformation is expressed by the following equations [1] [21]:

$$\begin{aligned} V_a &= V_{\alpha} \\ V_b &= \frac{[-V_{\alpha} + \sqrt{3}V_{\beta}]}{2} \\ V_c &= \frac{[-V_{\alpha} - \sqrt{3}V_{\beta}]}{2} \end{aligned} \quad (2.5)$$

## 2.4 LITERATURE REVIEW

Speed control of PMSMs (permanent magnet synchronous motors) has been a topic of interest for the last thirty years [3], [16]. And to achieve high-performance vector control for PMSMs, different papers have been published reporting different controller design for such motors.

**In 2007 Marek Stulrajter, Val'eria Hrabovcov'a and Marek Franko** [18] presented on the control theory of PMSM. This paper presents different approaches of PMSM control strategies. Scalar control and vector control were chosen as possible methods for AC motors control. Theoretical background briefly describes the properties of these control techniques. Several advantages and disadvantages are highlighted and confirmed by simulations. In the paper there was shown that it is possible to use a scalar control which brings an advantage of sensorless control. On the other hand, the fact that the scalar control in an open loop does not provide a possibility to control the currents causes different behaviours of PMSM during various operation cycles like the start-up and loading. The vector control as the most common method of SM control was used to compare the performances of the proposed scalar control approach. The paper confirms not only better dynamics performances of vector control but also shows that AC motors with PM must be controlled by the vector control to avoid oversaturation of the magnetic circuit, increasing the magnetizing current and, hence, losses [18]. Thus this thesis is done and focused on the vector control technique since it helps to transform the time varying dynamic model to the d-q reference frame dynamic model.

**In 2011 Cunjian Xia, Xiaocui Wang, Shihua Li and Xisong Chen [13]** presented on an Improved Integral Sliding Mode Control Methods for Speed Control of PMSM System to improve the disturbance rejection property of permanent magnet synchronous motor (PMSM) speed control system. However, the simulation and implementation results show that it is difficult to balance the chattering and the anti-disturbance capacity. To this end, three kinds of improved ISMC control methods are developed. First, ISMC using linear varying gain is developed. Using this method, the switching gain of ISMC controller can be smaller while still ensuring that the speed state reaches its steady state and the steady state fluctuations can thus be reduced. Moreover, the anti-disturbance capacity of the PMSM system can also be assured. Second, an integral sliding mode control based on extended state observer (ESO) is developed. ESO can estimate both of the states and the disturbances simultaneously. By using ESO, an estimate of the lumped disturbances is obtained, which is employed for the feed forward compensation design of the composite ISMC control law. In this case, the controller may take a smaller value for the switching gain without sacrificing disturbance rejection performance, which helps to reduce large chattering caused by high control gains. Third, an adaptive composite control method combining linear varying gain and ESO is developed to take advantages of both improved methods. These improved methods show advantages in reducing the chattering while ensuring the dynamic and disturbance rejection performance. Both of simulation and experiment results are provided [13]. But chattering is reduced but not eliminated on the proposed paper. So by transferring the discrete items to the higher derivatives of output variables, higher order sliding modes remove the chattering effect.

**In 2011 Mahlet Legesse Gebresilassie [17]** presented on the Speed Control of Vector Controlled PMSM Drive using Fuzzy Logic-PI Controller. This study is focused on the speed control performances of a Permanent Magnet Synchronous Motor (PMSM). Usually, a proportional-integral (PI) controller is used as a speed controller for a Permanent Magnet Synchronous Motor (PMSM) in high performance drive system despite the existence of many modern nonlinear control techniques. However, a PI controller is sensitive to speed changes, load disturbances and parameters variation without continuous tuning of its gains. The conventional approach to these issues is to tune the proportional and integral gains manually by observing the response of the system. The tuning of the PI parameters must be made on-line and automatic in order to avoid tedious tasks in manual control. The well-known Ziegler-Nichols method to tune

the coefficients of a PI controller is very simple, but cannot guarantee to be always effective. For this reason, this thesis proposed the design of an on-line self-tuning PI controller scheme using fuzzy logic controller (FLC). The performance of the proposed controller is tested through a wide range of reference speeds as well as with load variations through simulation using MATLAB/SIMULINK and Real Time Workshop (RTW) packages. The developed controller algorithm with field oriented control (FOC) is then simulated on a Texas Instruments TMS320F2812 Digital Signal Processor (DSP) device simulator. In this thesis work a FLC comprising 49 rules is proposed to improve the dynamic performance of the drive system. The simulation results show that the speed response of the proposed drive is unaffected by load torque & parameters variation. Furthermore, the developed controller works well even when the reference speed is changed, i.e., the developed controller speed regulation is unaffected by the nature of the reference speed. These shows remarkable performance improvement compared to conventional PI speed controller [17]. In this paper, it makes the design simple but it uses the PI controller for torque and flux control. Thus the performance is affected in some extent. So to improve the performance, I used the PID controller for the torque and flux control.

**In 2011 Huangfu Yigeng<sup>1</sup>, S. Lagrouche, Liu Weiguo<sup>1</sup> and A. Miraoui [29]** presented on Robust High Order Sliding Mode Control of Permanent Magnet Synchronous Motors. This paper applies the research of nonlinear control and high order sliding mode control theory in PMSM control, and achieves robust control for a PMSM in spite of the internal parameter uncertainties and unknown external disturbance load torque. The simulation results show good performance; in addition, the estimation online of system state variables is also one of the hot issues in the control field. In this paper, a new design based on high order sliding mode with differentiator for PMSM, access to the state variable estimation; besides, unknown uncertain load impacts the performance of motor control. In order to improve system performance, this paper also uses external disturbance load estimation online. It makes sure the load can be accurately estimated. It described dSPACE physics experiment control platform the build and development process in detail. Through the dSPACE real-time control platform, the nonlinear high order sliding mode control theory research is applied to the control of permanent magnet synchronous motor. The experimental results and simulation results are consistently indicate that synchronous motor has better dynamic performance and steady accuracy, proves the feasibility of this technology in practical application systems; It is also verified by high order sliding mode control technique that

preserves the robustness of traditional sliding mode control. The high order sliding mode essentially eliminates the chattering caused by discrete control law. From another point of view, the simulation and physical experiment provide a certain reference value for the nonlinear systems high order sliding mode control further application [29]. In this paper it controls the position of the motor and shows the advantage of the HOSM controller. But the paper uses a differentiator and input/output feedback linearization. So, this makes the system more complicated. So I make the system design simple and reduced the chattering effect in this thesis.

**In 2014 Zhugang Ding, Guoliang Wei and Xueming Ding [14]** presented on Speed identification and control for permanent magnet synchronous motor via sliding mode approach. In this paper, speed identification and control problems are simultaneously considered for a permanent magnet synchronous motor (PMSM) based on the sliding mode technique. Thus, a variable-structure MRAS speed observer is designed to avoid the installation of the mechanical sensor. To reduce system chattering, a sigmoid function with an adaptive gain is employed as a continuous control in variable-structure control instead of the traditional signum function. Secondly, the conventional speed PI actuator is replaced by a sliding mode controller. It makes up the shortcomings of speed PI regulator to a certain extent. Finally, the reasonability and validity of this method have been testified by the simulation results. The simulation results show that this speed observer can track the reference speed rapidly and precisely when the reference speed and load torque change, and the three-phase current and torque of the motor also have a good dynamic response. Obviously, simulation results indicate that the proposed control scheme eliminates the speed overshooting, and holds the advantages of fast response and strong robustness [14]. Here, chattering is not reduced and removing of the mechanical sensor causes instability of the system since it is very difficult to get accurate rotor position from the observer.

**In 2010 Marwa Ezzat, Nicolas Gonzalez and Alain Glumineau [15]** presented on an Observer-Controller Scheme using High Order Sliding Mode Techniques for Sensorless Speed Control of Permanent Magnet Synchronous Motor. This paper deals with the control problem of the speed for a sensorless permanent magnet synchronous motor (PMSM) using high order sliding mode techniques. An observer is designed via a super twisting algorithm in order to estimate the speed and the position of the motor from the currents and the voltages measurements. A quasi-continuous high order sliding mode controller is designed for the speed

track a desired reference under the presence of parameter uncertainties. Simulation results are shown in the framework of an industrial benchmark to illustrate the performance of the proposed scheme. The results are obtained in spite of parameter uncertainties on stator resistance and stator inductance that are bad known. Furthermore, the whole system (observer based controller and motor) is tested by using a specific industrial benchmark. Significant robustness tests are implemented and simulation results illustrate the efficiency of the proposed sensorless observer-controller scheme [15]. Even if this work solves the chattering problem, it doesn't assure on the stability of the system due to the accuracy of the observer on the rotor position. And the other thing is, the system design is more complicated due to the use of input output feedback linearization. It is better to use a PID controller to control the currents in order to make the system simple.

**In 2010 Yi-Geng Huangfu [22]** presented on the research of nonlinear system, high order sliding mode control and its applications for PMSM. The paper more focuses on the PMSM due to its inherent advantages. For improving performance, this paper applied nonlinear high order sliding mode research achievement to MIMO permanent magnet synchronous motor. It changes the coupling nonlinear PMSM to single input single output (SISO) linear subsystem control problem instead of near equilibrium point linearization. Thereby, the problem of nonlinear and coupling for PMSM has been solved. In addition, Uncertainty nonlinear robust control system has been well-received study of attention. Because the robust control theory is essentially at the expense of certain performance. This kind of robust control strategy often limits bandwidth of closed loop, so that system tracking performance and robustness will be decreased. So, sliding mode control is an effective approach for improving system robust. This thesis first proposed a robust high order sliding mode controller for PMSM. The system has good position servo tracking precision in spite of parameters uncertainties and external torque disturbance. The results of simulation indicate observe value has high precision when sliding mode variable and its differentials are convergent into zero. The same theory is used in external unknown torque disturbance estimation online for PMSM. As if, load torque will no longer be unknown disturbance. System performance can be improved greatly. It establishes theoretical foundation for the future applications. At the end of paper, using advanced half-physical platform controller dSPACE to drive a PMSM, hardware experiment implement is structured completely. The experiment results illustrate that PMSM adopting precious feedback linearization decoupling and

high order sliding mode controller can realize system servo tracking control with good dynamic and steady character. But in comparison with the traditional sliding mode control, high order sliding mode control technique uses high order differential of sliding mode variable to transfer the chattering phenomenon generated in discontinuous discrete control into higher order sliding surface, making the low order sliding surface eliminate the chattering and improving the control precision [22]. Considered all the above together, this paper basically focused on the position tracking, nonlinear control and high order sliding mode control with their full strengths to improve the overall performances of system but when we see the overall system design, it is more complicated and is not cost effective.

**In 2014 Romain Delpoux Thierry Floquet** [30] presented on High order sliding mode control for sensorless trajectory tracking of a PMSM. The paper presents a new sensorless approach for permanent magnet synchronous motor (PMSM). Current sensors are assumed available, but position and velocity sensors are not. Based on the electrical equations, sliding mode observers are designed to estimate the back-EMF of the motor. These estimations are used to reconstruct the position and the velocity. From this estimation, a robust sliding mode control is developed which ensures the position tracking of the motor. A new reference frame is used that presents advantages similar to the standard (d-q) frame, but without the need for a position sensor. The efficiency of the algorithm is shown through experimental results. The approach is potentially applicable to other types of synchronous motors as well [30]. Making a sensorless cause instability of a system because it is very difficult to get the accurate position of the rotor.

In this thesis, after the analysis/investigation, review and comparison of several speed controls of PMSM techniques, a combination of higher order sliding mode controller and PID controller is used. Because it reduces/eliminates the chattering effect, makes the system robust under parameter variation and load torque disturbance, gives a quick dynamic response, eliminates the basic steady-state error, makes the system design simple and stable.

## CHAPTER 3

# MODELING AND DESIGNING OF A SPEED CONTROL OF PMSM

The dynamic model of a PMSM is a nonlinear time-varying system with MIMO because the dynamic model includes highly coupling product terms of angular velocity and d-q axis current. It must achieve decoupling and linearization of the system to get the accurate control performance. In the past years, many different control algorithms have been used to improve the performance of the magnet motor. For example, as the dynamic model of the machine is nonlinear, a natural approach is the exact feedback linearization control method [22], by which the original nonlinear model can be transformed into a linear model through proper coordinate transformation. However, in general, the dynamics of the synchronous motors may not be fully known, since some of parameters appearing in the equations will vary. For instance, the resistance and inductance will be changed when the temperature alters. As a consequence, nonlinearities can only be partially cancelled by the feedback linearization technique, and parameters uncertainties act on the equations of the motion. Then an important aim of the control design is to develop a robust controller which ensures good dynamic performances in spite of parameters uncertainties and perturbation [22].

This chapter deals on the detailed mathematical modeling of a permanent magnet synchronous motor (PMSM) and controller design. Field oriented control of the motor in constant torque and flux-weakening regions are discussed. The outer loop is developed using a higher order sliding mode (HOSM) controller to control the motor speed whereas the inner loop is used to control both the torque and flux produced by the motor using PID controller. Design of the speed controller and different components which are used for the control design such as SVPWM, inverter, speed/position sensors and their principle of operations are illustrated and presented briefly in the next sections.

### 3.1 MATHEMATICAL MODEL OF PMSM

The PMSMs have an electromagnetic stator and a permanent magnet rotor. The magnetic flux of the rotor is caused by permanent magnets, rather than an alternating current, which is the case in DC motors. Since current isn't applied in the rotor, there is no need for mechanical contact between stator and rotor, hence described as brushless.

There are several design methods for building the rotor in a PMSM. The most notable method in design is the number of magnets or pole pairs mounted. The number of pole pairs will affect the synchronous speed and the amount of torque the motor yields for a certain current. Another design method is either the rotor magnets are mounted on the iron core (surface mounted PMSM) or the magnets are mounted inside the iron core (internal mounted PMSM). The type of mounting will affect the inductance created by the rotor. The inductance of stator winding is generally a function of rotor position. When a stator winding is energized, applying a DC voltage for a certain time, a magnetic field with a fixed direction will be established. The current response of the winding is different due to inductance variation. But if we assume that the rotor has the surface mounted design, which is the case of most brushless motors, there is no saliency such that the stator self-inductances are independent of the rotor position. Figure 3.1 below shows a PMSM which has surface-mounted permanent magnets rotor [23].

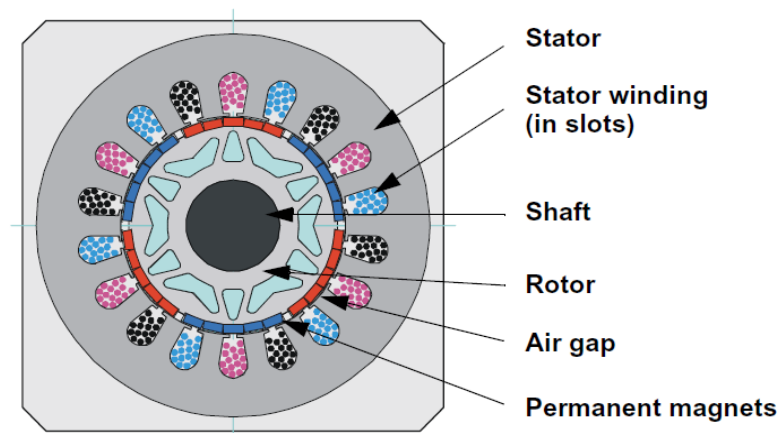


Figure 3.1: Cross Section of a PM Synchronous Motor [23]

The common analysis of permanent magnet synchronous motor is d-q axis mathematical model as discussed in chapter 2. It can be used to analyze not only the permanent magnet synchronous

motor steady state operating characteristics, but also to analyze the transient performance of the motor. The d-q model has been developed on rotor reference frame as shown in figure 3.2. At any time  $t$ , the rotating rotor d-axis makes an angle  $\theta_r$  with the fixed stator phase axis and rotating stator mmf makes an angle  $\alpha$  with the rotor d-axis. Stator mmf rotates at the same speed as that of the rotor [26].

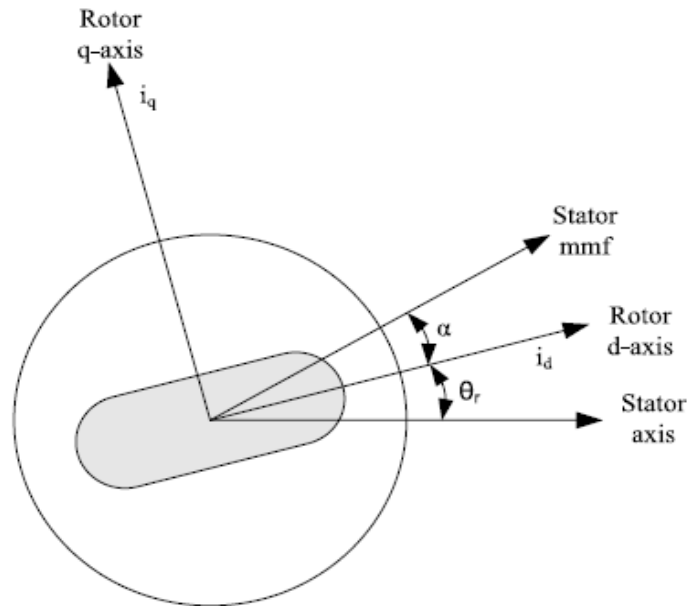


Figure 3.2: Motor Axis [26]

In order to establish sinusoidal PMSM d-q axis mathematical model and for design simplification, firstly I assume that:

- The saturation and parameter changes are neglected.
- The motor current is symmetrical three phase sine function.
- Stator resistances and inductances of all the windings are equal.
- Eddy currents and hysteresis losses are negligible.

### Model equations of the motor in stator reference frame (a, b, c) [1] [24] [25] [26]

Let,  $R$  – resistance,  $P$  – No of pole pairs;  $L$  - stator inductance;  $E_a, E_b, E_c$  - are back EMF in (a, b, c) axis respectively;  $\lambda_m$  - rotor flux linkage;  $\theta$  - the electrical angle between rotor axis and stator;  $\theta_r$  -mechanical angular position of motor;  $\omega_r$  – rotor mechanical speed;  $\omega$  -synchronous

speed;  $V_a, V_b, V_c$  &  $i_a, i_b, i_c$  are respectively stator voltages and currents in (a, b, c) axis;  $T_e$ - electromagnetic torque;  $T_m$ - electromechanical torque;  $T_L$  -load torque,  $J$ -moment of inertia,  $B$ -friction coefficient.

In (a, b, c) stationary reference frame, the voltage equation of a three phase PMSM can be described as:

$$V_{abc} = R' i_{abc} + \frac{\partial \lambda_{abc}}{\partial t} \quad (3.1)$$

Where,  $\lambda_{abc}$  is stator flux linkage,

$$(f_{abc})^T = [f_a \quad f_b \quad f_c], \text{ f can represent V, i or } \lambda \quad (3.2)$$

$$R' = \begin{bmatrix} R & 0 & 0 \\ 0 & R & 0 \\ 0 & 0 & R \end{bmatrix} \quad (3.3)$$

The flux linkage equation may be written as:

$$\lambda_{abc} = L \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \lambda_m \begin{bmatrix} \sin(\theta) \\ \sin\left(\theta - \frac{2\pi}{3}\right) \\ \sin\left(\theta + \frac{2\pi}{3}\right) \end{bmatrix} \quad (3.4)$$

Equation 3.1 can be also written as:

$$\begin{aligned} V_a &= R i_a + L \frac{di_a}{dt} + E_a; E_a = \omega \lambda_m \sin(\theta) \\ V_b &= R i_b + L \frac{di_b}{dt} + E_b; E_b = \omega \lambda_m \sin\left(\theta - \frac{2\pi}{3}\right) \\ V_c &= R i_c + L \frac{di_c}{dt} + E_c; E_c = \omega \lambda_m \sin\left(\theta + \frac{2\pi}{3}\right) \end{aligned} \quad (3.5)$$

The Electromagnetic torque:

$$T_e = (E_a i_a + E_b i_b + E_c i_c) / \omega \quad (3.6)$$

It can be written as:

$$T_e = \frac{p}{2} \lambda_m \left[ \left( i_a - \frac{1}{2} i_b - \frac{1}{2} i_c \right) \sin(\theta) - \frac{\sqrt{3}}{2} (i_b - i_c) \cos(\theta) \right] \quad (3.7)$$

**Model equations of the motor in the stationary orthogonal ( $\alpha, \beta$ ) reference frame [15] [24] [25] [26]**

$$\begin{aligned} V_{\alpha} &= Ri_{\alpha} + L \frac{di_{\alpha}}{dt} - \lambda_m \omega \sin(\theta) \\ V_{\beta} &= Ri_{\beta} + L \frac{di_{\beta}}{dt} + \lambda_m \omega \cos(\theta) \end{aligned} \quad (3.8)$$

Electromagnetic torque

$$T_e = \lambda_m \frac{p}{2} [-i_{\alpha} \sin(\theta) + i_{\beta} \cos(\theta)] \quad (3.9)$$

As we can see, the above equations are non-linear and strongly coupled with each other. These factors make it very difficult to perform direct control of flux and torque by varying the currents in the stator coils. Therefore; transforming to another coordinate system is needed.

**Model equations of the motor in the Rotary Orthogonal reference ( $d, q$ ) frame [15] [22] [25] [26]**

The common analysis of permanent magnet synchronous motor is  $d$ - $q$  axis mathematical model. It can be used to analyze not only the permanent magnet synchronous motor steady state operating characteristics, but also the transient performance motor.

Finally, the model of the motor in rotary time invariant reference frame will be:

Voltage equations are given by:

$$\begin{aligned} V_d &= Ri_d + \frac{\partial \lambda_d}{\partial t} - \omega \lambda_q \\ V_q &= Ri_q + \frac{\partial \lambda_q}{\partial t} + \omega \lambda_d \end{aligned} \quad (3.10)$$

Where,  $\lambda_d = Li_d + \lambda_m$ ,  $\lambda_q = Li_q$

When the value of  $\lambda_d$  and  $\lambda_q$  are substituted in equation (3.10) and rearranged, the following differential equations is obtained.

$$\frac{di_d}{dt} = -\frac{R}{L}i_d + \omega i_q + \frac{V_d}{L}$$

$$\frac{di_q}{dt} = -\frac{R}{L}i_q - \omega i_d - \frac{\lambda_m}{L}\omega + \frac{V_q}{L} \quad (3.11)$$

Electromagnetic torque is given by

$$T_e = \frac{3P}{2}(\lambda_d i_q - \lambda_q i_d) \quad (3.12)$$

In order to produce maximum torque, to avoid reluctance effects and torque ripple, optimal operation is achieved by vector control, which ensures that the stator- current space vector contains only a quadrature component by pushing d-axis current towards zero [15] [20] [23] [25]. As a result the torque equation becomes:

$$T_e = \frac{3P}{2}\lambda_m i_q \quad (3.13)$$

Thus the torque can be controlled directly by the current  $i_q$  only.

And for all reference frame the electromechanical torque is

$$J \frac{d\omega_r}{dt} = T_e - B\omega - T_m; \quad \text{Where } \omega_r = \frac{d\theta_r}{dt} \quad (3.14)$$

Thus, from equation (3.13) and (3.14)

$$\frac{d\omega_r}{dt} = \frac{1}{J} \left( \frac{3P}{2}\lambda_m i_q - B\omega_r - T_m \right) \quad (3.15)$$

Set of equations (3.11) and  $\frac{d\theta}{dt} = \omega$  we can get the state equation expression of PMSM as following.

$$\begin{bmatrix} \frac{di_d}{dt} \\ \frac{di_q}{dt} \\ \frac{d\theta}{dt} \end{bmatrix} = \begin{bmatrix} -\frac{R}{L}i_d + \omega i_q + \frac{V_d}{L} \\ -\frac{R}{L}i_q - \omega i_d - \frac{\lambda_m}{L}\omega + \frac{V_q}{L} \\ \omega \end{bmatrix} \quad (3.16)$$

Using equation (3.14), (3.15),  $\omega = P\omega_r$  and  $\theta = P\theta_r$ , the state equation expression of the mechanical motion of PMSM is

$$\begin{bmatrix} \frac{d\theta_r}{dt} \\ \frac{d\omega_r}{dt} \\ \frac{di_d}{dt} \\ \frac{di_q}{dt} \end{bmatrix} = \begin{bmatrix} \omega \\ \frac{3P}{2J}\lambda_m i_q - \frac{B}{J}\omega_r - \frac{T_l}{J} \\ -\frac{R}{L}i_d + P\omega_r i_q + \frac{V_d}{L} \\ -\frac{R}{L}i_q - P\omega_r i_d - P\frac{\lambda_m}{L}\omega_r + \frac{V_q}{L} \end{bmatrix} \quad (3.17)$$

$V_d, V_q, i_d$  &  $i_q$  are d-axis and q-axis voltage and current component.

From the equation (3.17) we can see that PMSM is a multi-variable, coupling, nonlinear time varying systems.

Taking  $i_d, i_q, \omega_r, \theta_r$  as state vector  $V_d, V_q$  as input vector and  $i_d, i_q$  as output vector, The input output state space representation of the system in load condition becomes:

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ \omega_r \\ \theta_r \end{bmatrix} = \begin{bmatrix} -\frac{R}{L} & P\omega_r & 0 & 0 \\ -P\omega_r & -\frac{R}{L} & -P\frac{\lambda_m}{L} & 0 \\ -P\omega_r & \frac{3P}{2J}\lambda_m & -\frac{B}{J} & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ \omega_r \\ \theta_r \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{L} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} V_d \\ V_q \end{bmatrix} \quad (3.18)$$

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ \omega_r \\ \theta_r \end{bmatrix} \quad (3.19)$$

### 3.2 CONTROLLER DESIGN OF THE SYSTEM

The general system block diagram in vector control method is as shown in figure 3.3. The system basically includes speed sensor and Controllers (higher order sliding mode controller i.e a quasi-continuous sliding mode controller, generation of reference currents and PID controller). The system also includes different co-ordinate transformation, such as Clarke, Park and IPark for the purpose of vector control. Space vector pulse width modulation (SVPWM) is also part of the system which calculates the operation time of the inverter gates from the controlled d-q space vector voltage value for proper application voltages on the stator windings. The controller attains the required value of set point by using measured stator currents and speed as feedback value.

## Speed Control of PMSM using HOSM Controller

The controller is Quasi-continuous high order sliding mode controller which is recently used in industrial application to reduce chattering effect.

A properly designed feedback controller makes the system insensible to disturbance and changes of the parameters. The purpose of a motor speed controller is to take a signal representing the demanded speed, and to drive a motor at that speed. Closed Loop speed control systems have fast response, but become expensive due to the need of feedback components such as speed sensors.

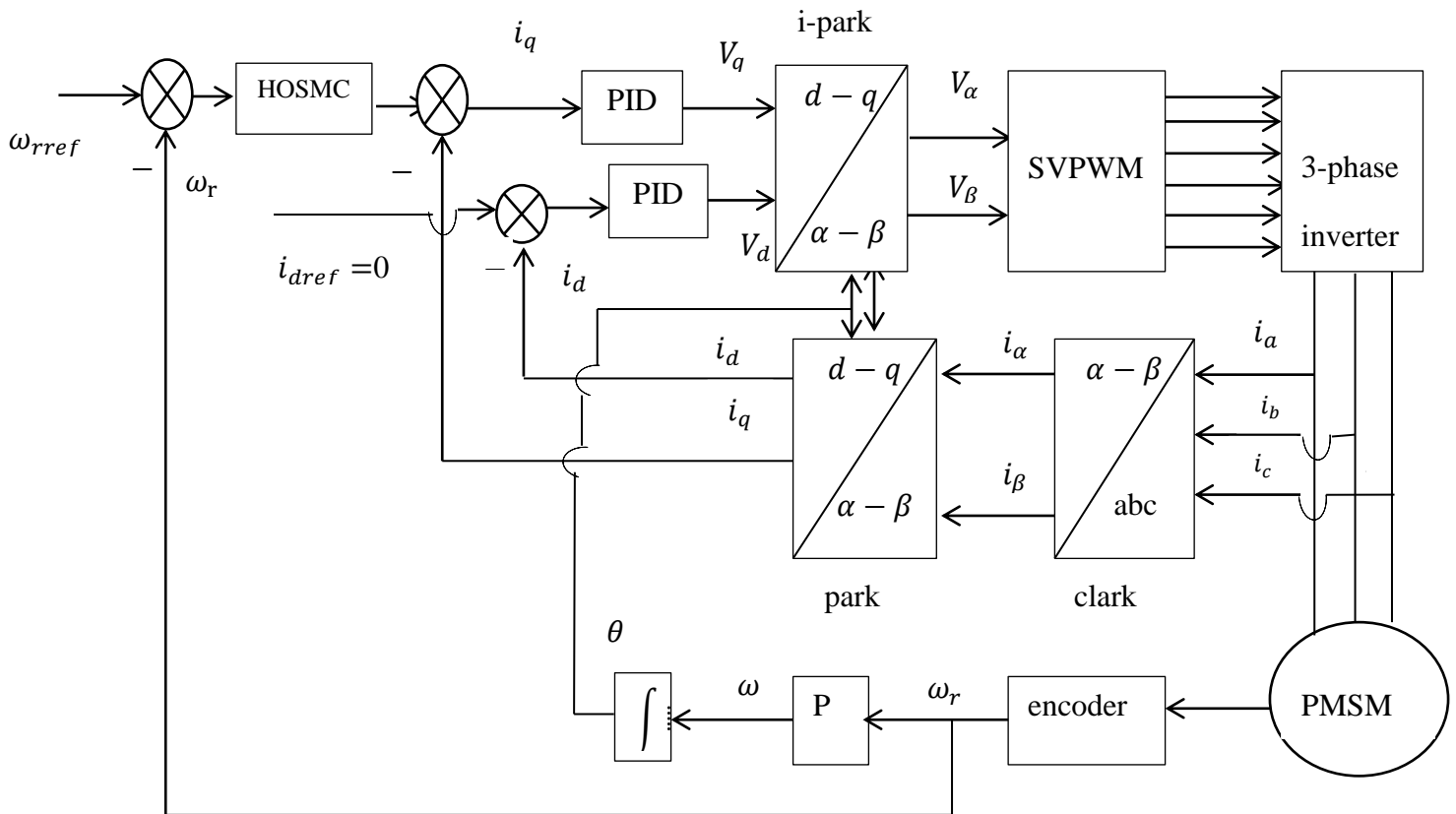


Figure 3.3: system model for speed control of PMSM using HOSM controller

To control the system speed and torque, higher order sliding mode controller especially quasi-continuous sliding mode controller and Proportional, Integral and Derivative (PID) controller is used respectively.

### 3.2.1 High order sliding mode control for PMSM

Because of their own advantages of permanent magnet synchronous motor, it is very suitable for modern industrial control and applications. But the synchronous motor is a typical nonlinear, multivariable system with coupling. So, in recent years, there are a lot of algorithms to improve the performance of motor control, For instance, exact feedback linearization control. But in a dynamic system, it is impossible to accurately inform of all the parameters of the synchronous motor. The system parameters in the equation will change due to the environment, such as resistors and inductors will be changed as different external temperature. Therefore, the design of robust controller is very important to ensure good dynamic performance with the parameter uncertainties and external disturbances. It is based on this consideration the sliding mode control proposed and applied.

The sliding mode control is the essence of high-frequency switching feedback control. The switching value of its control law switches over according to the system state. The design the sliding mode controller is composed mainly by two parts, the sliding surface and control law. So that the dynamic system have been bound at the sliding mode surface. This kind of state also is known as sliding mode state, which is only relative with the choice of sliding surface. The dynamic system of the sliding state is without outside influence. The main idea of sliding mode control is to keep a properly chosen constraint at zero in sliding mode. It is natural to classify sliding modes by the smoothness degree of the constraint function calculated along the system trajectories [15] [27].

In recent years, a new type of high order sliding mode control is proposed, and has been widely used. The characteristic of the control is high order derivative of sliding mode variables in place of the original discrete control, so that the chattering disappears in the high order differentiation. The switch function in traditional sliding mode control is generally dependent on the system state, has nothing to do with the control input. Discontinuous items will be directly transferred to the controller. But, the high order sliding mode control constitutes a new sliding mode surface using the sliding mode control variable and its differentiations, transfers the discrete items to the first order or higher order derivative of sliding mode variable.

Using high order sliding mode control, it not only eliminates the chattering problems that exist in the traditional sliding mode, but also avoids the requirements of relative degree that must be equal to 1 in the control law design. The technology not only retains advantage of strong robustness in the traditional sliding mode control, but also enables discontinuous item variables transmit into the first order or higher order sliding mode derivative to eliminate the chattering. Besides, the design of the controller no longer must require relative degree to be 1. Therefore, it is greatly simplified to design parameters of sliding mode surface.

The two most well-known r-sliding mode controller families are Nested Sliding mode Controllers and Quasi-continuous Sliding mode Controllers. The controllers of the form

$$u = -\alpha\psi_{r-1,r}(\sigma, \dot{\sigma}, \dots, \sigma^{(r-1)}) \quad (3.20)$$

are defined by recursive procedures, have magnitude  $\alpha > 0$ , and solve the general output regulation problem for system  $(\sigma^{(r)} \in [-C, C] + [K_m, K_M]u)$ . The parameters of the controllers can be chosen in advance for each relative degree r. Only the magnitude  $\alpha$  is to be adjusted for any fixed  $C, K_m, K_M$ , most conveniently by computer simulation, thus avoiding complicated and redundantly large estimations [15] [22] [27].

### **Quasi-continuous Sliding Mode Controllers**

The obvious disadvantage of the nested sliding mode controllers is the high frequency leaping of transient trajectories when they approach the HOSM. Although the leaping magnitude tends to zero during the transient, it may cause problems in practice [27].

Consider the following nonlinear system with a smooth output function  $\sigma$  [22] [27]

$$\begin{aligned} \dot{x} &= f(t, x) + g(t, x)u \\ y &= \sigma(t, x) \end{aligned} \quad (3.21)$$

Where  $x \in \mathbb{R}^n$ ,  $f$  and  $g$  and  $\sigma: \mathbb{R}^{n+1} \rightarrow \mathbb{R}$  are unknown smooth functions,  $u \in \mathfrak{R}$ ,  $n$  is also uncertain. Introducing a new change of coordinates  $\xi_1 = \sigma, \xi_2 = \dot{\sigma}, \dots, \xi_r = \sigma^{(r-1)}$  where  $r$  is the relative degree of the system which is assumed to be constant and known, i.e.

$$\sigma^{(r)} = F(t, x) + G(t, x)u \quad (3.22)$$

Where  $F(t, x) = \sigma^{(r)}|_{u=0}$ ,  $G(t, x) = \frac{\partial \sigma^r}{\partial u}|_{u \neq 0}$ . It is supposed that for some  $K_m, K_M, C > 0$ , the following inequalities hold at least locally  $0 < K_m \leq \frac{\partial \sigma^r}{\partial u} \leq K_M$ ,  $\sigma^r|_{u=0} \leq C$ .

The control law in Eq. (3.20) should be discontinuous at least at the HOSM:

$$\sigma = \dot{\sigma} = \dots = \sigma^{(r-1)} = 0 \quad (3.23)$$

An r-sliding controller Eq. (3.20) is called quasi-continuous, if it is continuous everywhere except at Eq. (3.23).

The finite stabilization of system at the origin has been proved in the following.

**Theorem:** Assuming that the constants  $\beta_1, \beta_2, \dots, \beta_{r-1}, \alpha > 0$  are chosen sufficiently large in the lexicographic order, the controller Eq. (3.20) is a r-sliding homogeneous controller and provides for the finite-time stability of the Eq. (3.20) and (3.21) where [15]

$$\begin{aligned} \varphi_{0,r} &= \sigma, N_{0,r} = |\sigma|, \psi_{0,r} = \varphi_{0,r}/N_{0,r} = \text{sign}(\sigma). \\ \varphi_{i,r} &= \sigma^{(i)} + \beta_i N_{(i-1,r)}^{(r-i)/(r-i+1)} \psi_{i-1,r} \\ N_{i,r} &= |\sigma^{(i)}| + \beta_i N_{(i-1,r)}^{(r-i)/(r-i+1)} \psi_{i-1,r} \\ \psi_{i,r} &= \varphi_{i,r}/N_{i,r} \end{aligned} \quad (3.24)$$

Furthermore,  $N_{i,r}$  is positive definite, ( $N_{i,r} = 0$  iff  $\sigma = \dot{\sigma} = \dots = \sigma^{(r-1)} = 0$ ). The inequality  $|\psi_{r-1,r}(\sigma, \dot{\sigma}, \dots, \sigma^{(r-1)})| \leq 1$  holds whenever  $N_{i,r} > 0$  for  $i = 1, \dots, r-1$ . The function  $\psi_{r-1,r}(\sigma, \dot{\sigma}, \dots, \sigma^{(r-1)})$  is continuous everywhere except the point  $\sigma = \dot{\sigma} = \dots = \sigma^{(r-1)} = 0$ .

The choice of parameters  $\beta_1, \beta_2, \dots, \beta_{r-1}, \alpha > 0$  determines a controller family applicable to all systems (3.21) of relative degree r. The parameter  $\alpha$  is chosen specifically for any fixed  $C, K_m, K_M$ , most conveniently by computer simulation, avoiding redundantly large estimations of  $C, K_m, K_M$ .

The following are quasi-continuous controllers of eq.(3.20) with  $r \leq 3$ , which have simulation tested  $\beta_i$  [15] [27]:

$$r = 1: u = -\alpha \text{sign}(\sigma)$$

$$r = 2: u = -\alpha \frac{\dot{\sigma} + |\sigma|^{1/2} \text{sign}(\sigma)}{|\dot{\sigma}| + |\sigma|^{1/2}} \quad (3.25)$$

$$r = 3: u = -\alpha \frac{\ddot{\sigma} + 2(|\dot{\sigma}| + |\sigma|^{2/3})^{-1/2} (\dot{\sigma} + |\sigma|^{2/3} \text{sign}(\sigma))}{|\ddot{\sigma}| + 2(|\dot{\sigma}| + |\sigma|^{2/3})^{1/2}}$$

### 3.2.2 PID controller

Speed control of motors mainly consist of two loops the inner loop for current and the outer loop for speed. The order of the loops is due to their response, how fast they can be changed. This requires a current loop at least 10 times faster than the speed loop.

Since the PMSM is operated using field oriented control, it can be modeled like a dc motor. The design begins with the innermost current loop by drawing the block diagram. But in PMSM system the motor has current controllers which make the current loop. The current control is performed by the comparison of the reference currents with the actual motor currents is done by PID controller.

PID controller has the optimum control dynamics including zero steady state error, fast response (short rise time), no oscillations and higher stability. The necessity of using a derivative gain component in addition to the PI controller is to eliminate the overshoot and the oscillations occurring in the output response of the system. One of the main advantages of the PID controller is that it can be used with higher order processes including more than single energy storage.

**Proportional Term:** The main usage of the P controller is to decrease the steady state error of the system. As the proportional gain factor K increases, the steady state error of the system decreases.

**Integral Term:** it is mainly used to eliminate the steady state error.

**Derivative Term:** D mode is designed to be proportional to the change of the output variable to prevent the sudden changes occurring in the control output resulting from sudden changes in the error signal.

**Controllers loop design and calculations of gains**

The rotor speed is controlled by quasi-continuous sliding mode controller and, both rotor flux and rotor torque are each controlled by a separate PID module.  $K_p + \frac{K_I}{S} + K_d s$  is the PID controller; where  $K_p$ -proportional gain,  $K_I$ - is integral gain  $K_d$ - derivative gain.

**Current controller**

For the design of PID-controllers it is necessary to know the closed loop transfer function. The d-axis current closed loop transfer function from equation (3.17) will be:

$$\frac{i_d}{V_d + PL\omega_r i_q} = \frac{1}{LS + R} \tag{3.26}$$

The Block diagram of the closed loop transfer function with the PID controller in d-axis is shown in Figure 3.4.

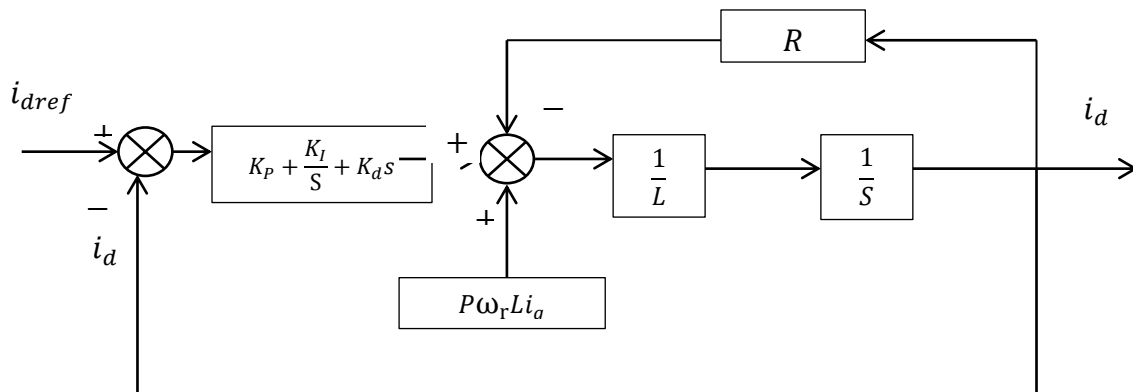


Figure 3.4: d –axis current closed loop transfer function with PID controller block diagram

Thus, the closed loop current control equation is.

$$\frac{i_d}{i_{dref}} = \frac{K_d s^2 + K_p s + K_I}{(K_d + L)s^2 + T_i (K_p + R)s + K_I} \tag{3.27}$$

Similarly, the q-axis current closed loop transfer function from equation (3.16) will be:

$$\frac{i_q}{V_q - P\omega_r(Li_d + \lambda_m)} = \frac{1}{LS + R} \quad (3.28)$$

The Block diagram of the closed loop transfer function with the PID controller in q-axis is shown in Figure 3.5.

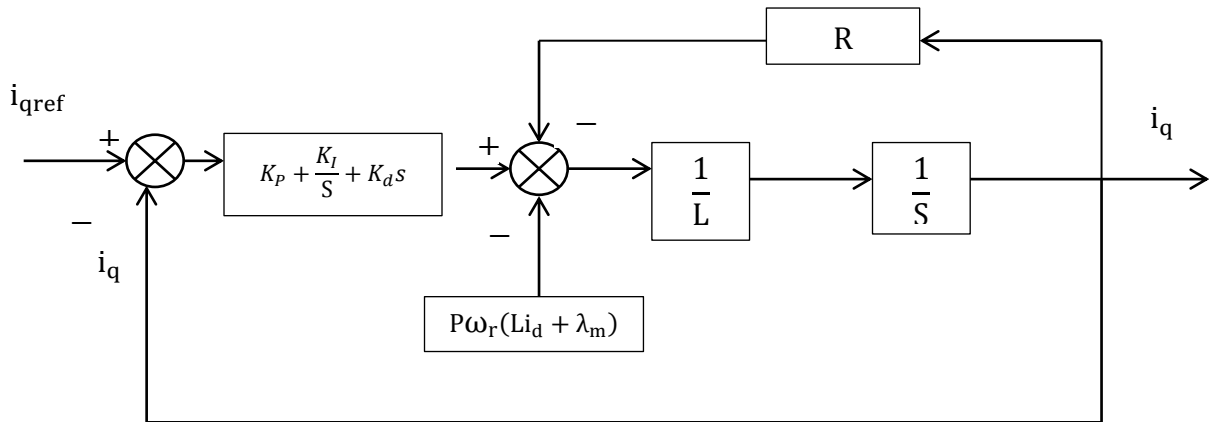


Figure 3.5: q-axis current closed loop transfer function with PID controller block diagram

As we can see from the above two block diagrams,  $P\omega_r Li_q$  and  $P\omega_r(Li_d + \lambda_m)$  block makes the closed loop transfer function complicated. For simplicity of PID controller parameter calculation, the  $P\omega_r Li_q$  and  $P\omega_r(Li_d + \lambda_m)$  part of the equation is ignored. Its effect can be compensated by tuning the parameters after calculation.

From Figure 3.5 the overall transfer function between the torque component  $i_q$  and its reference  $i_{qref}$  is as:

$$\frac{i_q}{i_{qref}} = \frac{K_d s^2 + K_p s + K_I}{(K_d + L)s^2 + T_i(K_p + R)s + K_I} \quad (3.29)$$

For proper simulation of the system, both the Proportional, Integral and Derivative gain values are determined by using Ziegler-Nichols method (see appendix A) and tuning method in the simulation to get the appropriate value.

### Speed controller

The aim is to design an appropriate control which guarantees robust performance in presence of parameters and load variations. The rotor mechanical angular velocity  $\omega_m$  must track a reference mechanical angular velocity  $\omega_{rref}$ . To track the mechanical angular velocity  $\omega_m$ , another sliding manifold is proposed so that the error dynamics follows a desired second order dynamic. Denoting  $\omega_{mref}$  the desired velocity, the following form can be obtained.

$$\sigma = \omega_{rref} - \omega_r \quad (3.30)$$

Considering load torque as external disturbance, derivative of  $\sigma$  continuously until control input appears.

$$\begin{aligned} \dot{\sigma} &= \dot{\omega}_{rref} - \dot{\omega}_r \\ \dot{\sigma} &= \dot{\omega}_{rref} - \left( \frac{3P}{2J} \lambda_m i_q - \frac{B}{J} \omega_r - \frac{T_l}{J} \right) \end{aligned} \quad (3.31)$$

The control input  $i_q$  appears in the first order derivative of  $\sigma$ , so the relative degree of  $\sigma$  equals 1. But in order to eliminate the chattering effect from the system, this thesis uses a second order derivative of  $\sigma$ .

$$\ddot{\sigma} = \ddot{\omega}_{rref} - \frac{3P}{2J} \lambda_m \dot{i}_q + \frac{B}{J} \dot{\omega}_r \quad (3.32)$$

Thus by using eq. (3.25) and (3.32), the control input  $i_q$  can be expressed as follows:

$$\begin{aligned} -\alpha \frac{\dot{\sigma} + |\sigma|^{\frac{1}{2}} \text{sign}(\sigma)}{|\dot{\sigma}| + |\sigma|^{\frac{1}{2}}} &= \ddot{\omega}_{rref} - \frac{3P}{2J} \lambda_m \dot{i}_q + \frac{B}{J} \dot{\omega}_r \\ \dot{i}_q &= \frac{2J}{3P\lambda_m} \left( \alpha \frac{\dot{\sigma} + |\sigma|^{\frac{1}{2}} \text{sign}(\sigma)}{|\dot{\sigma}| + |\sigma|^{\frac{1}{2}}} + \ddot{\omega}_{rref} + \frac{B}{J} \dot{\omega}_r \right) \\ i_q &= \int \frac{2J}{3P\lambda_m} \left( \alpha \frac{\dot{\sigma} + |\sigma|^{\frac{1}{2}} \text{sign}(\sigma)}{|\dot{\sigma}| + |\sigma|^{\frac{1}{2}}} + \ddot{\omega}_{rref} + \frac{B}{J} \dot{\omega}_r \right) dt \end{aligned} \quad (3.33)$$

The value of  $\alpha$  will determine through simulation.

### 3.3 SPACE VECTOR PULSE WIDTH MODULATION(SVPWM)

Space vector modulation is a PWM method that has more advantages when compared to classical sinusoidal weighted modulation PWM. The main advantage is by using SVPWM we can have higher bus voltage utilization (86%) and lower total harmonic Distortion (THD) [28]. One common way to represent the phase voltages A, B, C is the space vector model. The three legs of the three phase inverter can connect the phases of the motor to positive or negative terminal of DC bus voltage as shown in Figure 3.6.

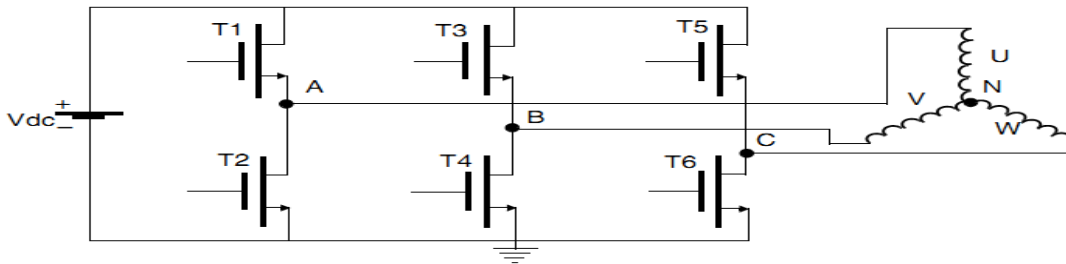


Figure 3.6: Three phase inverter

Table 3.1: Switching patterns and resulting line-to-neutral voltages

State	ON IGBT	$V_{an}$	$V_{bn}$	$V_{cn}$	Space Vector Voltage
0	T2, T4, T6	0	0	0	$U_{000}(-, -, -)$
1	T1, T4, T6	$2V_{dc}/3$	$-V_{dc}/3$	$-V_{dc}/3$	$U_0(+, -, -)$
2	T1, T3, T6	$V_{dc}/3$	$V_{dc}/3$	$-2V_{dc}/3$	$U_{60}(+, +, -)$
3	T2, T3, T6	$-V_{dc}/3$	$2V_{dc}/3$	$-V_{dc}/3$	$U_{120}(-, +, -)$
4	T2, T3, T5	$-2V_{dc}/3$	$V_{dc}/3$	$V_{dc}/3$	$U_{180}(-, +, +)$
5	T2, T4, T5	$-V_{dc}/3$	$-V_{dc}/3$	$2V_{dc}/3$	$U_{240}(-, -, +)$
6	T1, T4, T5	$V_{dc}/3$	$2V_{dc}/3$	$V_{dc}/3$	$U_{300}(+, -, +)$
7	T1, T3, T5	0	0	0	$U_{111}(+, +, +)$

Considering that one and only one switch per leg must be closed, 8 different states are possible. It is possible to associate a reference vector to each of the 8 states. In order to generate a rotating

field, the inverter has to be switched in six of the eight states. This mode of operation is called six-step mode. The remaining two states are called zero vectors. The angle between any two vectors is  $60^\circ$ . For example, the state when IGBTs T1, T4 and T6 are “on” is represented with the notation (+,-,-). The detail of this representation the resulting line-to-neutral voltages are shown in Table 3.1

Running the inverter through the sequence of (+, -, -), (+, +, -), (-, +, -), (-, +, +), (-, -, +), (+, -, +) will produce the line-to-neutral voltages shown in Figure 2.5. This strategy of operation is called “six-step mode”. Operating in this mode allows you to use the full capabilities of the inverter.

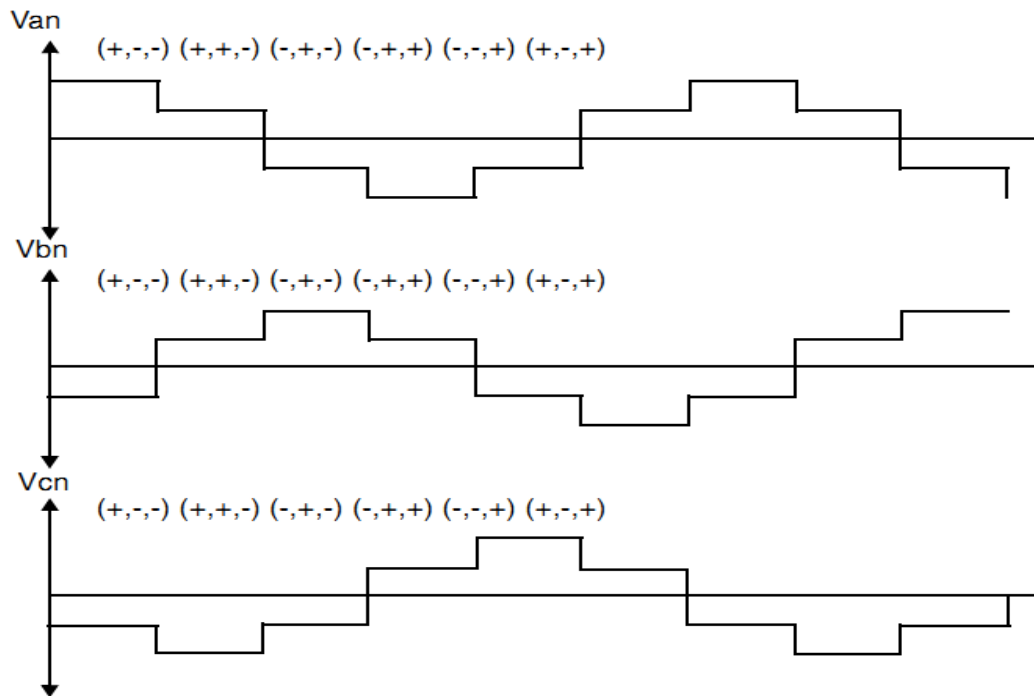


Figure 3.7: Line to neutral voltage in six-step mode

Space vector modulation uses six-step mode, but smooth out the steps through some sophisticated averaging techniques. Space vectors are similar to phasors and they are denoted by a magnitude and an angle. Space vectors are used to represent three spatially separated time variant quantities. Since the three line-to-neutral voltages sum to zero, they can easily be converted into a space vector using the following transformation:

$$U = V_{an}(t)e^{j0} + V_{bn}(t)e^{j\frac{2}{3}} + V_{cn}(t)e^{-j\frac{2}{3}} \quad (3.34)$$

Usually, when creating space vectors, the three time-varying quantities are sinusoids of the same amplitude and frequency that have  $120^\circ$  phase shifts. When this is the case, the space vector at any given time maintains its magnitude. As time increases, the angle of the space vector increases, causing the vector to rotate with a frequency equal to the frequency of the sinusoid. The purpose of space vector modulation is to generate the appropriate PWM signals so that any vector ( $U_s$ ) can be produced. Consider a space vector voltage  $U_{out}$  located in the sector defined by  $U_1$  and  $U_2$ . We can approximate  $U_s$  by applying  $U_1$  for a percentage of time ( $t_1$ ) and  $U_2$  for a percentage of time ( $t_2$ ) such that:

$$U_s = t_1 * U_0 + t_2 * U_{60} \quad (3.35)$$

$t_1$  and  $t_2$  are calculated using equation 3.36 and 3.37

$$t_1 = m \left[ \cos(\alpha) - \left( \frac{1}{\sqrt{3}} \right) \sin(\alpha) \right] \quad (3.36)$$

$$t_2 = 2m \left( \frac{1}{\sqrt{3}} \right) \sin(\alpha) \quad (3.37)$$

So, given a space vector of angle  $\alpha$ , bounded by  $U_0$ ,  $U_{60}$  and modulation index  $m$ , the approximation can be constructed by applying vectors  $U_0$  and  $U_{60}$  for percentage of times  $t_1$  and  $t_2$  respectively. If the vector is in another sector it can be rotated by a multiple of  $\pi/3$  radians until it is inside the first sector.

To approximate  $U_s$  the inverter state that corresponds to  $U_0$  should be active for  $t_1 * T_0$  seconds, and the inverter state that corresponds to  $U_{60}$  should be active for  $t_2 * T_0$  seconds. When the modulation index is sufficiently small (less than 0.866), the sum of  $t_1$  and  $t_2$  will be less than one. This means that  $t_1 * T_0 + t_2 * T_0$  is less than  $T_0$ .  $T_0$  is the sampling time, typically the PWM frequency.

Graphically, this means that for space vector modulation to work properly, the magnitude of the reference space vector,  $U_s$ , must be totally contained inside the hexagon shown in Figure 3.8. Each of the vectors  $U_0$ ,  $U_{60}$  etc., in Figure 3.8 represent the six voltage steps developed by the inverter where the zero voltages  $U_{000}$  and  $U_{111}$  are located at the origin. At each of these states the IGBT are in steady state. In order to develop a sine wave at the motor then we must devise a

switching pattern that produces a voltage at not only the six vectors states but also one which transitions in between these states. This effectively means producing a continuously rotating vector  $U_{out}$  that transition smoothly from state to state. SVPWM seeks to average out the adjacent vectors for each sector.

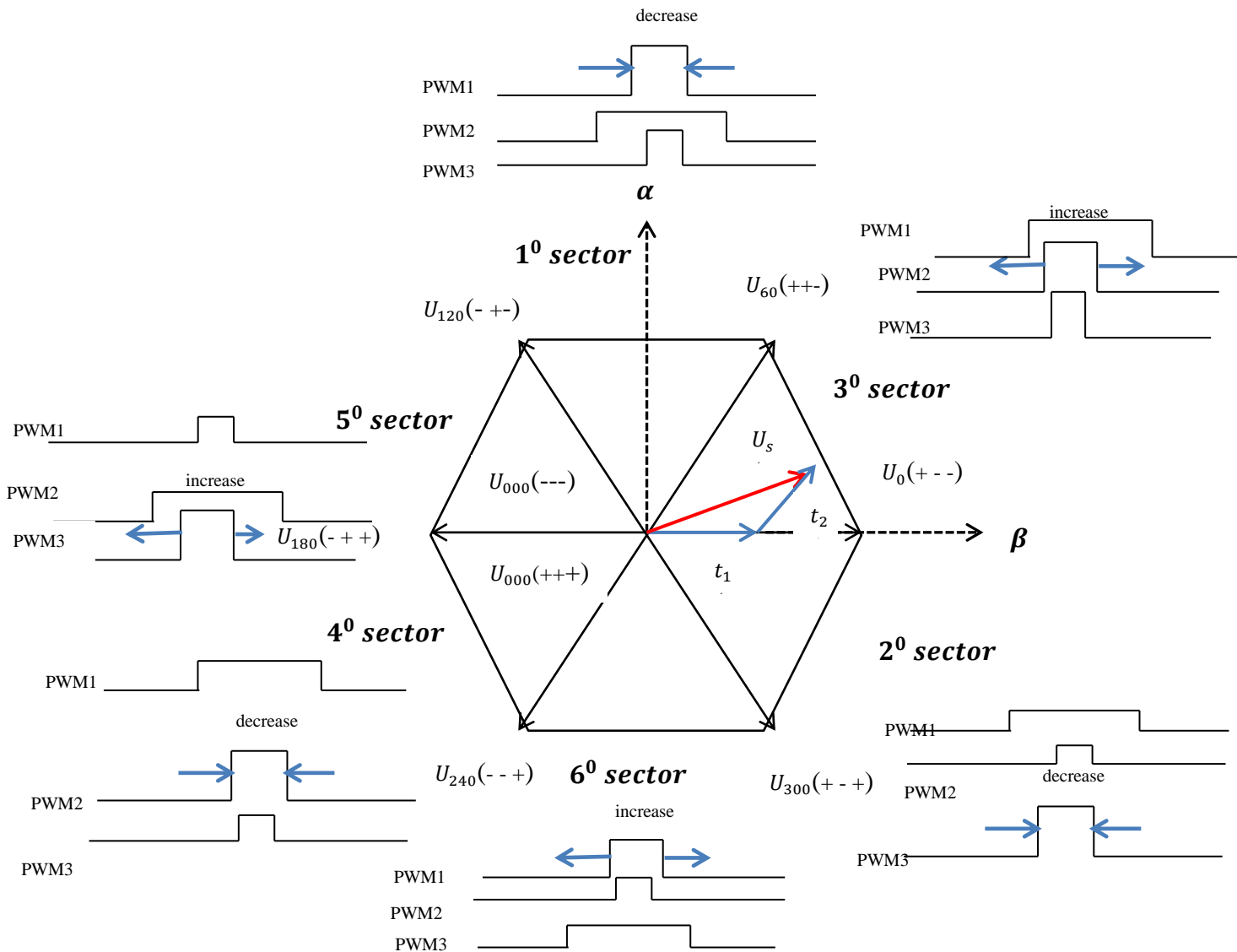


Figure 3.8: Voltage space vectors representing eight possible switching states

Using the appropriate PWM signals a vector is produced that transitions smoothly between sectors and thus provide sinusoidal line to line voltages to the motor.

### 3.4 POSITION/SPEED SENSOR

Accurate measurement of rotor position/speed is needed for applying a vector control to AC motors. The need of knowing the rotor position/speed requires the development of devices for position/speed measurement. There are four main devices for the measurement of position, the potentiometer, linear variable differential transformer, optical encoder and revolvers. The ones most commonly used for motors are encoders and revolvers. Depending on the application and performance desired by the motor a position/speed sensor with the required accuracy can be selected.

#### 3.4.1 Optical Encoder

The most popular type of encoder is the optical encoder as shown in figure 3.9, which consists of a rotating disk, a light source, and a photo detector (light sensor). The disk, is mounted on the rotating shaft, has coded patterns of opaque and transparent sectors. As the disk rotates, these patterns interrupt the light emitted onto the photo detector, generating a digital pulse or output signal. Encoders transform rotary movement into a sequence of electrical pulses. Optical encoders offer the advantages of digital interface [17]. There are two types of optical encoders, incremental encoder and absolute encoder. In this thesis a simulated incremental encoder is used.

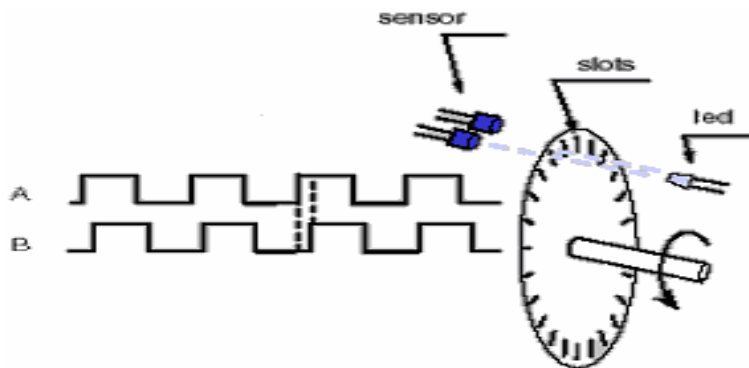


Figure 3.9: Optical encoder [17].

### Incremental encoders

Incremental encoders have good precision and are simple to implement. The disk of an incremental encoder is patterned with a single track of lines around its periphery. The disk count is defined as the number of dark/light line pairs that occur per revolution ("cycles / revolution"). The number of square wave cycles produced per one turn of the shaft is called the encoder resolution.

The most common type of incremental encoder uses two output channels (A and B) to sense position. Using two code tracks with sectors positioned  $90^\circ$  degrees out of phase, the two output channels of the quadrature encoder indicate both position and direction of rotation as shown in figure 3.10. If A leads B, for example, the disk is rotating in a clockwise direction. If B leads A, then the disk is rotating in a counter-clockwise direction. By monitoring both, the number of pulses and the relative phase of signals A and B, it's possible to track position and direction of rotation. Some quadrature encoders also include a third output channel, called a zero or index or reference signal, which supplies a single pulse per revolution. This single pulse is used for precise determination of a reference position. The precision of the encoder is fixed by its code disk but it can be increased by detecting the Up and Down transitions on both the A and B channels.

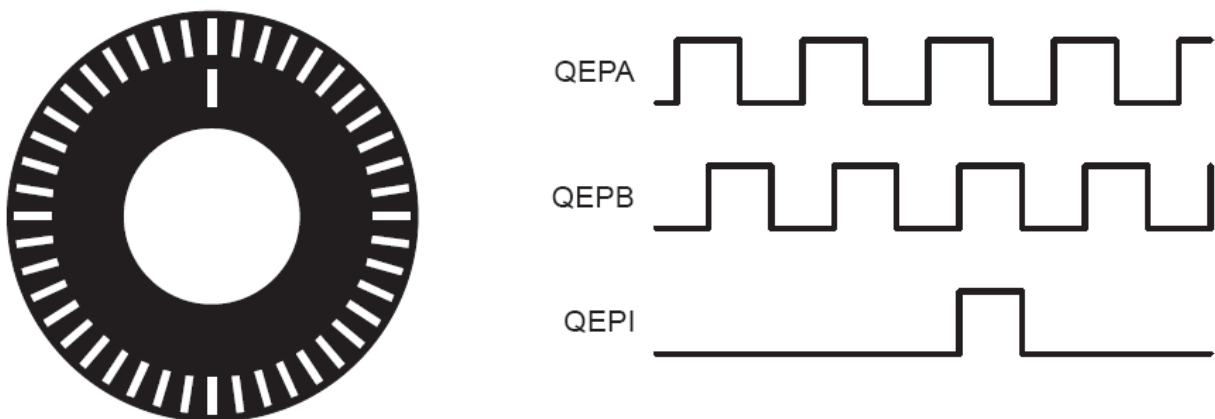


Figure 3.10: Output pulses of incremental encoder [17]

As a rule, a second track is added to generate a signal that occurs once per revolution (QEPI), which can be used to indicate an absolute position. The precision of the encoder is fixed by its code disk but it can be increased by detecting the up and down transitions on both the A and B channels. To derive direction information, the lines on the disk are read out by two different photo-elements that "look" at the disk pattern with a mechanical shift of  $\frac{1}{4}$  of a cycle (90 degrees) between them. As the disk rotates, the two photo-elements generate signals that are shifted  $90^\circ$  out of phase from each other. These are commonly called the quadrature "A" and "B" signals. The clockwise direction for most encoders is defined as the "A" channel going positive before the "B" channel [17].

## CHAPTER 4

### SIMULATION RESULTS AND DISCUSSION

#### 4.1 SIMULATION SETUP

The simulator for speed control of PMSM using HOSM controller was developed in MatLab/Simulink R2014a by using components from the Power System's Block set as shown in Figure 4.1. The overall system simulation block includes different sub-functional blocks: second order quasi-continuous sliding mode controller, PID controller, different transformation, and 3-phase inverter block are the main sub-functional blocks. The simulation is based on discrete system with sampling period of  $10\mu\text{s}$ .

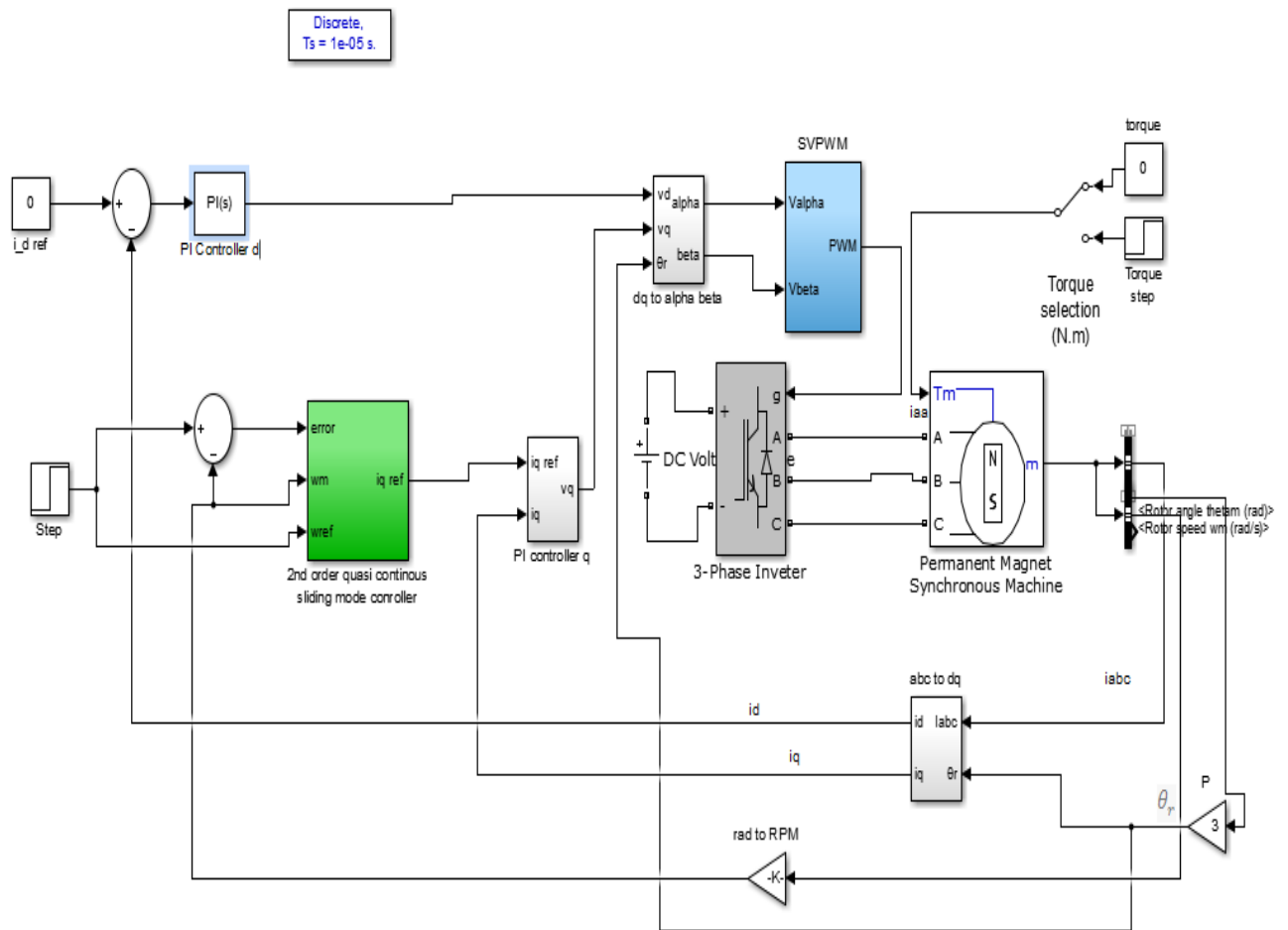


Figure 4.1: speed control of PMSM using HOSM controller system simulation block

## Speed Control of PMSM using HOSM Controller

The second order quasi-continuous sliding mode controller and the coordinate transformations block are set to implement the principle of vector control discussed in chapter 2 and the structure of the block shown in Figure 4.2 is the direct implementation of Figure 3.3. Different coordinate transformations (Clarke-park and inverse Park transforms) are vector transforms as discussed in section 2.3.2 and are taken from Matlab Simulink library. The PID blocks are PID controller in which the Proportional and Integral gain parameters are adjusted by using Ziegler and Nichols simple mathematical procedures for tuning PID controllers. These procedures are now accepted as standard in control systems practice. Ziegler-Nichols formulae for specifying the controllers are based on plant step responses (see appendix B).

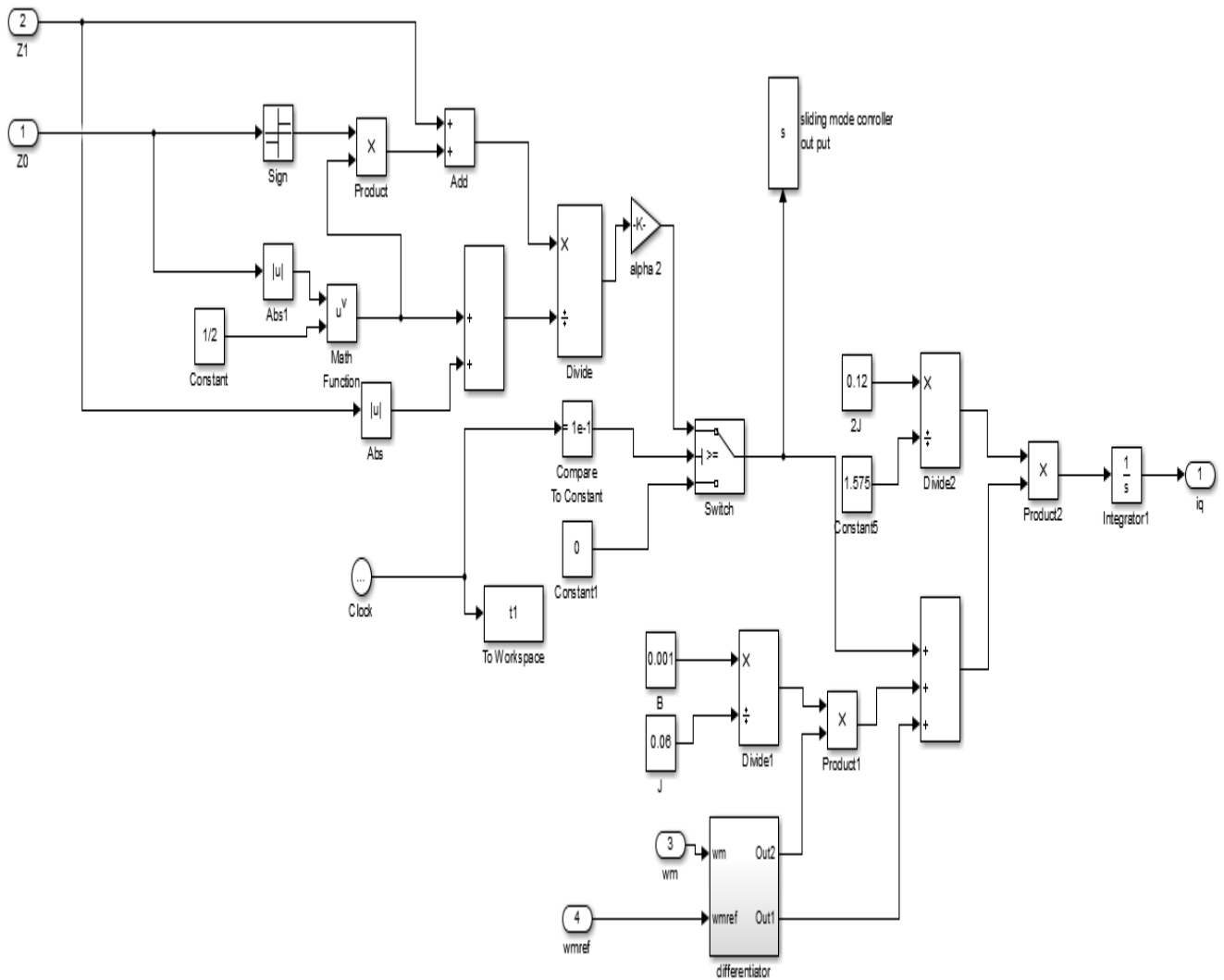


Figure 4.2: 2<sup>nd</sup> order quasi-continuous sliding mode controller block in the simulation

Figure 4.3 shows simulation block diagram of Space Vector Pulse Width Modulated signal. SVPWM block which is a Matlab source code embedded on simulation block taking state voltage in alpha and beta direction as input, the 6 PWM generated pulses as outputs. The output operation times  $T_{cma}$ ,  $T_{cmb}$  and  $T_{cmc}$  are compared with a saw tooth repeating sequence signal with the same frequency as PWM to generate pulse width modulated signal. Finally the three output signals are passed through inverter logic operator to produce the inverse of them. These six pulse width modulated signals are used to drive a three phase voltage source inverter.

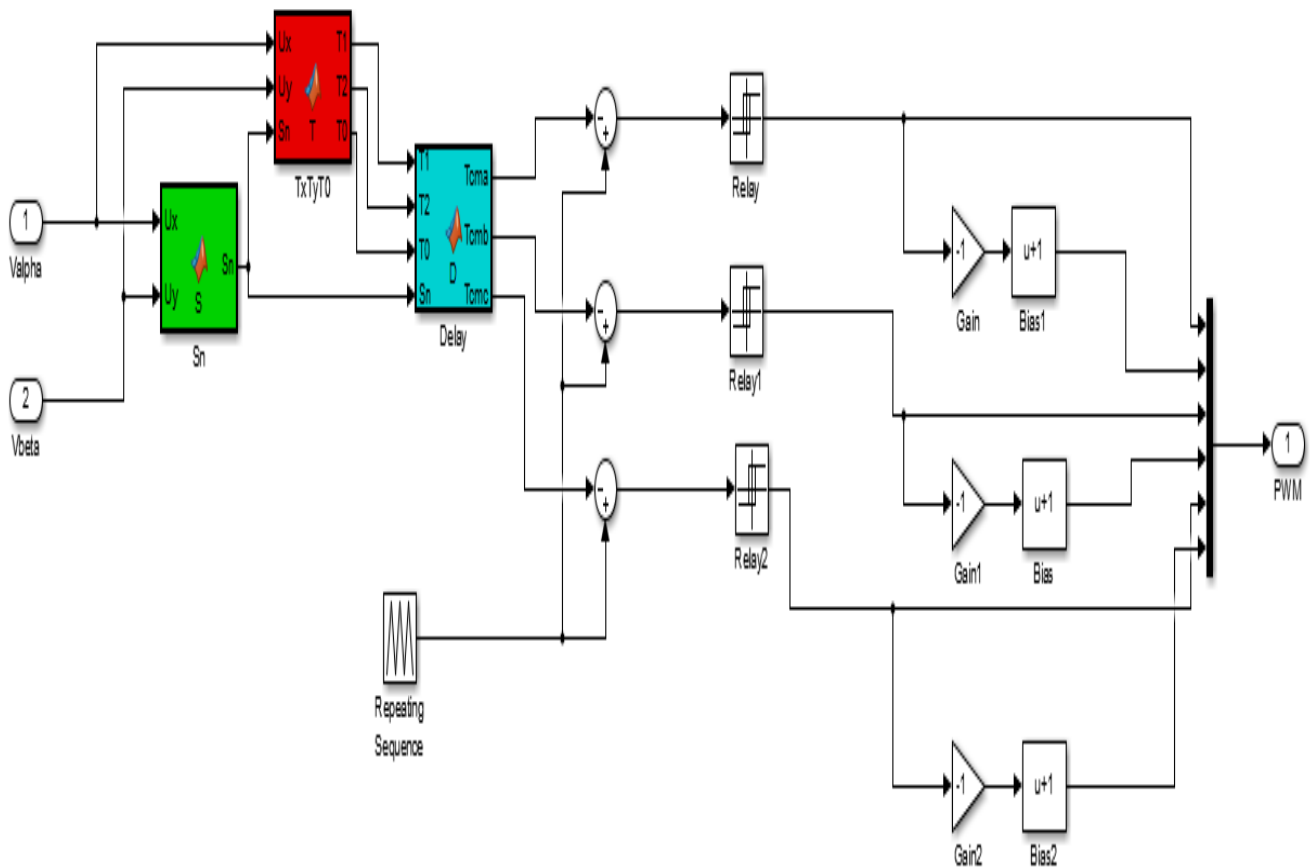


Figure 4.3: SVPWM generation simulation diagram

## 4.2 SIMULATION RESULTS

The overall Matlab/Simulink model of the PMSM allows simulating the behavior of the machine using unity feedback and quasi-continuous sliding mode controller follow with PI controller for

different operating modes. Thus, the simulation results were carried out to assess its performance. The PMSM parameters used in the simulation are found in Table A1 of Appendix A. First of all, the control law of the 2<sup>nd</sup> order quasi-continuous sliding mode controller is shown. Second, the response of the system is tested for step input. Thirdly, the three phase currents from the inverter are measured. Fourthly, the electromagnetic torque developed by the system is measured at different load. Then, the effect of parameters variation (moment of inertia, stator resistance and stator inductance) on the speed response is simulated.

To achieve a better efficiency of the proposed scheme and to convince the robustness of the system, the parameters of the controllers were chosen as follows:

- Quasi-continuous sliding mode controller gain  $\alpha = 15000$  which was obtained through simulation.
- PID controller for  $i_d$  current  $P = 47, I = 5$  and  $D = 3$ .
- PID controller for  $i_q$  current  $P = 25, I = 0.2$  and  $D = 35$ .

Figure 4.4 shows the output of the 2<sup>nd</sup> order quasi-continuous sliding mode controller. The control law still is high frequent discrete signal. But the aim of high order sliding mode makes it act on high order sliding mode surface, so that it eliminates chattering in lower order sliding mode surface.

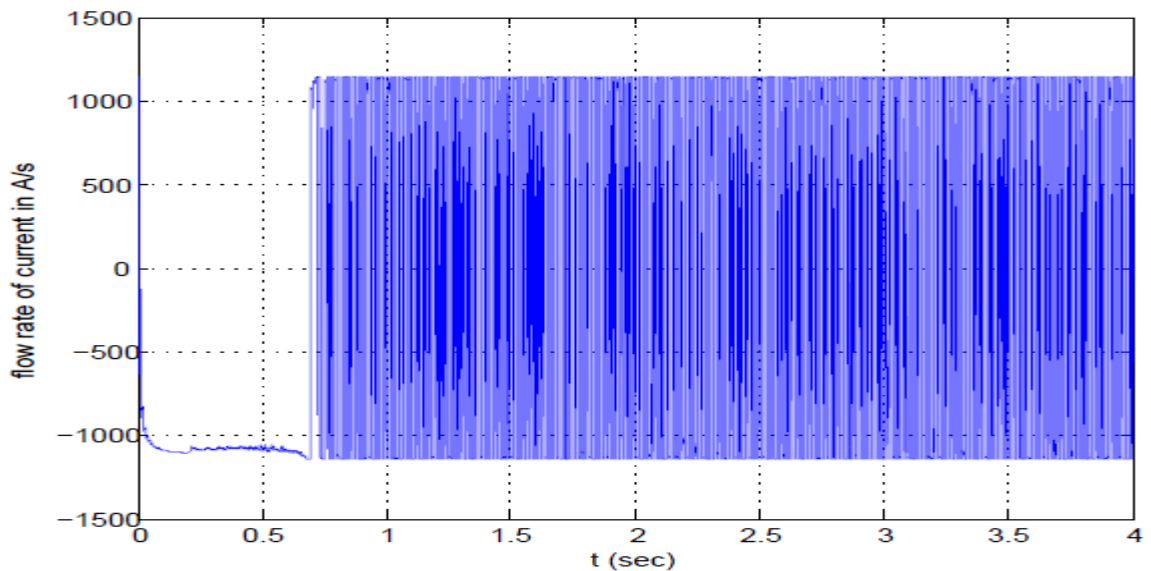


Figure 4.4: output of the 2<sup>nd</sup> order quasi-continuous controller

As shown the above figure, the controller eliminates chattering existed in traditional sliding mode essentially.

Figure 4.5: shows the speed response of the motor when the motor is commanded at a reference speed of 300 rad/sec (2865 RPM).

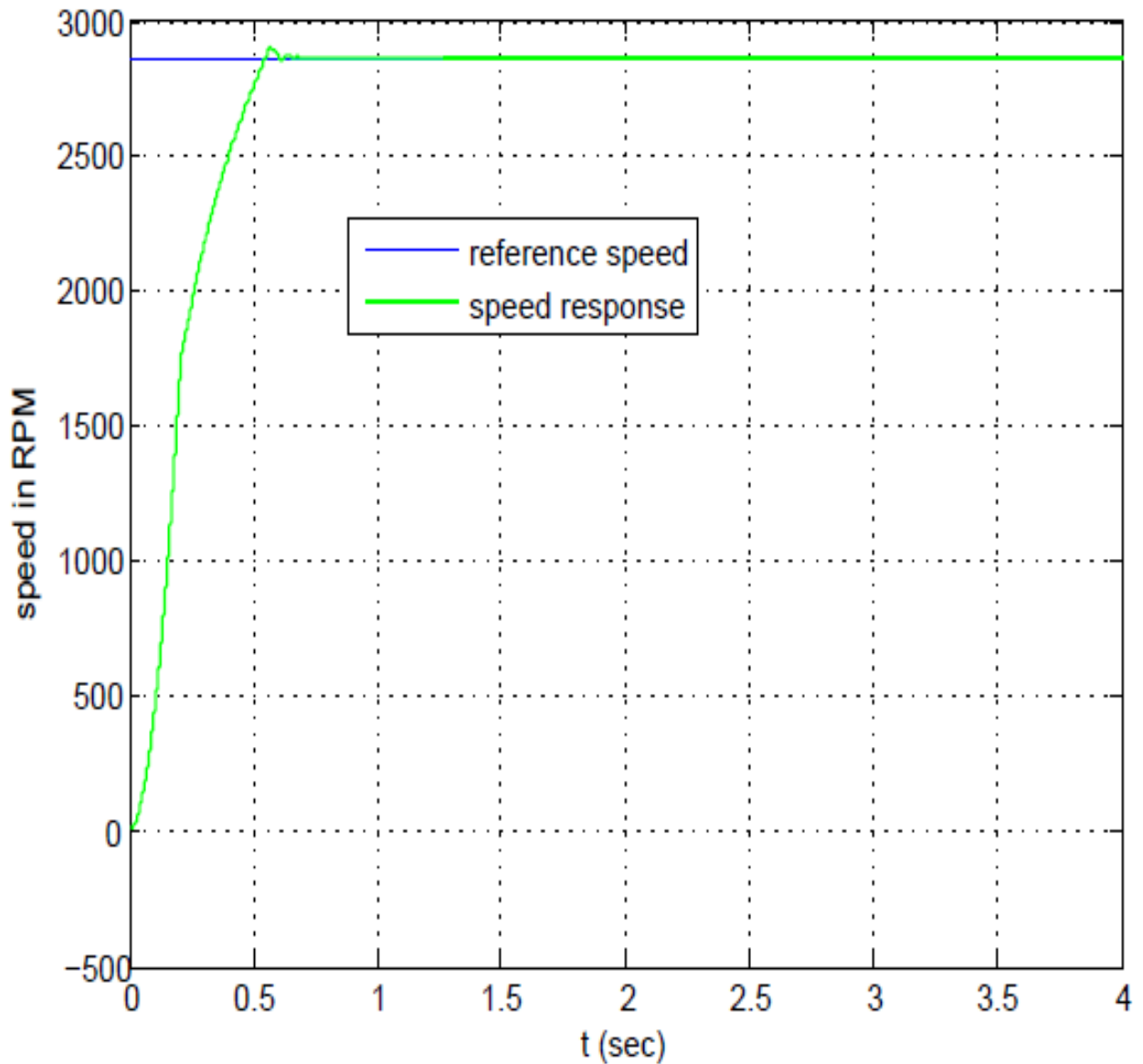


Figure 4.5: Speed response of the PMSM at no load conditions.

As we see from the above figure a variation of the motor speed with time. The steady state speed is the same as that of the commanded reference speed. The system can follow the reference

signal with a rise time of less than 0.54 second, settling time of 0.65 second and has 1.08% maximum overshoot value and this indicates that the system has good transient and steady state response.

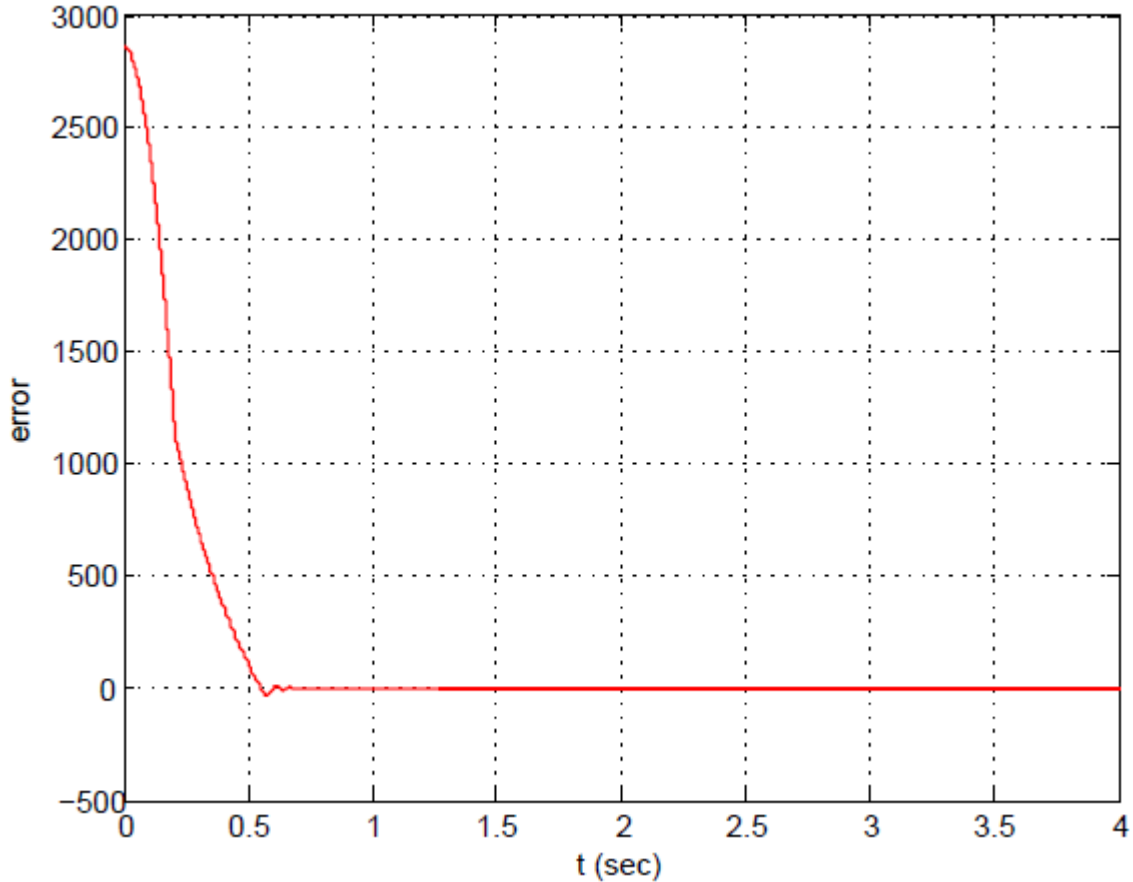


Figure 4.6: error between Speed response of the PMSM and reference speed at no load.

Another simulation result is the three phase stator currents which are generated by the three phase inverter. This three phase inverter is controlled by SVPWM block for appropriate stator current and voltage generation. These three phase currents should be equal magnitude and  $60^\circ$  phase shift with each other for appropriate rotating flux generation as shown in Figure 4.7, and the detail view it is also shown in Figure 4.8 below.

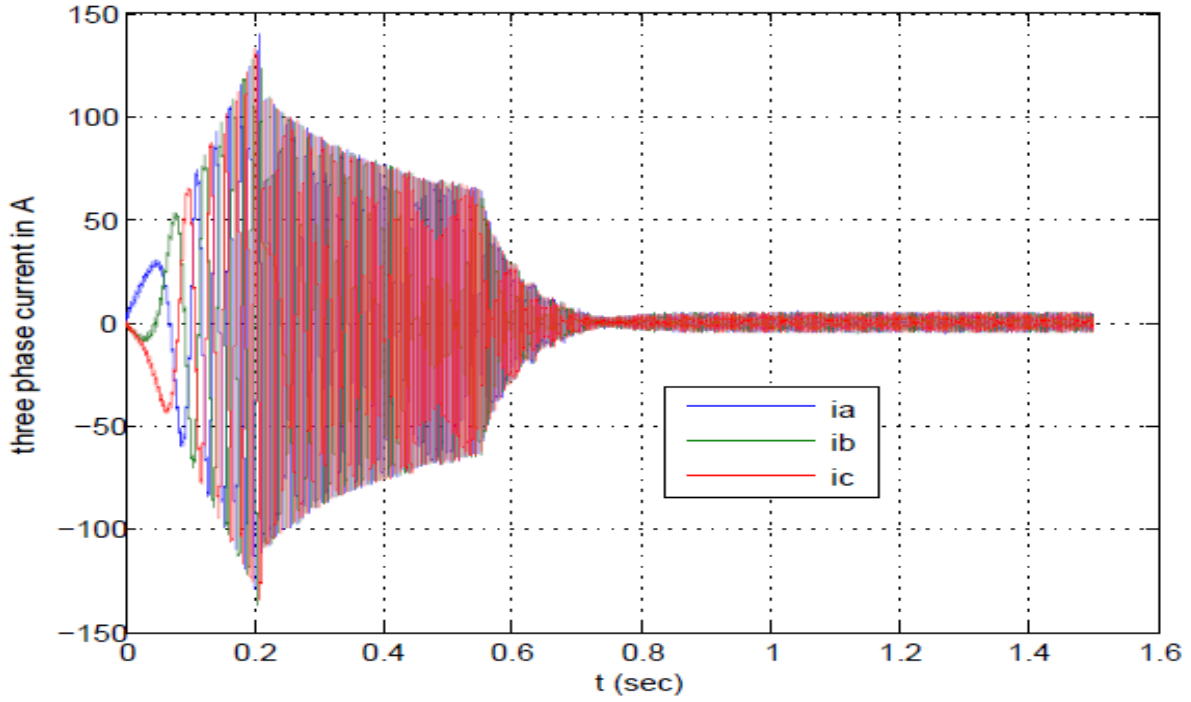


Figure 4.7: Three phase stator current while the motor is running at no-load

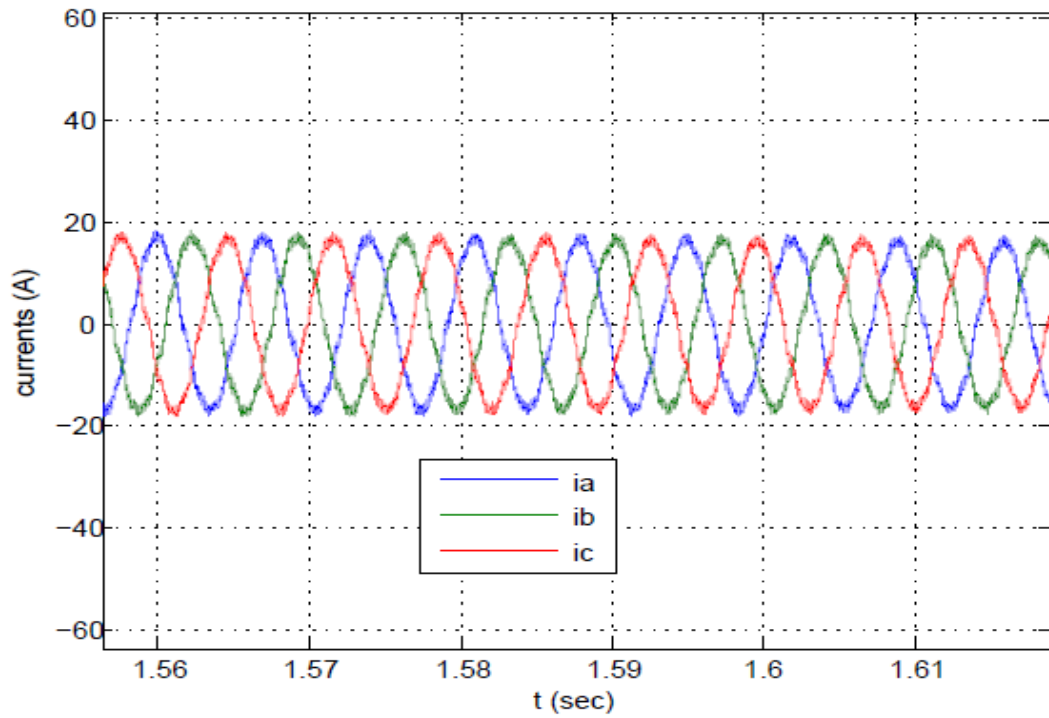


Figure 4.8: Three phase currents zoom out between 1.55 and 1.62 sec at no load

As we see in the above Figure, the appropriate stator phase current value is generated with good accuracy. Hence the system can feed the appropriate stator voltage to the motor.

Figure 4.9 shows the developed electromagnetic torque of the motor at no load condition. The starting torque until 0.66 sec is varied which is due to primarily the acceleration of the rotor to reach to the steady speed of 2865 rpm. The starting torque is reached 98.8 Nm which is higher compare to the steady state value. However, after 0.66sec the generated torque is reduced almost to 0 Nm to support only the approximately zero retard friction.

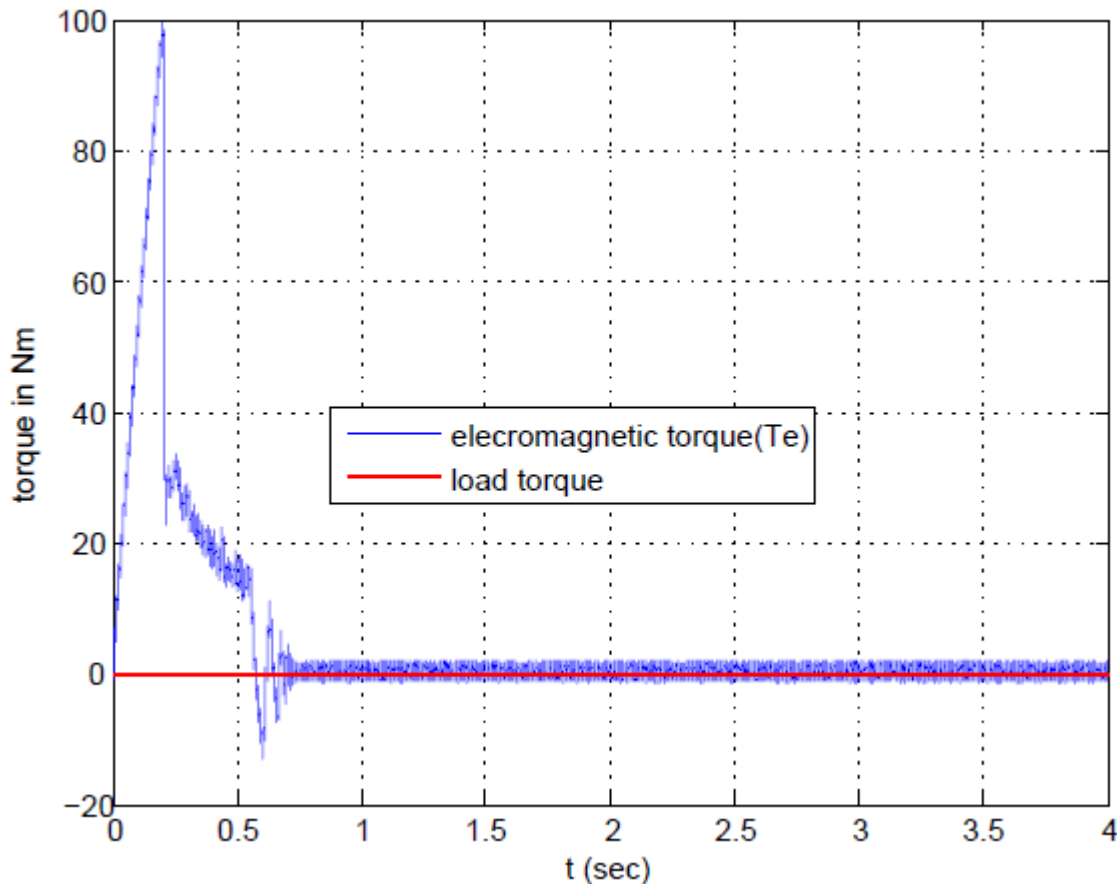


Figure 4.7: Developed electromagnetic torque at no load condition.

Figure 4.10 shows the speed response tracks a given step input from 900 RPM to 2865 RPM. As shown in the figure, the speed response of the system desires for two step speed level. First the speed is set to be 900 RPM. Then the speed changes instantly at  $t = 2$  sec and attains its steady speed of 2865 RPM. When we see the performance of the system, it has a better performance in

settling time, rise time and maximum over shoot when the motor runs on the first level step input i.e at 900 RPM which is below the rated speed as compared to the rated speed response as shown in figure 4.5. So the result indicates that the system is reduced chattering effect as shown in the figure.

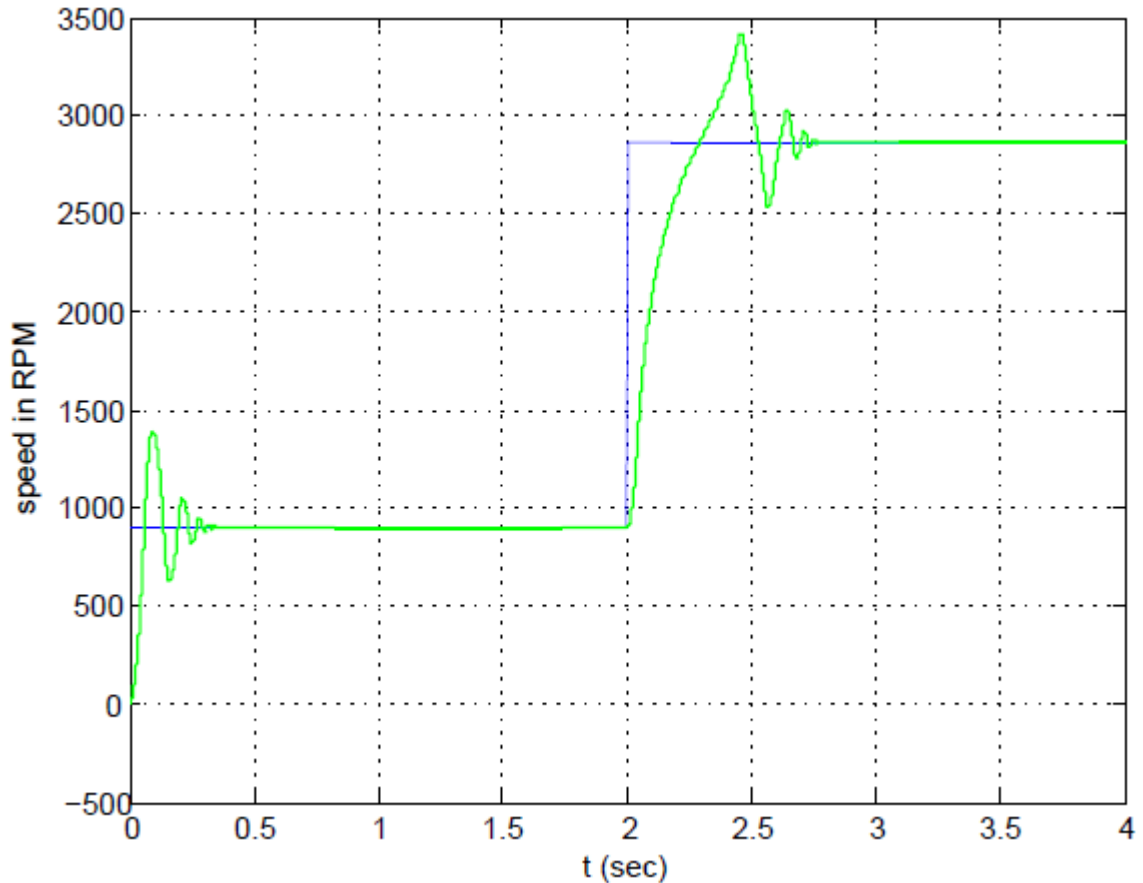


Figure 4.10: Speed response for step input from 900 RPM to 2865 RPM at no load.

The system also works in forward and reverse condition and has good speed tracking in both cases. Figure 4.11 shows simulation result while the motor is running in 2865 rpm forward and reverse condition. When we compare the performance of the system, during in the forward condition the system has a better performance compared to the reverse condition in some control parameters such as on rising time and settling time. But the maximum overshoot in both conditions are similar.

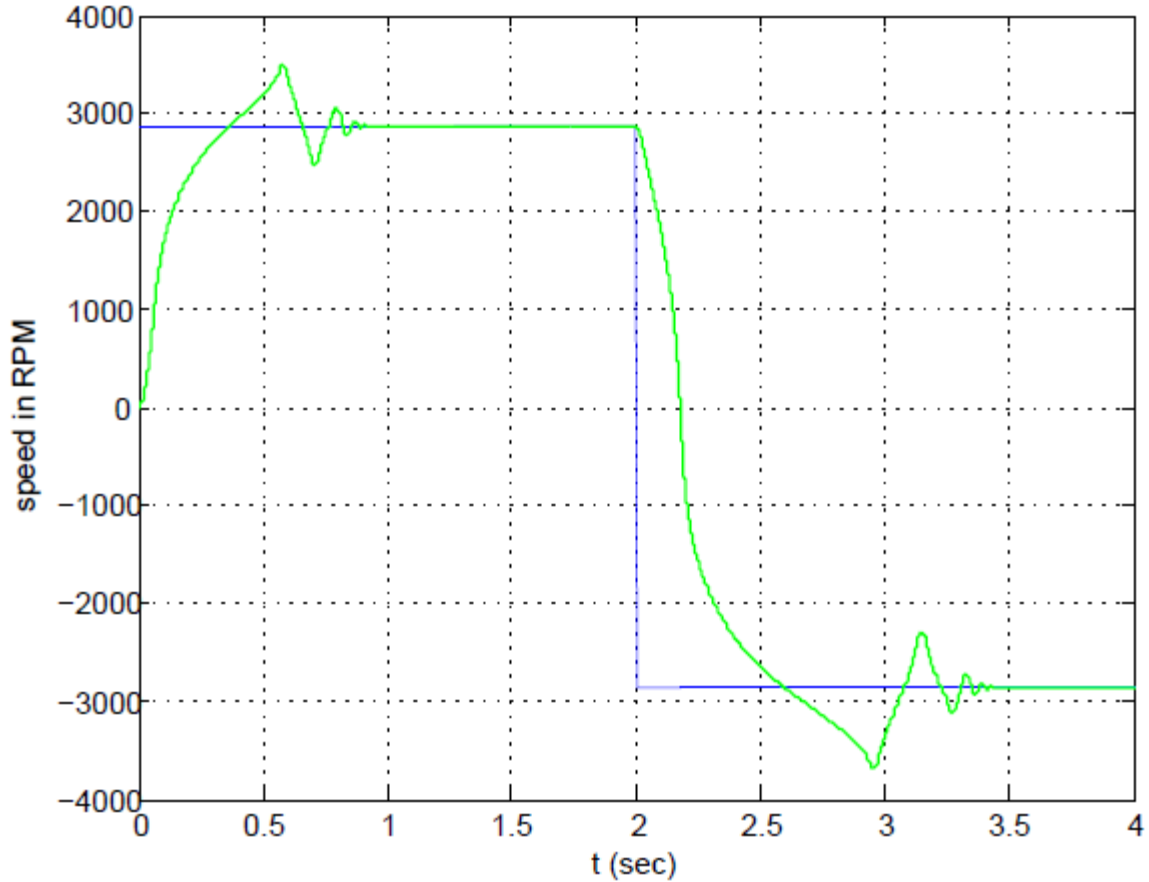


Figure 4.11: speed response for forward and reverse condition at no load condition.

Figure 4.12 illustrates the developed electromagnetic torque and load torque of the system in load torque conditions. Still the developed electromagnetic torque is varied on the transient response due to the acceleration of the motor as discussed above in figure 4.9. But the electromagnetic torque takes 0.78 sec to come to the steady state response (5Nm) which is a longer time as compared to the no load conditions.

As stated in Chapter Two, the advantage of vector control is directly controlling the motor torque by using the stator current like DC motor control. Figure 4.12 and Figure 4.13 shows this fact in which the Electromagnetic torque is directly proportional with the q-axis current. The simulation was taken place with a load torque of 5Nm.

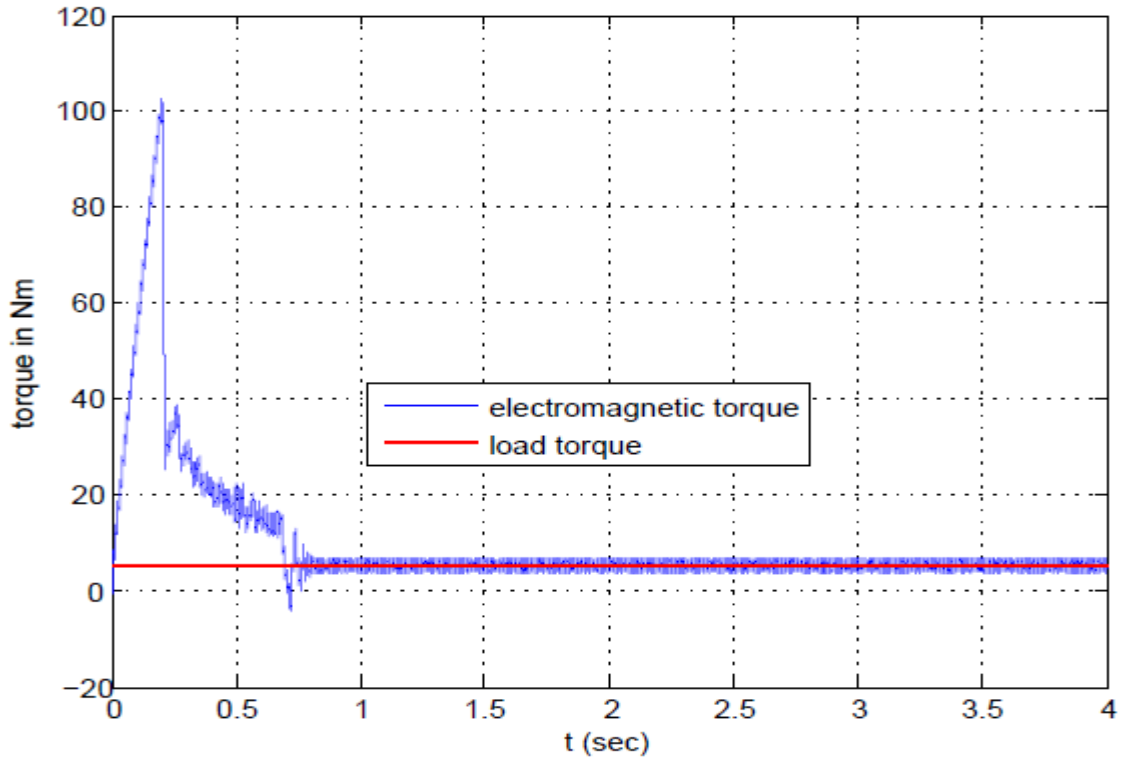


Figure 4.12: Developed electromagnetic torque at load torque ( $T_l=5\text{Nm}$ ) condition.

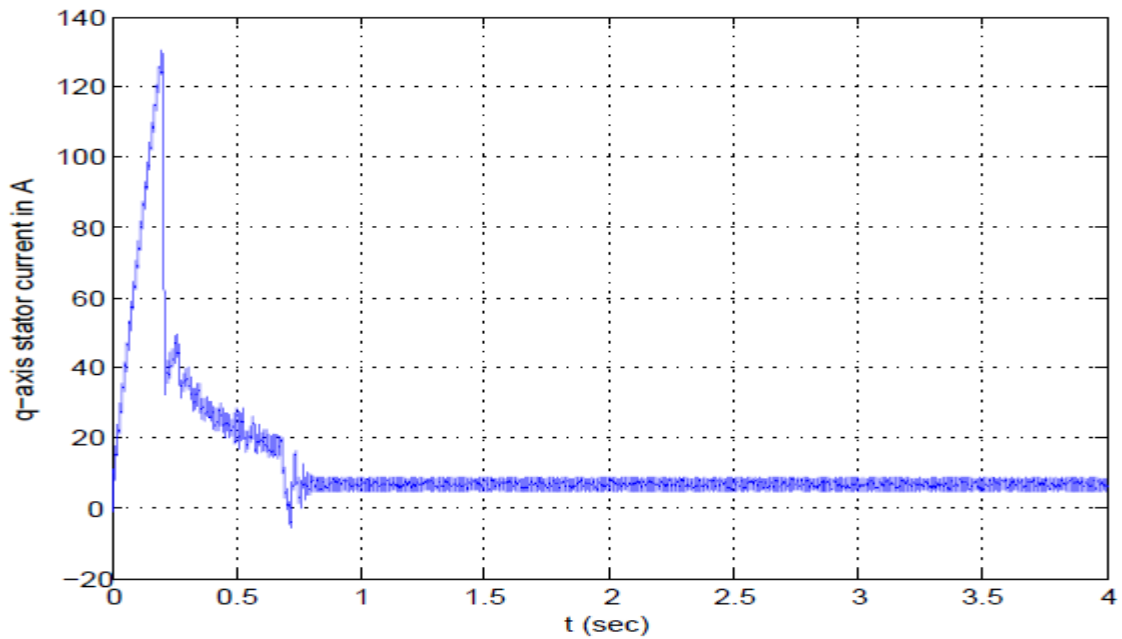


Figure 4.13 q-axis stator current in A

Figure 4.14 shows the speed response of the system at 5 Nm torque load. The performance of the system is also good in control parameters which have a rise time of less than 0.68 sec, a settling time of 0.75 sec, maximum over shoot value of 0.48%.

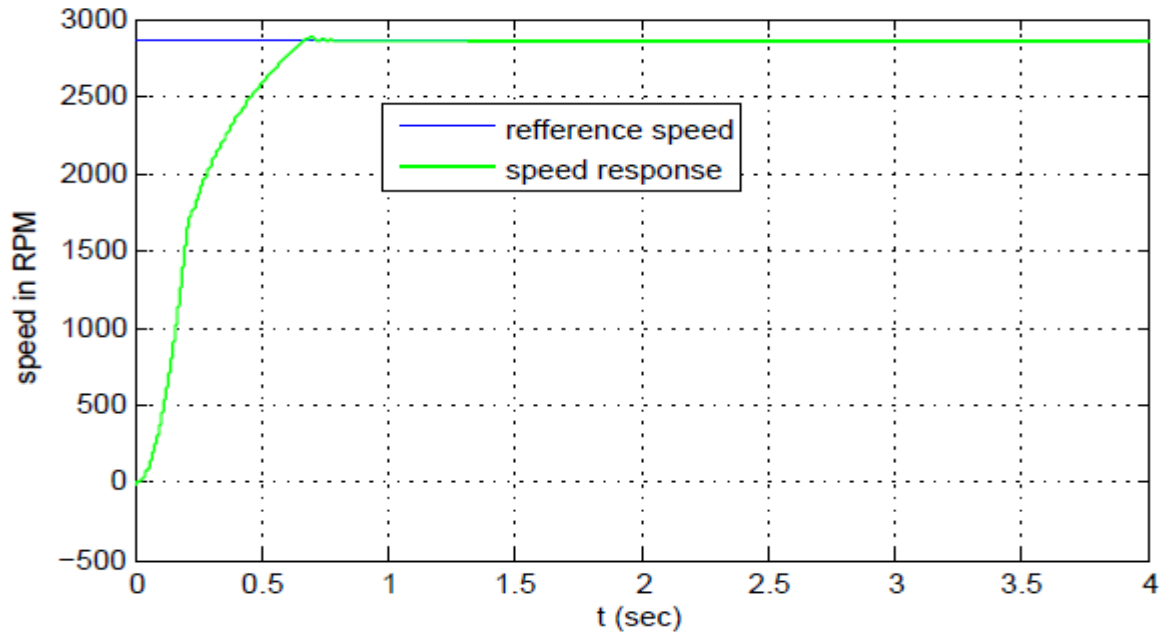


Figure 4.14 Speed response of the PMSM at 5Nm load torque.

The Stator phase currents of the system at loaded condition also have good transient and steady state response. The following figure 4.15 and figure 4.16 show these properties.

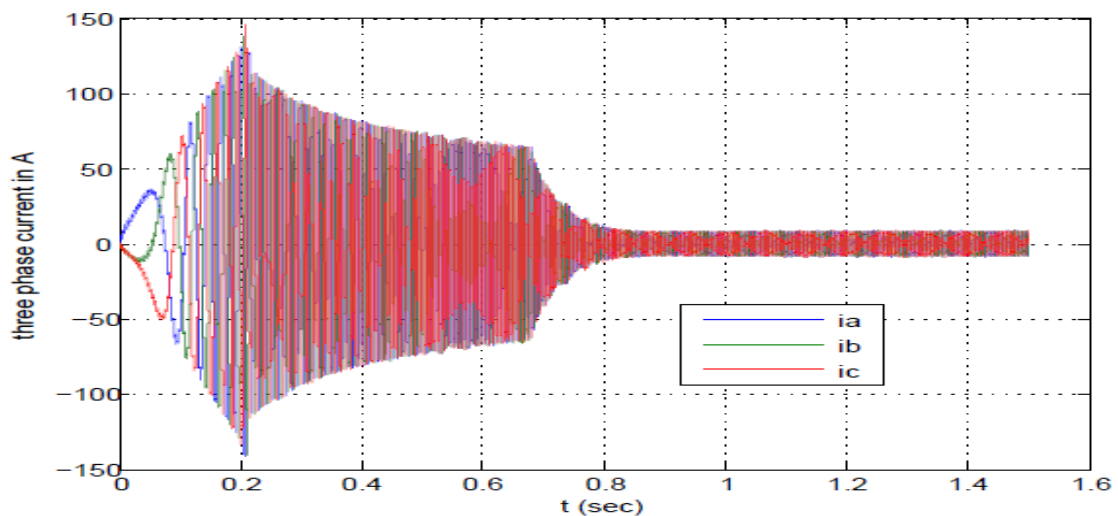


Figure 4.15: Three phase stator current while the motor is running at 5Nm load torque.

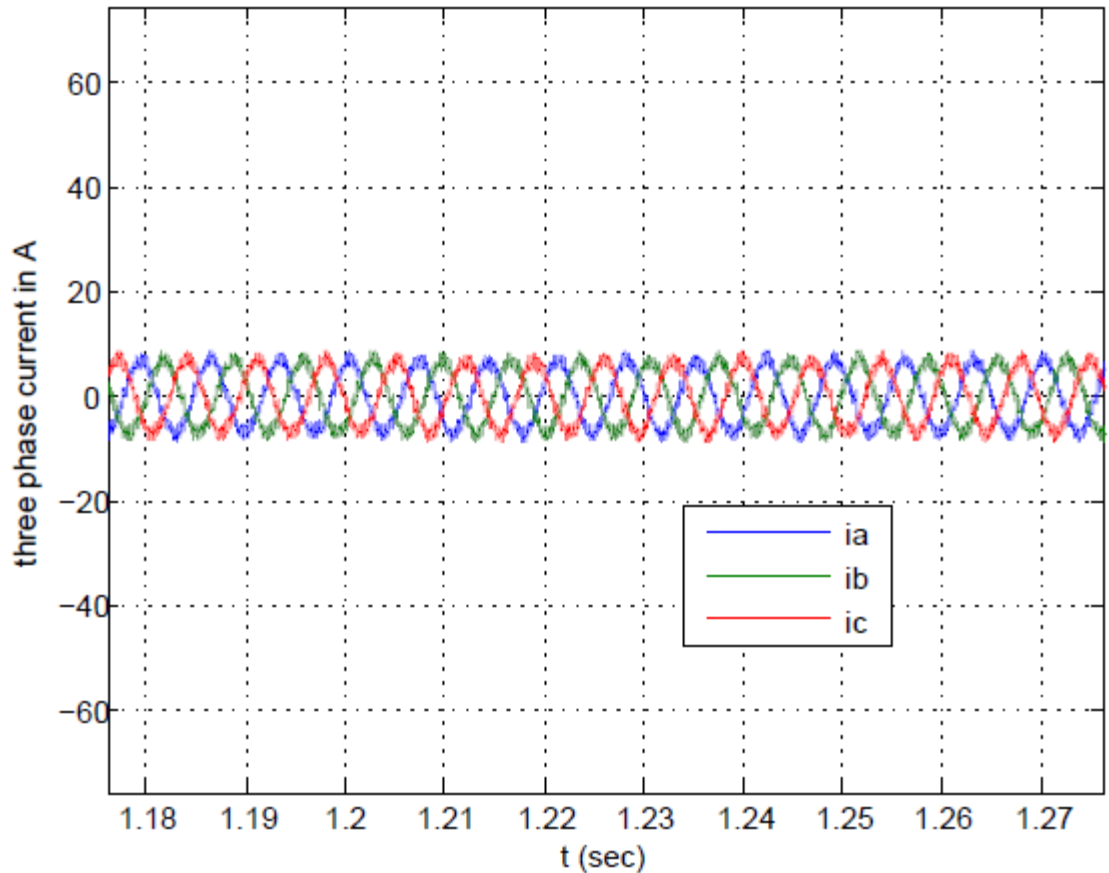


Figure 4.16: Three phase currents between 1.17 and 1.28 sec at 5 Nm load condition.

From figure 4.17 up to figure 4.19 shows the performance of the system because of the parameter variation of the motor due to different reasons such as temperature. Figure 4.17, 4.18 and 4.19 present the speed response of the system when the motor stator resistance, inductance and moment of inertia are increased by 50% of the nominal value without the load torque respectively. From those figures we can see that the PMSM control system has good performance and robust. As it is shown from the three figures, the permanent magnet synchronous motor can precisely regulate the given speed.

From the figures we can clearly see that the discrete control law acts on the high order sliding mode surface, which makes the lower sliding mode surface smooth. That is the reason why high order sliding mode control can reduce the chattering. Generally, the system has good transient and steady state response in all conditions.

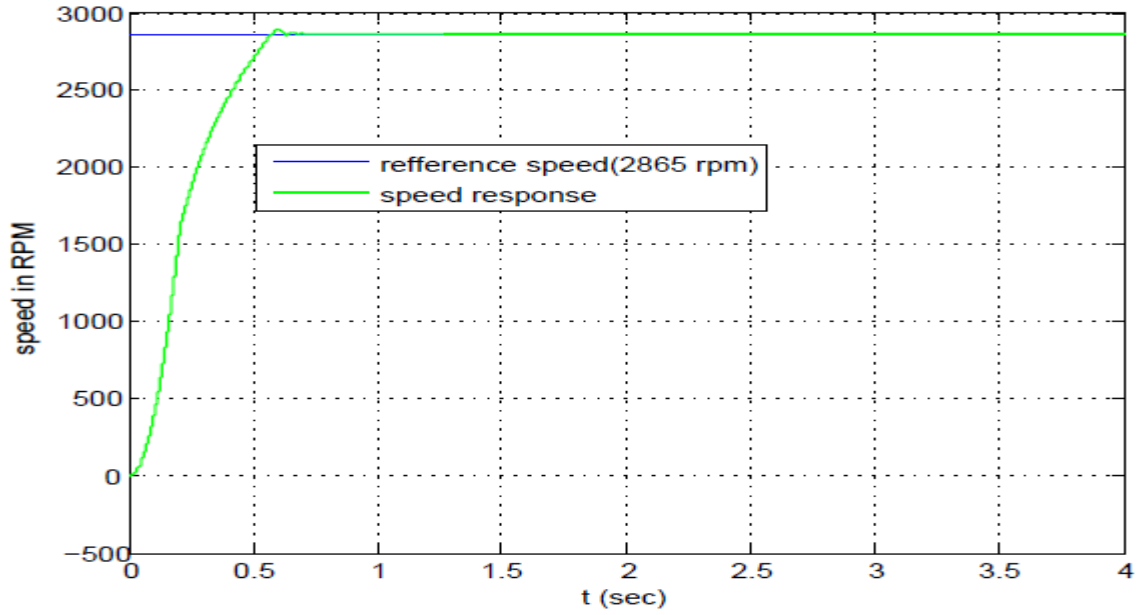


Figure 4.17: Speed response of the motor when the stator resistance increased by 50%.

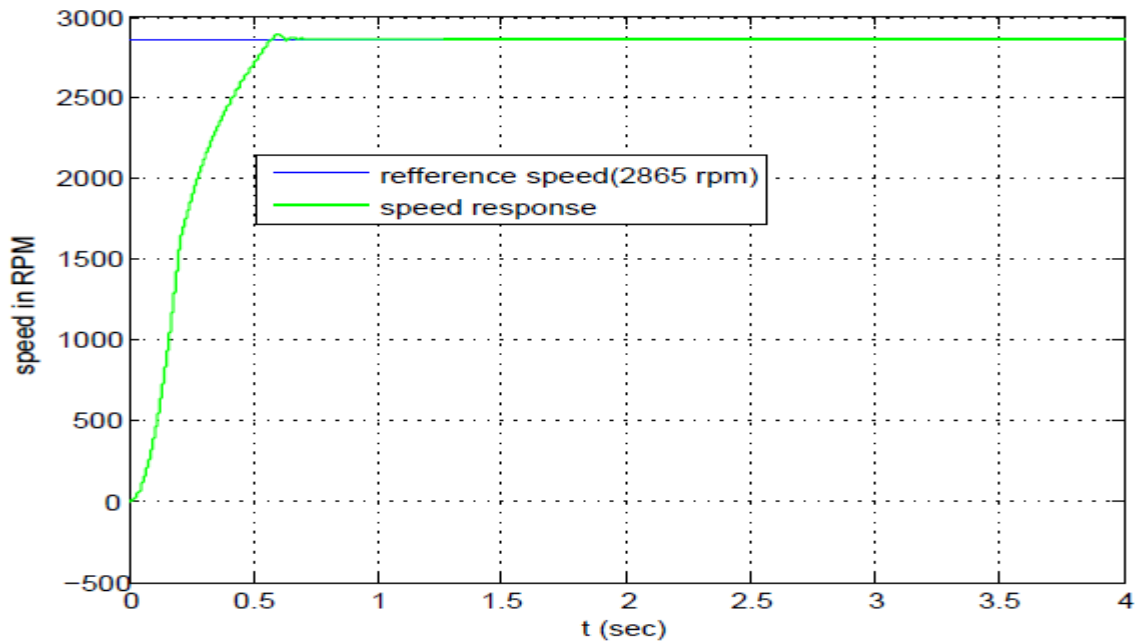


Figure 4.8: Speed response of the motor when the stator inductance increased by 50%.

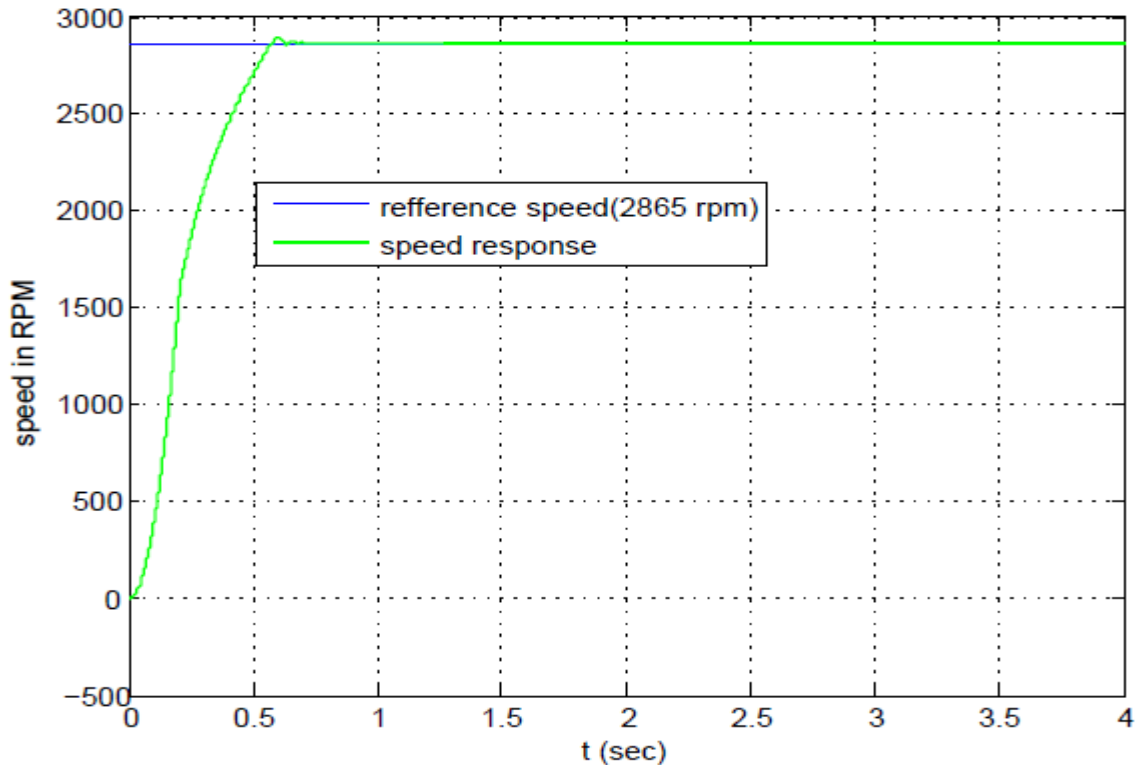


Figure 4.9: Speed response of the motor when the moment of inertia increased by 50%.

### 4.3 SUMMARY ON THE SIMULATION RESULTS

On the whole the simulation results were very positive. These are the findings:

- The system achieves robust control for a PMSM in spite of the internal parameter uncertainties and unknown external disturbance load torque.
- The simulation results showed that our controller has a good performance for a given reference input in terms of settling time, rising time, steady state error and maximum over shoot.
- The simulation results showed that the motor has good speed tracking.
- Three phase inverter is controlled by SVPWM algorithm has a good performance as we can see from Figure 4.7 and 4.16 the generated current has  $60^\circ$  phase shift and equal magnitude.

From these results we can conclude that the FOC or vector control algorithm has a good performance.

## CHAPTER 5

### CONCLUSIONS AND FUTURE WORKS

#### 5.1 CONCLUSIONS

This thesis presented a speed control of a permanent magnet synchronous motor using higher order sliding mode controller. The PMSM system is a MIMO nonlinear and time-varying complex system. Using PI controller as speed control of PMSM is not satisfactory to the higher degree of accuracy condition. So a high order sliding mode control is designed, in order to reduce the chattering effect and assure the robustness of the system so that it gives a better response and good result in all conditions.

Thus a higher order sliding mode controller not only has the prominent advantage in complex, time-varying and nonlinear system control but also preserves the strong robustness of the other control techniques, which is not sensitive to the uncertain parameters of systematic model and the unknown external disturbances. But in comparison with the conventional sliding mode control, high order sliding mode control especially the quasi-continuous sliding mode controller technique uses high order differential of sliding mode variable to eliminate the chattering effect.

The complete system has been simulated using the software package Matlab/Simulink. The simulation encompasses the entire system: the controller, different coordinate transformation, the inverter, and the PMSM.

The simulations are performed and investigated on the following mode:

- System performance and the speed response, the developed electromagnetic torque and the three phase stator currents at no load and loaded torque condition.
- System performance and the speed response by varying the plant parameters such as stator resistance, stator inductance and moment of inertia.
- The speed response of the system at reverse, forward and a step input at instant of time.

Generally, the simulation results show that a speed control of a PMSM using HOSM controller have good performance since from figure 4.5 & 4.14 the system has a settling time of less than 1 second, maximum overshoot of 1.08%, and fast recovery from all modes which are listed above. Hence the system can work with different load torque and different environment condition. Significant robustness tests are implemented and simulation results illustrate the efficiency of the controller scheme.

### **5.2 Future works**

Recommendations for future works are listed as follows:

- An online parameter identification of the system should be applied in order to improve the efficiency of the system.
- Performing experimental setup and analyzing the result on the lab room.

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## APPENDICES

### Appendix A

#### PMSM PARAMETERS

Table A1: PMSM Motor Parameters.

Parameters	Values	Parameters	Values
Number of poles ( $p$ )	2x3	friction coefficient	0.001Nms
Stator Inductances	0.008H	rotor inertia	0.06JKgm <sup>2</sup>
Flux distribution type:	sinusoidal	Voltage constant	95.2445V/krpm
Stator resistance	1.456	rotor flux linkage	0.175 Vs
Torque constant:	0.7875Nm/A		

## Appendix B

### Ziegler-Nichols method

The Ziegler-Nichols method is common used in tuning of PI/PID controllers. The proposed ruled of Ziegler-Nichols method is for determining values of the proportional  $k_p$  and integral time  $T_i$  based on the transient response characteristics of the plant. Such as the determination of the parameters of PI controllers or tuning of PI controller can be made on site by experiment on plant. The technique is designed to result in a closed loop system with 25% overshoot. The steps for tuning a PI controller via the 2<sup>nd</sup> method are as follows:

Using only proportional feedback control:

1. Reduce the integrator and derivative gains to 0.
2. Increase  $k_p$  from 0 to some critical value  $k_p = k_{cr}$  at which sustained oscillations occur.
3. Note the value  $k_{cr}$  and the corresponding period of sustained oscillation,  $P_{cr}$

The controller gains are now specified as follows:

Table B1: Ziegler Nichols method

PID type	$k_p$	$T_i$	$T_d$
P controller	$0.5 k_{cr}$	0	0
PI controller	$0.45 k_{cr}$	$\frac{P_{cr}}{1.2}$	0
PID controller	$0.6 k_{cr}$	$\frac{P_{cr}}{2}$	$\frac{P_{cr}}{8}$

## DECLARATION

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

Derege Eskezia

Name

\_\_\_\_\_

Signature

**Place:** Addis Ababa

**Date of Submission:** \_\_\_\_\_

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

Dr.- Ing. Mengesha Mamo

Advisor's Name

\_\_\_\_\_

Signature

Ato Andinet Negash

Co-dvisor's Name

4/30/2016

Signature