SLIDING MODE VECTOR CONTROL OF THREE PHASE INDUCTION MOTOR

A THESIS SUBMITTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF

Master of Technology in 'Power Control and Drives'
By

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Department of Electrical Engineering
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Under the guidance
of

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CERTIFICATE

This is to certify that the work in this thesis entitled "**Sliding Mode Vector Control of three phase induction motor**" by **Sobha Chandra Barik**, has been carried out under my supervision in partial fulfillment of the requirements for the degree of Master of Technology in '***Power Control and Drives***' during session 2006-2007 in the Department of Electrical Engineering, National Institute of Technology, Rourkela and this work has not been submitted elsewhere for a degree.

To the best of my knowledge and belief, this work has not been submitted to any other university or institution for the award of any degree or diploma.

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Sobha Chandra Barik
M.Tech (Power Control and Drives)

ABSTRACT

Sliding Mode Control (SMC) is a robust control scheme based on the concept of changing the structure of the controller in response to the changing state of the system in order to obtain a desired response. A high speed switching control action is used to switch between different structures of the controller and the trajectory of the system is forced to move along a chosen switching manifold in the state space.

This thesis work presents a new sensorless vector control scheme consisting on the one hand of a speed estimation algorithm which overcomes the necessity of the speed sensor and on the other hand of a novel variable structure control law with an integral sliding surface that compensates the uncertainties that are present in the system.

In this work, an indirect field-oriented induction motor drive with a sliding-mode controller is presented. The design includes rotor speed estimation from measured stator terminal voltages and currents. The estimated speed is used as feedback in an indirect vector control system achieving the speed control without the use of shaft mounted transducers. Stability analysis based on Lyapunov theory is also presented, to guarantee the closed loop stability. The high performance of the control scheme under load disturbances and parameter uncertainties is also demonstrated.

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LEGENDS

B	Viscous friction coefficient
$d^e - q^e$	Synchronously rotating reference frame (or rotating frame) direct and quadrature axes
$d^s - q^s$	Stationary reference frame direct and quadrature axes.
f	Frequency (Hz)
I_f	Machine field current.
I_s	Rms stator current.
i_{dr}^s	d^s - axis rotor current.
i_{ds}^s	d^s - axis stator current
i_{qr}^s	q^s - axis rotor current.
i_{qs}^s	q^s - axis stator current.
i_{qr}	q^e - axis rotor current.
i_{dr}	d^e - axis rotor current
i_{ds}	d^e - axis stator current
i_{qs}	q^e - axis stator current
i_{abc}^*	Command current
i_{abc}	Actual current
J	Moment of inertia.
K	Constant gain.
K_T	Torque constant
β	Switching gain
θ_e	Angle of Synchronously rotating frame
θ_r	Rotor angle

θ_{sl}	Slip angle
L_m	Magnetizing inductance
L_r	Rotor inductance
L_{lr}	Rotor leakage inductance
L_{ls}	Stator leakage inductance
P	Number of poles
R_r	Rotor resistance
R_s	Stator resistance
S	Sliding variable
T_e	Developed torque
T_L	Load torque
T_{em}	Electro magnetic torque
T_r	Rotor time constant
V	Lyapunov function
v_{dr}^{s}	d^s -axis rotor voltage
v_{qr}^{s}	q^s - axis rotor voltage
v_{ds}^{s}	d^s - axis stator voltage
v_{qs}^{s}	q^s - axis stator voltage
v_{qr}	q^e - axis rotor voltage
v_{dr}	d^e - axis rotor voltage
v_{qs}	q^e - axis stator voltage
v_{ds}	d^e - axis stator voltage
Ψ_a	Armature reaction flux linkage
Ψ_f	Field flux linkage

Ψ_m	Airgap flux linkage
Ψ_r	Rotor flux linkage
Ψ^s_{dr}	d^s - axis rotor flux linkage
Ψ^s_{qr}	q^s - axis rotor flux linkage
Ψ^s_{ds}	d^s - axis stator flux linkage
Ψ^s_{qs}	q^s - axis stator flux linkage
Ψ_{qr}	q^e - axis rotor flux linkage
Ψ_{dr}	d^e - axis rotor flux linkage
Ψ_{qs}	q^e - axis stator flux linkage
Ψ_{ds}	d^e - axis stator flux linkage
ω_e	Stator frequency
ω_m	Rotor mechanical speed
ω_r	Rotor electrical speed
ω_{sl}	Slip frequency
σ	Motor leakage coefficient

ABBREVIATION AND ACRONYMS

AC	Alternating Current
CEMF	Counter Electromotive Force
DC	Direct Current
DSP	Digital Signal Processing
EKF	Extended Kalman Filter
FOP	Field Oriented Principle
IGBT	Insulated Gate Bipolar Transistor
IM	Induction Motor
MRAC	Model Reference Adaptive Control
MRAS	Model Reference Adaptive System
PWM	Pulse Width Modulation
SMC	Sliding Mode Control
VR	Vector Rotation

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CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION:

Power semiconductor devices constitute the heart of modern power electronic apparatus. They are used in power electronic converters in the form of a matrix of on off switches, and help to convert power from ac to dc,dc to dc and ac to ac at the same or different frequencies. The switching mode power conversion gives high efficiency; the disadvantage is that due to the nonlinearity of the switches, harmonics are generated at both the supply and load sides. The switches are not ideal, and they have conduction and turn on and turn off switching losses. Converters are widely used in application such as heating and lighting controls,ac and dc power supplies, electrochemical processes, dc and ac motor drives, static var generation, active harmonic filtering,etc. Although the cost of the power semiconductor devices in power electronic equipment may hardly exceed 20-30 percent, the total equipment cost and performance may be highly influenced by the characteristic of the devices. An engineer designing equipment must understand the devices and their characteristics thoroughly in order to design efficient, reliable, and cost effective system with optimum performances. It is interesting to note that the modern technology evolution in power electronics has generally followed the evolution of power semiconductor devices. The advancement of microelectronics has greatly contributed to the knowledge of power device materials, processing, fabrication, packaging, modeling, and simulation.

In the past, dc motors were used extensively in areas where variable speed operation was required, since their flux and torque could be controlled easily by the field and armature current. In particular, the separately exited dc motor has been used mainly for applications where there was a requirement of fast response and four quadrant operation with high performance near zero speed. However the dc motors have certain disadvantage, which are due to the existence of the commutator and the brushes. That is, they require periodic maintenance; they can not be used in explosive or corrosive environments and they have limited commutator capability under high speed, high voltage operational conditions. These problems can be overcome by the application of alternating current motors, which can be simple and rugged structure, high maintainability and economy; they are also robust and immune to heavy overloading.

Their small dimension compared with dc motors allows ac motors to be designed with substantially higher output ratings for low weight and low rotating mass.

Variable speed ac drives have been used in the past to perform relatively undemanding roles in application which preclude the use of dc motors, either because of the working environment or commutator limits. Because of the high cost of efficient, fast switching frequency static inverter, the lower cost of ac motors has also been a decisive economic factor in multi motor systems. However as a result of the progress in the field of power electronics, the continuing trend is towards cheaper and more effective power converters, and a single motor ac drives complete favorably on a purely economic basis with a dc drives.

Among the various ac drive systems, those which contain the cage induction motor have a particular cost advantage. The cage motor is simple and rugged and is one of the cheapest machines available at all power ratings. Owing to their excellent control capabilities, the variable speed drives incorporating ac motors and employing modern static converters and torque control can well complete with high performance four quadrant dc drives.

It is expected that with the rapid developments in the field of microelectronics ,torque control of various types of ac machines will become a commonly used technique when, even though high dynamic performance is not required, servo like high performance plays a secondary role to reliability and energy saving. It is possible to contribute to the energy saving by the application of intelligent control of the flux and torque producing components of the stator currents.

In the case of dc drives, the power circuits are relatively uniform and in most cases contain a line commutated thyristor converter or a transistorized chopper for low power application. However, for ac drives there is much greater variety, due to the different types of converter voltage source, current source, natural commutation, forced commutation, dc link, cycloconverter, which can be combined with various type of ac machines.

Adjustable speed AC drives have become the preferred choice in many industrial applications where controlled speed is required. At the same time, the maturing of the technology and the availability of fast and efficient solid state power semiconductor switches (IGBTs) has resulted in voltage source, PWM controlled inverters becoming a standard configuration in the power range to 500kW. While high frequency PWM control represents the most advanced drive concept, when inappropriately applied, it also generates side effects, some which have been recognized only recently.

This course presents a comprehensive coverage of application issues of PWM inverter controlled ac motor drives which include: damage to motor insulation due to reflected voltages caused by long motor leads; the mechanism of motor bearing failures due to excessive common mode dv/dt and leakage currents to ground.

The control and estimation of ac drives in general are considerably more complex than those of dc drives, and this complexity increases substantially if high performances are demanded. The main reasons for this complexity are the need of variable frequency harmonically optimum converter power supplies, the complex dynamics of ac machines, machine parameter variation, and the difficulties of processing feedback signals in the presences of harmonics. While considering drive application we need to address followings.

- One , two or four quadrant drive
- Torque, speed or position control in the primary or outer control loop.
- Single or multi motor drive.
- Range of speed control.
- Accuracy and response time.
- Robustness with load torque and parameter variation.
- Control with speed sensor or sensor less control.
- Type of front end converter.
- Efficiency, cost, reliability, and maintainability consideration.
- Line power supply, harmonics, and power factor consideration.

In general an electric motor can thought of as a controlled source of a torque. Accurate control of instantaneous torque produced by a motor is required in high performance drive system, e.g. those used for position control. The torque developed in the motor is a result of the interaction of between the current in the armature winding and the magnetic field produced in the field system of the motor. The field should be maintained at a certain optimal level, suffiently high to yield a high torque per unit ampere, but not too high to result in excessive saturation of the magnetic circuit of the motor. With fixed field, the torque is proportional to the armature current.

Independent control of the field and armature current is feasible in separately exited dc motor where the current in the stator field winding determines the magnetic field of the motor, while the current in the rotor winding can be used as a direct means of torque control. The physical disposition of the brushes with respect to the stator field ensures optimal condition for torque production under all condition. Even today most high performance drive performance is still based on dc motor.

Vector control techniques incorporating fast microprocessor and DSPs have made possible the application of the induction motor and synchronous motor drives for high performance applications where traditionally only dc drives were applied. In the past such control technique would have not been possible because of the complex hardware and software required to solve the complex control problem. As for dc machine, torque control in ac machine is achieved by controlling the motor currents. However, in contrast to a dc machine, in an ac machine, both the phase angle and the modulus of the current has to be controlled, or in other words, the current vector has to be controlled. This is the reason for the terminology ‘vector control’. Furthermore, in dc machines, the orientation of the field flux and the armature mmf is fixed by the commutator and the brushes, while in ac machines the field flux and the spatial angle of the armature mmf require external control. In the absence of this control, the spatial angle between the various fields in ac machines vary with the load and yield unwanted oscillating dynamic response. With vector control of ac machines, the torque and flux producing current components are decoupled and the transient response characteristics are similar to those of a separately excited dc machine, and the system will adapt to any load disturbance or reference value variations as fast as a dc machine.

Following the early works of Blaschke and Hasse and largely due to the pioneering work of Professor Leonhard, vector control of ac machines has become a powerful and frequently adopted technique worldwide. In recants years, numerous important contributions have been made in this field by contributors from many countries, including Canada, Germany, Italy, Japan, the UK, and the USA. Many industrial companies have marketed various forms of induction motor and synchronous motor drives using vector control.

1.2 Motivation of work

The field of ac drives has experienced an explosive growth in recent years and its demand reaches at the peak as additional services are being added to existing infrastructure. The control and estimation of induction motor drives constitute a vast subject, and the technology has further advanced in recent years. Induction motor drives with cage type machines have been the workhorses in industry for variable speed application in a wide power range that covers from fractional horsepower to multimegawatts. These applications includes pumps and fans, paper and textile mills, subway and locomotive propulsion, electric and hybrid vehicles, machine tools and robotics, home appliances, heat pumps and air conditioners, rolling mills, wind generation system, etc. In addition to process control, the energy saving aspects of variable frequency drives is getting a lot of attention nowadays.

In many books and survey papers the authors mentioned that the research in the area of Variable Structure Systems and Sliding Mode Control (SMC) was initiated in the former Soviet Union about 40 years ago and then the developed control methodology has been receiving much more attention of the international control community within the last two decades. So first, the events happened before these “last two decades” starting from the late fifties are surveyed to demonstrate the initial ideas and hopes of the first researchers. The retrospective view will be helpful to establish the bridges between those early steps and the scientific arsenal accumulated in the sliding mode control design methodology by now.

1.3 LITERATURE OVERVIEW

The vector control technique has been widely employed in several electric drive applications . By providing decoupling of torque and flux control commands, the vector control can navigate an AC motor drive similar to a separately excited DC motor drive without sacrificing the quality of the dynamic performance. Within this scheme, a rotational transducer such as a tachogenerator, an encoder or a resolver, was often mounted to establish the speed feedback. In this manner, the speed information can be obtained. The close-loop control paradigm is also reached. However, this speed sensor may lower the system reliability, increase device investment and complicate the implementation. Therefore, a speed sensor less drive has been used in modern industrial application.

The method of vector control allows high-performance control of torque, speed, or position to be achieved from an induction machine. This method can provide at least the same performance from an inverter-driven induction machine as is obtainable from a separately excited dc machine. Vector control provides decoupled control of the rotor flux magnitude and the torque-producing current, with a fast, near-step change in torque achievable. The fast torque response is achieved by estimating, measuring, or calculating the magnitude and position of the rotor flux in the machine. In the indirect method of vector control presented here, the calculation of the rotor flux position is dependent on the rotor resistance value. A parameter identification algorithm, such as the extended Kalman filter (EKF), can be used for the estimation of rotor resistance.

The evaluation of a vector-controlled induction machine a system can be performed in two ways. The first approach uses a real machine with an appropriate vector controller and a current-regulated pulse width modulation (PWM) inverter. The vector controller is usually implemented on a microprocessor or microcontroller and the algorithm coded in assembler or “C” if a compiler is available. A digital signal processor (DSP) is required if rotor resistance estimation is also to be implemented. The second approach uses a simulation of the system. It is therefore necessary to be able to model the components of the system.

Advantages of the first approach are that it naturally includes factors such as the actual noise present, the PWM waveform and nonlinear device, and sensor characteristics that are difficult or impossible to include in a simulation. Advantages of using a simulation environment are that all quantities can be readily observed and parameters altered to investigate their effect and to help debug estimation and control routines. Sensor noise can be added to simulate the performance of real sensors in addition to testing the ability of parameter identification techniques to cope with noisy measurements. Process noise can be added to include model imperfections in the machine model. If tests on more powerful or higher rated speed machines are required, this can be readily implemented by changing only the parameters of the machine model. The ability to run the simulation slowly also aids diagnostics.

This drive application allows vector control of the AC induction motor running in a closed-speed loop with the speed / position sensor coupled to the shaft. The application serves as an example of AC induction vector control drive design using a free scale hybrid controller with PE support. It also illustrates the use of dedicated motor control libraries that are included in PE. This application note includes a description of free scale hybrid controller features, basic AC induction motor theory, the system design concept, hardware implementation and software design, including the PC master software visualization tool.

Induction motor drives are becoming the standard for variable speed applications because of the robustness of squirrel cage induction motor, their stiff torque speed characteristics and the development of high performance control algorithm and power electronics. The rotor speed is used for flux estimation in high performance drives and for speed feedback when open loop control is sufficient. However rotor speed measurement is always troublesome, very often not practical and sometimes impossible.

For these reasons there is a growing interest in estimating the rotor speed by using only the stator voltage and current measurements. This estimator could be useful, not only for speed feedback as part of speed control loop in servo drives, but also for diagnosis of the drive when speed measurement is available. Since all the necessary electrical variables are available with in the power electronics controller, speed estimates and measurements could be compared. Note that estimation algorithms require calculation of the variable of interest, plus some averaging method to filter noise.

Indirect field-oriented control (vector control) of induction motors has been widely used in high-performance ac drives. To orient the injected stator current vector and to establish speed loop feedback, knowledge of rotor speed is necessary in these applications. Tachogenerators or digital shaft-position encoders are usually used to detect the rotor speed of motors. These speed sensors lower the system reliability and require special attention to noise. In addition, for some special applications such as very high-speed motor drives, there are difficulties in mounting these speed sensors. In recent years, several tacholess vector control schemes have been proposed. The traditional approach to tacholess vector control uses the method of field orientation of rotor flux, which has difficulty in determining the instantaneous orientation of the flux vector. Other approaches are based on the model-reference adaptive system (MRAS), These MRAS schemes require pure integrators in their reference models, and are affected by stator resistance thermal variations. These problems have limited the applications of tacholess vector control in high-performance and low-speed drives. Consequently, tacholess vector control has an image of providing a low class of ac drives.

In modern adjustable-speed alternating current (ac) drives an accurate and fast speed sensing is highly desirable for the optimum operation of the drive. In most drives, an optoelectrical or electromechanical speed transducer is used. The conventional transducer suffers from at least one of the various problems such as nonlinearity, drift, low resolution and accuracy, and poor performance at low speeds. In addition to these problems and the initial high cost, the transducer requires mounting, wiring, calibration, and maintenance. These shortcomings spoil the praised characteristics of robustness and mechanical simplicity of an ac drive.

During the last two decades, attempts have been made to replace the conventional speed transducers in adjustable- speed drives, by sensing the speed from the electrical quantities applied to the motor, i.e., voltage and current which are already available for the drive assessment and control. An early effort was made by Abbondanti and Brennen who designed in 1975 an analog slip calculator based on the processing of the motor input quantities, voltages, currents, and phases [l]. In 1979, it was followed by the work of Ishida *et al.* [2], who used rotor slot harmonic voltages in slip frequency control. They reported success for speeds over 300 r/min. Ishida and Iwata endorsed this technique in [3], [4], without any further progress in improving the speed range.

Hammerli *et al. [5]* presented a method, in 1987, based on detecting the speed in the range 20-100% of nominal speed from the rotor slot harmonics and, in the range under 30% of nominal speed, by injecting an additional signal of constant frequency (from the inverter) into the machine to produce rotor slot modulation and therefore enhance the speed detection at low frequencies. A different approach was reported by Beck and Naunin in which they described a sensor less speed control of a squirrel cage induction motor, based on the calculation of the rotor frequency from the phase angle between the stator voltage and current [6].

In 1990, Williams *etal.* [7] Used rotor slot harmonics in the stator current reconstructed from the direct current (dc) link current to detect the speed of an inverter-fed induction motor using switched-capacitor filters. In the published paper no practical assessment was shown to evaluate the detector with respect to static and dynamic performance. Recently, sensor less field-oriented control systems were introduced. In these drives the speed is estimated digitally from terminal voltages and currents. However, the techniques used rely heavily on the estimated motor parameters which change considerably with frequency and temperature .

Modern control techniques for ac motor drives were developed largely as a result of the search for low-cost alternatives to high-performance four-quadrant dc servo drives. In these applications, the use of shaft mounted tacho generators and resolvers are established practice, and the digital shaft-position encoder used in the most effective vector-control schemes is considered acceptable. Nevertheless, the shaft encoder does present problems. Delicate optical encoders with internal signal conditioning electronics are widely used. These lower the system reliability, especially in hostile environments, and require careful cabling arrangements with special attention to electrical noise. There are also situations where the positional feedback is extremely difficult to obtain. This is particularly true for the case of linear-motor drives such as for transportation vehicles. Finally, the encoder is a cost factor since the provision of special motor-shaft extensions and encoder-mounting surfaces leads to more expensive machines.

` Until recently, the rapid developments in vector control technology have had little impact on “adjustable-speed” ac drives. These are typically simple voltage-source inverters with variable output frequency and are used for applications requiring little dynamic control, such as pumps and fans. It has now become clear that these drives can benefit from the closed-loop current-control techniques that have evolved for use in vector control systems. Current control is readily applicable to existing voltage-source inverters, where it reduces the incidence of over current tripping and improves inverter utilization.

Once the inverter is current controlled, additional controls must be provided to specify the magnitude and slip frequency of the injected current vector and hence regulate the flux and torque of the motor. Motor speed feedback is typically required for outer-loop speed control as well as in the flux and torque control algorithm. This presents a problem in low-performance systems where motor speed transducers are not usually available. This has led to a renewed interest in “tacholess” vector control with the objective of providing an intermediate class of ac drives with enhanced performance and a wider range of applications than “adjustable-speed” drives but about the same cost except for the small additional cost of more sophisticated control algorithms. . The difficulty in either case lies in determining the instantaneous orientation of the relevant vector. Tamai et al. have described an approach in which the field-oriented reference frame is identified by means of a model-reference adaptive system (MRAS). This paper presents an alternative tacholess control method that uses an MRAS to determine the motor speed and thereby establishes vector control of the motor as well as overall speed control. The new MRAS scheme is thought to be less complex and more effective than the previous approach. It has been implemented in a 30-hp drive, which has proved its viability and robust nature.

1.4 SPEED DETERMINATION USING ROTOR SLOT HARMONICS:

For a given pole pitch and supply frequency f_0 the speed of the traveling (rotating) field is constant and is known as the synchronous speed. It is given in revolutions per minute (r/min) by the well-known equation

$$n_s = \frac{60 f_0}{p} \tag{1.1}$$

The difference between the actual mechanical speed n of the rotor and the synchronous speed n_s is known as slip. This difference in speeds depends upon the load on the motor and is often expressed as a fraction of the synchronous speed, as follows

$$s = \frac{n_s - n}{n_s} \tag{1.2}$$

The relative speed n_r given by

$$n_r = n_s - n \tag{1.3}$$

And, known as slip speed. If at any slip speed, the frequency of the rotor current is f_r then f_r will be expressed as

$$f_r = s f_0 \tag{1.4}$$

And it is referred to as slip frequency. Since the rotor current is proportional to the slip frequency, to restrict the current drawn by the stator from exceeding its rated value, the steady-state slip frequency in commonly used induction motor drives is always kept small. In this work, in order to define a detection range for slot harmonics, the maximum slip is assumed equal to 5%, which is usually the case for most drives over 10 kW. Combining the two last equations to rotor rotational frequency f_0 will be given by

$$f_{r0} = (1 - s) f_0 \tag{1.5}$$

$$n = \frac{60}{z} (f_{sh} \pm f_0) \tag{1.6}$$

Where, f_{sh} is the slot harmonic frequency. If Z is a number other than a multiple of three, only one slot harmonic component is detected [3], which is normally the case for all squirrel-cage motors.

1.5 PARAMETER SENSITIVITY EFFECT:

The indirect vector controller is dependent on the rotor resistance R_r the mutual inductance L_m and the self inductance of the rotor L_r. The inductances change with the saturation of the magnetic material while the rotor resistance is affected by the variation in the temperature. The increase in temperature increases the rotor resistance. In case of the current source inverters used in high performance drive systems, if the stator current is maintained constant this has the effect of increasing the magnetizing current and hence saturating the machine. Core losses in the motor increase with the saturation of the magnetic path and in the end necessitates a considerable derating of the machine. The effects of this mismatch between the controller and motor parameters both for steady-state and dynamic conditions are summarized.

1.5.1 STEADY STATE EFFECTS:

In a torque controlled induction motor drive, i.e., with the outer speed loop open, the rotor flux linkages and the electromagnetic torque are the external command signals while the actual slip speed is maintained equal to the command value. This mismatch between the controller and the motor parameters introduces deviations in the flux and torque producing components of the stator current and hence the torque angle. The parameter α is the temperature factor and defined as the ratio between the rotor time constant of the motor and its instrumented value in the vector controller. The parameter p is the saturation factor and defined as the ratio between the actual mutual inductance and the instrumented value in the vector controller. For α < I which signifies an increase in the rotor temperature, the electromagnetic torque is greater than the commanded value. An increase in α indicates an increase in motor saturation at ambient temperature and for such operating conditions the torque is less than the commanded value.

This results in increased stator copper losses which has a detrimental effect on the thermal rating of the machine. α can be greater than l due to any of the following factors:

- Field weakening.
- Ambient less than the assumed values.
- Wrong instrumentation of the controller parameters.

Under this circumstance, the torque command has to be higher than for that required with α = 1, and hence the stator current and slip speed will be higher than the nominal values. This would result in higher copper losses in the machine having a detrimental effect on the rating of the motor drive.

1.5.2 DYNAMIC EFFECTS:

In the dynamic operation of the vector controlled induction motor drive it can be shown that the flux and torque have a time constant equal to the rotor time constant and a natural frequency of oscillation with a value equal to the slip frequency. In torque controlled drives, the developed rotor flux and torque increase with rising temperature but the oscillatory responses are undesirable and make the induction motor drive a non-ideal torque amplifier. In a speed controlled drive, the oscillations in the torque are not transmitted to the rotor shaft due to the following reasons.

- The high bandwidth of the outer speed loop forces the torque command to match the load torque in very short times.
- The moment of inertia of the induction motor drive and the load introduces filtering.

1.6 MRAS SPEED ESTIMATION:

The speed estimate $\hat{\omega}_r$ can be made to track the actual speed very closely and can be used both for speed loop feedback and for orienting the injected stator current vector for torque and flux control. An important feature of this system is that, provided the same value of T, is used in the MRAS adjustable model, perfect orientation of the injected current vector is achieved, in theory, even if the value of T2 used is quite wrong. If the MRAS successfully maintains nearly zero error, then the adjustable model accurately replicates the dynamic relationship between the stator current vector and the rotor flux vector that exists in the actual motor. Examination shows that this is only possible if

$$T_2(\omega - \hat{\omega}_r) = T_{2actual}(\omega - \hat{\omega}_r) \quad \textbf{(1.7)}$$

$$T_2\omega_{sl} = T_{2actual}\omega_{slactual} \quad \textbf{(1.8)}$$

Thus, if $T_2 \neq T_{2actual}$ then $\omega_{sl} \neq \omega_{slactual}$ the product $T_2\omega_{sl}$ is used in setting up the injected current vector, this is always done correctly for the actual slip, even under dynamic conditions. Naturally, the error in the value of T_2 will reflect as an error in the speed feedback, which will affect the accuracy of the speed control slightly.

1.7 VARIABLE STRUCTURE SYSTEM:

A high level of scientific and publication activity, an unremitting interest in variable structure control enhanced by effective applications to automation problems most diverse in their physical nature and functional purposes are a cogent argument to consider this class of nonlinear systems as a prospective area for study and applications. The term "variable structure system" (VSS) first made its appearance in the late 1950's. Since that time, the first expectations of such systems have naturally been reevaluated, their real potential has been revealed, new research directions have been originated due to the appearance of new classes of control problems, new mathematical methods, recent advances in switching circuitry, and (as a consequence) new control principles. The paper is oriented to base-stone ideas of VSS design methods and selected set of applications rather than the survey information or a historical sequence of the events accompanying. Furthermore, it will be shown that the dominant role in VSS theory is played by sliding modes, and the core idea of designing VSS control algorithms consists of enforcing this type of motion in some manifolds in system state spaces. Implementation of sliding mode control implies high-frequency switching. It does not cause any difficulties when electric drives are controlled since the "on-off" operation mode is the only admissible one for power converters. This reason predetermined both the high efficiency of sliding mode control for electric drives and the author choice of the application selection topic in this paper.

1.8 SPEED ESTIMATION USING VOLTAGE AND CURRENT:

Three main type of algorithm have been proposed related to this paper, based on the ideal model of the induction motor.

- Extended estimator for joint state and rotor speed estimation(considering the latter as a parameter) (Kim et al,1994; Do et al., 1995)
- Linear regression models. (Velez Reyes, 1992) and
- Model reference adaptive systems (Schauder, 1992; Peng and Fukao, 1994)

Algorithm of the first type require a high sampling frequency so that a simple model discretization algorithm can be used, and are computationally intensive, anyway algorithm of type two are very simple and easy to implement in real time. In addition they are well known and have been used successfully in many applications, calculation of the integrals and derivative of signals is required. However it will be shown that a state variable filter can be used for this purpose. Algorithm of type three has been widely applied in this field. Their main drawbacks are their sensitivity to inaccuracy in the reference model, and the difficulties of designing the adaptation block.

In this paper algorithm two is utilized. The proposed algorithm estimates rotor speed together with slip frequency and rotor flux frequency. This method is similar to that proposed by Vas (1993), but includes modification that ease the implementation of a state variable filters to obtained the derivative and integrals involved during calculation.

The paper intended not only to describe in detail an alternative for rotor speed estimation, but also to give a good inside into the features and limitations of the motor model used which could be applicable to other estimation algorithms described.

Typically, the rotor speed varies slowly, and can be considered as a parameter. In this can the motor admittance can be written as the frequency response transfer function between stator current and voltage

$$Y_s(\omega, \omega_R) = \frac{i_s(\omega, \omega_R)}{v_s(\omega, \omega_R)} \qquad \textbf{(1.9)}$$

The identifiably of rotor speed by using stator currents and voltages has to be investigated by considering the sensitivity of motor admittance to rotor speed. The motor admittance is a complex function, with real and imaginary parts for each harmonic in supply voltage; therefore, the response of the induction motor for each frequency component could be used to calculate the parameters. However the asymptotic value of the motor admittance with increasing harmonic order is showing that the leakage inductance σL_s is the only motor parameter in this limit, where as the rotor speed is absent. Therefore, only fundamental and low order harmonics can be used to calculate the rotor speed.

$$Y_{s\lim} = \lim_{\omega \to \infty} Y_s = \frac{1}{j\sigma L_s \omega} \quad \textbf{(1.10)}$$

1.9 SLIDING MODES IN VSS:

Variable structure systems consist of a set of continuous subsystems with a proper switching logic and, as a result, control actions are discontinuous functions of system state, disturbances (if they are accessible for measurement), and reference inputs. In the course of the entire history of control theory, intensity of discontinuous control systems investigation has been maintained at a high enough level. In particular, at the first stage, on-off or bang-bang regulators are ranked highly due to ease of implementation and efficiency of control hardware. Furthermore, we shall deal with the variable structure systems governed by

$$\dot{x} = f(x,t,u) if x \in R^n, u \in R^m \quad \textbf{(1.11)}$$

$$u = \begin{cases} u^+(x,t) if s(x) > 0 \\ u^-(x,t) if s(x) < 0 \end{cases} \quad \textbf{(1.12)}$$

$$s(x)^T = (s_1(x)............, s_m(x)) \quad \textbf{(1.13)}$$

` The VSS (1.11) with continuous functions *f*, **s,** u^+, u^- consists of 2^m subsystems and its structure varies on m surfaces at the state space. From the point of view of our later treatment, it is worth quoting the elementary example of a second-order system with bang-bang control and sliding mode:

$$\ddot{x} + a_2 \dot{x} + a_1 x = u \quad \textbf{(1.14)}$$

$$u = -M sign s \quad \textbf{(1.15)}$$

$$s = cx + \dot{x}, M, c, a_1, a_2................const \quad \textbf{(1.16)}$$

. It follows from analysis of the (X, x) state plane that, in the neighborhood of segment *on* the switching line s = 0, the trajectories run in opposite directions, which leads to the appearance of a sliding mode along this line. The switching line equation S=0 may be treated as a motion with solutions depending only on the slope gain *and* invariant to plant parameters and disturbance.

Indirect field-oriented techniques microprocessors are now widely used for the control of induction motor servo drive in high-performance applications. With the field-oriented techniques [9,13,22], the decoupling of torque and flux control commands of the induction motor is guaranteed, and the induction motor can be controlled linearly as a separated excited dc motor. However, the control performance of the resulting linear system is still influenced by the uncertainties, which are usually composed of unpredictable parameter variations, external load disturbances, unmodeled and nonlinear dynamics. Therefore, many studies have been made on the motor drives in order to preserve the performance under these parameter variations and external load disturbances, such as nonlinear control, optimal control, variable structure system control, adaptive control and neural control [14–16].

In the past decade, the variable structure control strategy using the sliding-mode has been focused on many studies and research for the control of the ac servo drive systems [8, 10, 17, and 20]. The sliding-mode control can offer many good properties, such as good performance against unmodelled dynamics, insensitivity to parameter variations, external disturbance rejection and fast dynamic response [21]. These advantages of the sliding-mode control may be employed in the position and speed control of an ac servo system.

On the other hand, in indirect field-oriented control of induction motors, knowledge of rotor speed is required in order to orient the injected stator current vector and to establish speed loop feedback control. Tachogenerators or digital shaft-position encoders are usually used to detect the rotor speed of motors. These speed sensors lower the system reliability and require special attention to noise. In addition, for some special applications such as very high-speed motor drives, there exist difficulties in mounting these speed sensors.

Recently, much research has been carried on the design of speed sensor less control schemes [11, 12, 20, 18, and 23]. In these schemes the speed is obtained based on the measurement of stator voltages and currents. However, the estimation is usually complex and heavily dependent on machine parameters. Therefore, although sensor less vector-controlled drives are commercially available at this time, the parameter uncertainties impose a challenge in the control performance.

1.10 THESIS OUTLINE:

Apart from the introduction chapter 1 contains the literature survey of the thesis work followed by by the parameter sensitivity, steady state and dynamic affect and also MRAS speed estimation. Control. Chapter 2 contains the modeling of the induction motor such as scalar control, vector control and its principle, dynamic d-q model of the machine and the dc drive analogy to the induction machine control Chapter 3 contains the sliding mode control and its principle and the theory behind it. In this section we wish to estimate the speed with out any sensor so called sensorless control. Chapter 4 explains the parameters value taken following by results and discussions, and finally the achievements, future work, research and conclusion is discussed.

CHAPTER 2

MODELLING AND FIELD ORIENTED CONTROL OF INDUCTION MOTOR

2.1 INTRODUCTION:

The Induction motors (IM) for many years have been regarded as the workhorse in industry. Recently, the induction motors were evolved from being a constant speed motors to a variable speed. In addition, the most famous method for controlling induction motor is by varying the stator voltage or frequency. To use this method, the ratio of the motor voltage and frequency should be approximately constant. With the invention of Field Orientated Control, the complex induction motor can be modeled as a DC motor by performing simple transformations.

In a similar manner to a dc machine in induction motor the armature winding is also on the rotor, while the field is generated by currents in the stator winding. However the rotor current is not directly derived from an external source but results from the emf induced in the winding as a result of the relative motion of the rotor conductors with respect to the stator field. In other words, the stator current is the source of both the magnetic field and armature current. In the most commonly used, squirrel cage motor, only the stator current can directly be controlled, since the rotor winding is not accessible. Optimal torque production condition are not inherent due to the absence of a fixed physical disposition between the stator and rotor fields, and the torque equation is non linear. In effect, independent and efficient control of the field and torque is not as simple and straightforward as in the dc motor.

The concept of the steady state torque control of an induction motor is extended to transient states of operation in the high performance, vector control ac drive system based on the field operation principle (FOP). The FOP defines condition for decoupling the field control from the torque control. A field oriented induction motor emulates a separately exited dc motor in two aspects:

- Both the magnetic field and torque developed in the motor can be controlled independently.
- Optimal condition for the torque production, resulting in the maximum torque per unit ampere, occurs in the motor both in steady state and in transient condition of operation.

2.2 INDUCTION MACHINE CONTROL:

Squirrel cage induction machines are simple and rugged and are considered to be the workhorses of industry. At present induction motor drives dominate the world market. However, the control structure of an induction motor is complicated since the stator field is revolving, and further complications arises due to the fact that the rotor currents or rotor flux of a squirrel cage induction motor can not be directly monitored

The mechanism of torque production in an ac machine and in a dc machine is similar. Unfortunately this similarity was not emphasized before the 1970s, and this is one of the reasons why the technique of vector control did not emerge earlier. The formulae given in many well known textbook of the machine theory have also implied that, for the monitoring of the instantaneous electromagnetic torque of an induction machine, it is also necessary to monitor the rotor currents and the rotor position. Even in the 1980s some publications seemed to strengthen this false conception, which only arose because the complicated formulae derived for the expression of the instantaneous electromagnetic torque have not been simplified. However by using fundamental physical laws or space vector theory, it is easy to show that, similar to the expression of the electromagnetic torque of a separately exited dc machine, the instantaneous electromagnetic torque of an induction motor can be expressed as the product of a flux producing current and a torque producing current, if a special flux oriented reference is used.

2.2.1 Scalar control:

Scalar control, as the name indicates, is due to magnitude variation of the control variables only, and disregarding the coupling effect in the machine. For example, the voltage of the machine can be controlled to control the flux, and the frequency and slip can be controlled to control the torque. However, flux and torque are also functions of frequency and voltage, respectively. Scalar control in contrast to the vector control or field oriented control, where both the magnitude and phase is controlled. Scalar controlled drives give some what inferior performance, but they are easily implemented. Scalar controlled drives have been widely used in industry. How ever their importance has diminished recently because of the superior performance of vector controlled drives, which is demanded in many applications.

2.2.2 Vector or field oriented control:

Scalar control is somewhat simple to implement, but the inherent coupling effect i.e. both torque and flux are functions of voltage or current and frequency gives the sluggish response and the system is easily prone to instability because of high order system harmonics. For example, if the torque is increased by incrementing the slip or slip the flux tends to decrease. The flux variation is sluggish. The flux variation then compensated by the sluggish flux control loop feeding additional voltage. This temporary dipping of flux reduces the torque sensitivity with slip and lengthens the system response time.

These foregoing problems can be solved by vector control or field oriented control. The invention of vector control in the beginning of 1970s, and the demonstration that an induction motor can be controlled like a separately excited dc motor, brought a renaissance in the high performance control of ac drives. Because of dc machine like performance, vector control is known as decoupling, orthogonal, or transvector control. Vector control is applicable to both induction and synchronous motor drives. Vector control and the corresponding feedback signal processing, particularly for modern sensor less vector control, are complex and the use of powerful microcomputer or DSP is mandatory. It appears that eventually, vector control will oust scalar control, and will be accepted as the industry standard control for ac drives.

2.3 DC DRIVE ANALOGY:

Ideally, a vector controlled induction motor drive operates like a separately excited dc motor drive in fig 1.1. In a dc machine, neglecting the armature reaction effect and field saturation, the developed torque is given by

$$T_e = K_t \Psi_f \Psi_a = K_t^{'} I_a I_f \qquad \textbf{(2.1)}$$

Where I_a = armature current

And I_f = field current

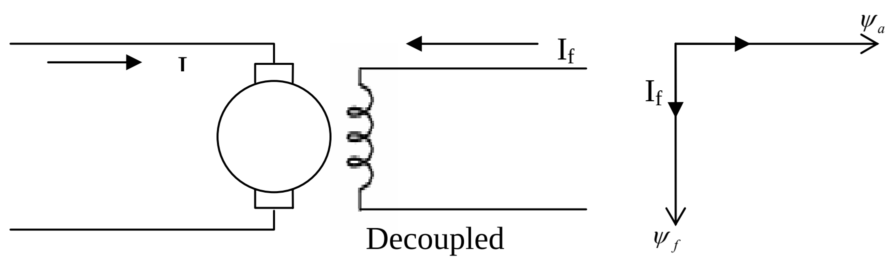

Fig 2.1-separately excited dc motor

This construction of dc machine is such that the field flux Ψ_f produced by the current I_f is perpendicular to the armature flux Ψ_a, which is produced by the armature current. These space vectors, which are stationary in space, are orthogonal and decoupled in nature. This means that when torque is controlled by controlling the current I_a, the flux Ψ_f is not affected and we get the fast transient response and high torque ampere ratio. Because of decoupling, when the field current I_f is controlled, it affects the field flux Ψ_f only, but not the Ψ_a flux. Because of the inherent coupling problem, an induction motor can not generally give such fast response.

DC machine like performance can also be extended to an induction motor if the machine is considered in a synchronously rotating reference frame ($d^e - q^e$), where the sinusoidal variables appears as dc quantity in steady.

In fig 1.2, the induction motor with the inverter and vector control in the front end is shown with two control current inputs i_{ds}^* and i_{qs}^* .These currents are the direct axis component and quadrature axis component of the stator current, respectively, in a synchronously rotating reference frame.

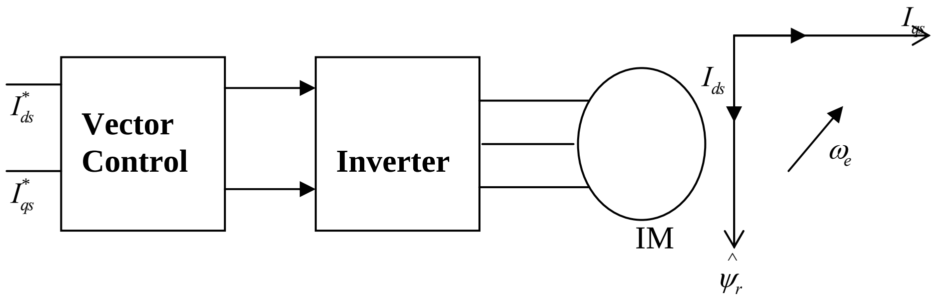

Fig 2.2 vector controlled induction motor

With vector control, i_{ds} is analogous to field current I_f and i_{qs} is analogous to armature current I_a of a dc machine. Therefore, the torque can be expressed as

$$T_e = K_t \hat{\Psi}_r I_{qs} = K_t' I_{qs} I_{ds} \qquad \textbf{(2.2)}$$

Where $\hat{\Psi}_r$ = absolute peak value of the sinusoidal space vector. This dc machine like performance is only possible if i_{ds} is oriented in the direction of $\hat{\Psi}_r$ and i_{qs} is established perpendicular to it, as shown by the space vector diagram of fig 1.2. This means that when i_{qs}^* is controlled; it affects the actual i_{qs} current only, but does not affect the flux $\hat{\Psi}_r$. Similarly, when i_{ds}^* is controlled, it controls the flux only and does not affect the i_{qs} component of current. This vector or field orientation of current is essential under all operating condition in a vector control drive. When compared to dc machine space vectors, induction machine space vector rotate synchronously at frequency ω_e as indicated in fig 1.2.

2.4 PRINCIPLE OF VECTOR CONTROL:

The fundamentals of vector control implementation can be explained with the help of fig 1.3. Where the machine model is represented in a synchronously rotating reference frame. The inverter is omitted from the figure, assuming that it has unity current gain, that is, it generates currents $i_a i_b and i_c$ as dictated by the corresponding commands currents i_a^*, i_b^* and i_c^* from the controller. A machine model with internal conversions is shown on the right. The machine terminal phase currents $i_a i_b and i_c$ are converted to I_{ds}^s and I_{qs}^s component by $\frac{3\varphi}{2\varphi}$ transformation.

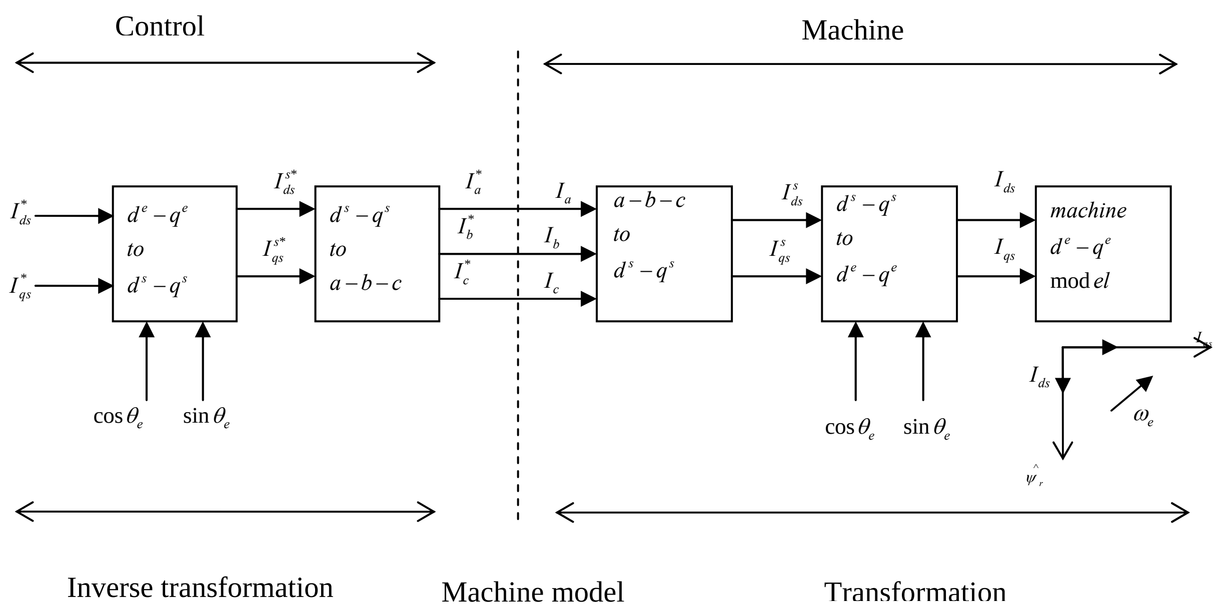

Fig 2.3 vector control implementation principle with machine ($d^e - q^e$) model.

These are then converted to synchronously rotating frame by the unit vector components $\cos\theta_e$ and $\sin\theta_e$ before applying them to the $d^e - q^e$ machine model as shown in the fig 1.3. The controller makes two stage of inverse transformation, as shown, so that the control currents $i_{ds}{}^*$ and $i_{qs}{}^*$ corresponds to the machine current i_{ds} and i_{qs}, respectively. The transformation and inverse transformation including the inverter ideally do not incorporate any dynamics therefore, the response to i_{ds} and i_{qs} is instantaneous.

2.5 DYNAMIC d-q MODEL:

The per phase equivalent circuit of the machine, which is only valid in steady state condition. In an adjustable speed drive, the machine normally constitute an element with in a feedback loop, and therefore its transient behavior has to be taken into consideration. Besides, high performance drive control, such as vector or field oriented control is based on the dynamic d-q model of the machine.

Basically, it can be looked on as a transformer with a moving secondary, where the coupling coefficients between stator and rotor phases changes continuously with the change of rotor position θ_r. The machine model can be described by the differential equations with timing varying mutual inductances, but such a model tends to be varying complex. A three phase machine can be represented by an equivalent two phase's machine, where $d^s - q^s$ correspond to stator direct and quadrature axes, and $d^r - q^r$ correspond to rotor direct and quadrature axes. Although it is somewhat simple, the problem of time varying parameters still remain. R. H Park in the 1920s, proposed a new theory of electric machine analysis to solve this problem. He formulated a change of variables which, in effect, replaced the variables voltage, current and flux linkage associated with stator windings of a synchronous machine with variables associated with fictitious windings rotating with the rotor at synchronous speed he referred, the stator variables to a synchronously rotating reference frame fixed in the rotor. With such transformation called Park's transformation, showed that all the time varying inductances that occur due to an electric circuit in relative motion and electric circuit with varying magnetic reluctance is eliminated.

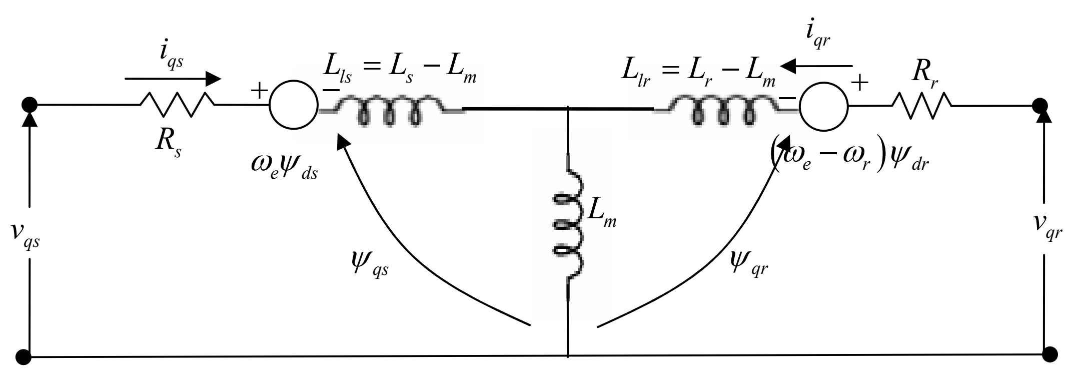

Fig 2.4 dynamic $d^e - q^e$ equivalent circuit of machine (q^e axis)

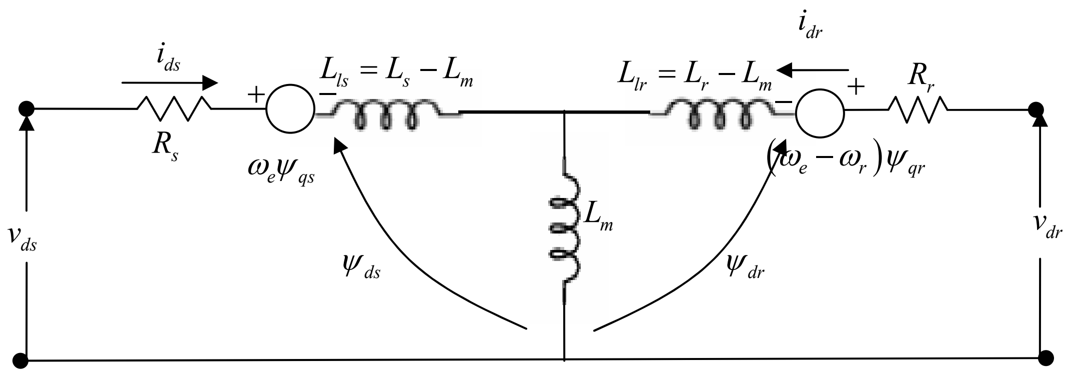

Fig 2.5 dynamic $d^e - q^e$ equivalent circuit of machine (d^e axis)

The above figure shows the $d^e - q^e$ dynamic model equivalent circuit. A special advantage of the $d^e - q^e$ dynamic model of the machine is that all the sinusoidal variables in stationary frame appear as dc quantity in synchronous frame. From the above model the electrical transient model in terms of voltage and currents can be given in matrix form,

$$\begin{bmatrix} v_{qs} \\ v_{ds} \\ v_{qr} \\ v_{dr} \end{bmatrix} = \begin{bmatrix} R_s + SL_s & \omega_e L_s & SL_m & \omega_e L_m \\ -\omega_e L_s & R_s + SL_s & -\omega_e L_m & SL_m \\ SL_m & (\omega_e - \omega_r) L_m & R_r + SL_r & (\omega_e - \omega_r) L_r \\ -(\omega_e - \omega_r) L_m & SL_m & -(\omega_e - \omega_r) L_r & R_r + SL_r \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{bmatrix} \quad \textbf{2.3}$$

There are basically two different types of vector control technique: direct and indirect techniques. The direct implementation relies on the direct measurement or estimation of rotor-, stator-, or magnetizing-flux-linkage vector amplitude and position. The indirect method uses a machine model, e.g. for rotor flux oriented control, it utilizes the inherent slip relation. In contrast to direct methods, the indirect methods are highly dependent on machine parameters. Traditional direct vector control schemes use search coils, tapped stator windings or Hall Effect sensors for flux sensing. This introduces limitation due to machine structural and thermal requirements. Many applications use indirect schemes, since these have relatively simpler hardware and better overall performance at low frequencies, but since these contain various machine parameters, which may vary with temperature, saturation level and frequency, various parameter adaptation schemes, have been developed. These include self tuning control application, model reference adaptive system (MRAS) applications, applications of observers, applications of intelligent controllers etc. to obtain a solution, in the development of theory it is sometimes assumed that the mechanical time constant is much greater than the electrical time constants, but this becomes an invalid assumption if the machine inertia is low. If incorrect modulus and the angle of the flux linkage space vector are used in a vector control scheme, then flux and torque decoupling is lost and the transient and steady state responses are degraded. Low frequency response, speed oscillations, and loss of input output torque linearity is major consequences of detuned operation, together with decreased drive efficiency.

2.6 Dq-abc TRANSFORMATION:

The relation between the synchronously rotating reference frame and the stationary reference frame is performed by the so-called reverse Park’s transformation:

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & -\sin(\theta_e) \\ \cos(\theta_e - 2\pi/3) & -\sin(\theta_e - 2\pi/3) \\ \cos(\theta_e + 2\pi/3) & -\sin(\theta_e - 2\pi/3) \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \tag{2.4}$$

Where θe is the angle position between the *d*-axis of the synchronously rotating reference frame and the *a*-axis of the stationary reference frame and it is assumed that the quantities are balanced.

This state transformation offers certain advantages, among which, the fact that, in this new reference frame, the electromagnetic torque is directly an image of the quadrature ("q”) component of the stator current.

2.7 DIRECT OR FEEDBACK VECTOR CONTROL:

The principal vector controls parameters, i_{ds}^* and i_{qs}^*, which are dc values in synchronously rotating frame, are converted to stationary frame as vector rotation (VR) with the help of unit vector generated from flux vector signals Ψ^s_{dr} and Ψ^s_{qr} . The resulting stationary frame signals are then converted into phase current commands for the inverter. The flux signal Ψ^s_{dr} and Ψ^s_{qr} .are generated from the machine terminal voltages and currents with the help of the voltage model estimator. A flux control loop has been added for precision control of flux. The torque component of current i_{qs}^* is generated from the speed control loop through a bipolar limiter. The torque proportional to i_{qs} ,can be bipolar.

The correct alignment of current i_{ds} in the direction of the flux $\hat{\Psi}_r$ and the current i_{qs} perpendicular to it are crucial in vector control. This alignment with the help of the stationary frame rotor flux vectors Ψ^s_{dr} and Ψ^s_{qr} is explained in fig 1.4. In this figure, the $d^e - q^e$ frame is rotating with a synchronous speed ω_e with respect to stationary frame $d^s - q^s$, and at any instant, the angular position of the $d^e - q^e$ to $d^s - q^s$ axis is θ_e , where $\theta_e = \omega_e t$.

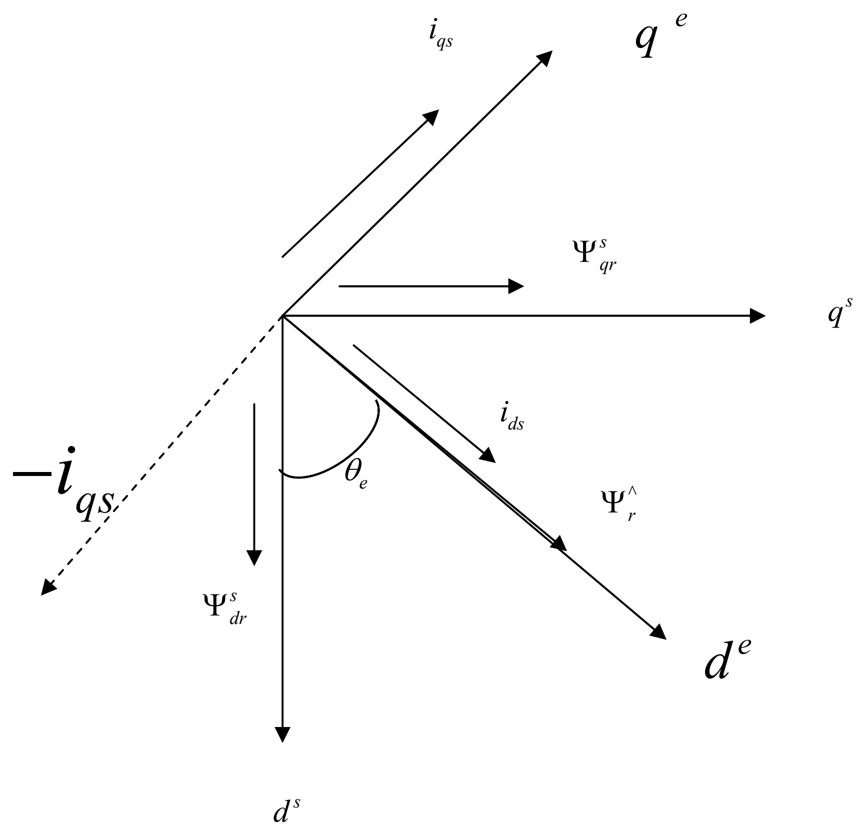

Fig 2.6 $d^s - q^s$ **and** $d^e - q^e$ **phasors with correct rotor flux orientation**

From the fig 1.6 we can write the following equations:

$$\Psi_{dr}^s = \hat{\Psi}_r \cos\theta_e \tag{2.5}$$

$$\Psi_{qr}^s = \hat{\Psi}_r \sin\theta_e \tag{2.6}$$

$$\cos\theta_e = \frac{\Psi_{dr}^s}{\hat{\Psi}_r} \tag{2.7}$$

$$\sin\theta_e = \frac{\Psi_{qr}^s}{\hat{\Psi}_r} \tag{2.8}$$

$$\hat{\Psi}_r = \sqrt{\Psi_{dr}^{s^2} + \Psi_{qr}^{s^2}} \tag{2.9}$$

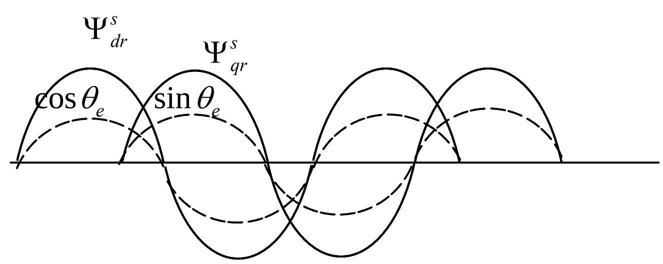

Fig 2.7 plot of unit vector signals in correct phase position

2.8 SALIENT FEATURES OF VECTOR CONTROL:

- ❖ The frequency ω_e of the drive is not directly controlled as in scalar control. The machine is essentially self controlled where the frequency as well as the phase are controlled indirectly with the help of unit vector.
- ❖ There is no fear of an instability problem by crossing the operating point beyond the breakdown torque T_{em} as in a scalar control. Limiting the total $\hat{I}_s$ with in the safe limit automatically limits operation with in the stable region.
- ❖ The transient response will be fast and dc machine like because torque control by i_{qs} does not affect the flux. However ideal vector control is not possible in practice, because of delays in converter and signal processing and the parameter variation effect.
- ❖ Like a dc machine, speed control is possible in four quadrants without any additional control elements like phase sequence reversing. In forward motoring condition, if the torque T_e in negative, the drive initially goes into regenerative braking mode, which slows down the speed. At zero speed, the phase sequence of the unit vector automatically reversus, giving reverse motoring operation.

The direct method of vector control is difficult to operate successfully at very low frequency including zero speed because of the following problems:

- At low frequency, voltage signals v_{ds}^{s} and v_{qs}^{s} are very low. In addition, ideal integration becomes difficult because dc offset trends to build up at the integrater output.
- The parameter variation effect of resistance R_s and inductance L_{ls}, L_{lr} and L_m tend to reduce accuracy of the estimated signals. Particularly temperature variation of R_s becomes more dominant. However, compensation of R_s is somewhat easier, which will be discussed later. At higher voltages, the effect of parameters can be neglected.

In industrial applications, vector drives often required operating from zero speed (including zero speed start up), hence direct vector control with voltage model signal estimation can not be used.

2.9 INDIRECT (FEED FORWARD) VECTOR CONTROL:

The indirect vector control method is essentially the same as direct vector control, except the unit vector signals are generated in feed forward manner. Indirect vector control is very popular in industrial application. The explanation of the fundamental principle of the indirect vector control with the help of phasor diagram is given by the fig 2.8.

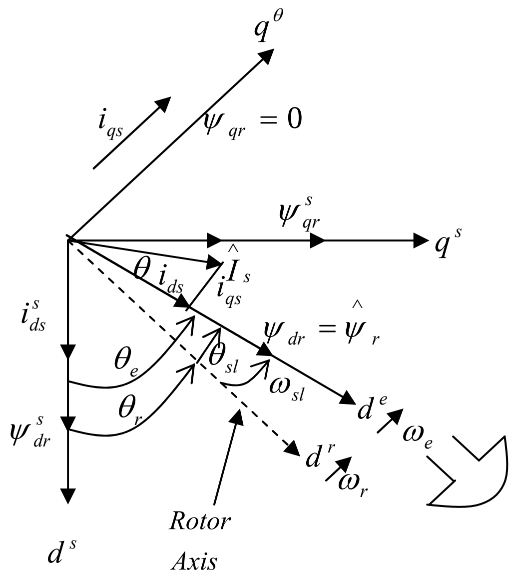

Fig 2.8 phasor diagram explaining indirect vector control

The $d^s - q^s$ axes are fixed on the stator, but the $d^r - q^r$ axes, which are fixed on the rotor, are moving at a speed ω_r as shown in the fig 1.6. Synchronously rotating axes $d^e - q^e$ are rotating ahead of the $d^r - q^r$ axes by the positive slip angle θ_{sl} corresponding to slip frequency ω_{sl}. Since the rotor pole is directed on the d^e axes and,

$$\omega_e = \omega_r + \omega_{sl} \tag{2.10}$$

We can write

$$\theta_e = \int \omega_e dt = \int (\omega_r + \omega_{sl}) dt = \theta_r + \theta_{sl} \tag{2.11}$$

The rotor pole position is not absolute, but is slipping with respect to the rotor at frequency ω_{sl}. The phasor diagram suggests that for decoupling control, the stator flux component of current i_{ds} should be aligned on the d^e axes, and the torque component of current i_{qs} should be on the q^e axes as shown in the fig 1.8.

From the voltage model of the flux vector estimation, the machine terminal voltage and currents are sensed and fluxes are computed from the stationary frame $d^s - q^s$ equivalent circuit. These fluxes equations are:

$$\Psi_{ds}^{s} = \int (v_{ds}^{s} - R_s i_{ds}^{s}) dt \qquad \textbf{(2.12)}$$

$$\Psi_{qs}^{s} = \int (v_{qs}^{s} - R_s i_{qs}^{s}) dt \qquad \textbf{(2.13)}$$

$$\Psi_{r}^{s\,2} = \Psi_{dr}^{s\,2} + \Psi_{qr}^{s\,2} \qquad \textbf{(2.14)}$$

$$\Psi_{dm}^{s} = \Psi_{ds}^{s} - L_{ls} i_{ds}^{s} = L_m (i_{ds}^{s} + i_{dr}^{s}) \qquad \textbf{(2.15)}$$

$$\Psi_{qm}^{s} = \Psi_{qs}^{s} - L_{ls} i_{qs}^{s} = L_m (i_{qs}^{s} + i_{qr}^{s}) \qquad \textbf{(2.16)}$$

$$\Psi_{dr}^{s} = L_m i_{ds}^{s} + L_r i_{dr}^{s} \qquad \textbf{(2.17)}$$

$$\Psi_{qr}^{s} = L_m i_{qs}^{s} + L_r i_{qr}^{s} \qquad \textbf{(2.18)}$$

Eliminating i_{dr}^{s} and i_{qr}^{s} from the equations 2.17and 2.18 with the help of the equations 2.15 and 2.16 respectively, we get,

$$\Psi_{dr}^{s} = \frac{L_r}{L_m} \Psi_{dm}^{s} - L_{lr} i_{ds}^{s} \qquad \textbf{(2.19)}$$

$$\Psi_{qr}^{s} = \frac{L_r}{L_m} \Psi_{qm}^{s} - L_{lr} i_{qs}^{s} \qquad \textbf{(2.20)}$$

For decoupling control, we can now make derivation of control equations of indirect vector control with the help of equivalent circuits. The rotor circuit equations can be written as

$$\frac{d\Psi_{dr}}{dt} + R_r i_{dr} - (\omega_e - \omega_r)\Psi_{qr} = 0 \quad \textbf{(2.21)}$$

$$\frac{d\Psi_{qr}}{dt} + R_r i_{qr} + (\omega_e - \omega_r)\Psi_{dr} = 0 \quad \textbf{(2.22)}$$

The rotor flux linkage expression can be given as

$$\Psi_{dr} = L_r i_{dr} + L_m i_{ds} \quad \textbf{(2.23)}$$

$$\Psi_{qr} = L_r i_{qr} + L_m i_{qs} \quad \textbf{(2.24)}$$

From the above equations we can write

$$i_{dr} = \frac{1}{L_r}\Psi_{dr} - \frac{L_m}{L_r} i_{ds} \quad \textbf{(2.25)}$$

$$i_{qr} = \frac{1}{L_r}\Psi_{qr} - \frac{L_m}{L_r} i_{qs} \quad \textbf{(2.26)}$$

The rotor currents in equations2.21 and 2.22 which are in accessible can be eliminated with the help of equations 2.25 and 2.26 as,

$$\frac{d\Psi_{dr}}{dt} + \frac{R_r}{L_r}\Psi_{dr} - \frac{L_m}{L_r} R_r i_{ds} - \omega_{sl}\Psi_{qr} = 0 \quad \textbf{(2.27)}$$

$$\frac{d\Psi_{qr}}{dt} + \frac{R_r}{L_r}\Psi_{qr} - \frac{L_m}{L_r} R_r i_{qs} - \omega_{sl}\Psi_{dr} = 0 \quad \textbf{(2.28)}$$

Where $\omega_{sl} = \omega_e - \omega_r$ has been substituted.

For decupling control, it is desirable that

$$\Psi_{qr} = 0 \quad \textbf{(2.29)}$$

That is

$$\frac{d\Psi_{qr}}{dt} = 0 \quad \textbf{(2.30)}$$

So that, the total rotor flux $\hat{\Psi}_r$ is directed on the d^e axis.

Substituting the above conditions in equations 2.29 and 2.30, we get

$$\frac{L_r}{R_r}\frac{d\hat{\Psi}_r}{dt}+\hat{\Psi}_r = L_m i_{ds} \qquad \textbf{(2.31)}$$

$$\omega_{sl} = \frac{L_m R_r}{\hat{\Psi}_r L_r} i_{qs} \qquad \textbf{(2.32)}$$

Where $\hat{\Psi}_r = \Psi_{dr}$ has been substituted.

If rotor flux $\hat{\Psi}_r$ =constants, which is usually the case, then from equation 1.30

$$\hat{\Psi}_r = L_m i_{ds} \qquad \textbf{(2.33)}$$

In other words, the rotor flux is directly proportional to current i_{ds} in steady state.

To implement the indirect vector control strategy, it is necessary to take 2.11, 2.31 and 2.32 into considerations. The speed control range in indirect vector control can easily be extended from stand still or zero speed to the field weakening region how ever in field weakening region, the flux is programmed such that the inverter all ways operated in PWM mode. In both the direct and indirect control methods instantaneous current control of inverter is necessary. Hysterisis bend PWM current control can be used but it harmonic content is not optimum. Besides at higher speed the current controller will tent to saturate in part the quasi PWM because of higher CEMF. In this conditions, the fundamental current magnitude its phase lose tracking with the command current and thus, vector control will not be valid.

CHAPTER 3

SENSORLESS SLIDING MODE VECTOR CONTROL

3.1 INTRODUCTION:

Control actions of the systems under study are assumed to be discontinuous function of the system state. For the principle operation mode the state trajectories are in the vicinity of discontinuity points. This motion is referred to as sliding mode. The scope of sliding mode control studies embraces - mathematical methods of discontinuous differential equations, - design of manifolds in the state space and discontinuous control functions enforcing motions along the manifolds, - implementation of sliding mode controllers and their applications to control of dynamic plants. The first problem concerns development of the tools to derive the equations governing sliding modes and the conditions for this motion to exist. Formally motion equations of SMC do not satisfy the conventional uniqueness-existence theorems of the ordinary differential equations theory. The reasons of ambiguity are discussed. The regularization approach to derive sliding motion equations i demonstrated and compared with other models of sliding modes. The sliding mode existence problem is studied in terms of the stability theory. Enforcing sliding modes enables decoupling of the design procedure, since the motion preceding sliding mode and motion in sliding mode are of lower dimensions and may be designed independently. On the other hand, under so called matching conditions the sliding mode equations depend neither plant parameter variations nor external disturbances. Therefore sliding mode control algorithms are efficient when controlling nonlinear dynamic plants of high dimension operating under uncertainty conditions. The design methods are demonstrated mainly for systems in the regular form. Component-wise and vector design versions of sliding mode control are discussed. The design methodology is illustrated by sliding mode control in linear systems. The concept "sliding mode control" is generalized for discrete-time systems to make feasible its implementation for the systems with digital controllers. New mathematical and design methods are needed for sliding mode control in infinite-dimensional systems including systems governed by PDE. The recent results in this area are briefly surveyed. The problem of chattering caused by unmodeled dynamics is discussed in the context of applications. The systems with asymptotic observers are shown to be free of chattering. Sliding mode control applications is useful to electric drives, mobile robots, flexible bar and plate.

Sliding mode control is a type of variable structure control where the dynamics of a nonlinear system is altered via application of high frequency switching control. This is a state feedback control scheme where the feedback is not a continuous function of time. The principle of sliding mode control is to forcibly constrain the system, by suitable control strategy, to stay on the sliding surface on which the system will exhibit desirable features. When the system is constrained by the sliding control to stay on the sliding surface, the system dynamics are governed by reduced order system.

Sliding mode techniques are one approach to solving control problems and are an area of increasing interest. This text provides the reader with an introduction to the sliding mode control area and then goes on to develop the theoretical results. Fully worked design examples which can be used as tutorial material are included. Industrial case studies, which present the results of sliding mode controller implementations, and are used to illustrate successful practical applications of the theory.

In the formulation of any control problem there will typically be discrepancies between the actual plant and the mathematical model developed for controller design. This mismatch may be due to any number of factors and it is the engineer's role to ensure the required performance levels exist despite the existence of plant/model mismatches. This has led to the development of so-called robust control methods.

3.2 SENSOR LESS CONTROL:

Sensorless vector control of an induction motor drive essentially means vector control without any speed sensor. An incremental shaft mounted speed encoder usually an optical type is required for close loop speed or position control in both vector and scalar control drives. A speed signal is also required in indirect vector control in the whole speed range, and in direct vector control for the low speed range, including the zero speed start up operation. A speed encoder is undesirable in the drive because it adds cost and reliability problem, besides the need for a shaft extension and mounting arrangement. It is possible to estimate the speed signal from machine terminal voltages and current with the help of a dsp. However the estimation is normally complex and heavily depends on the machine parameters. Although the sensorless vector controlled drives are commercially available this time, the parameter variation problem, particularly near zero speed, impose a challenge in the accuracy of speed estimation.

3.3 SLIDING MODE CONTROL:

A sliding mode control (SMC) with a variable structure is basically an adaptive control that gives robust performance of a drive with parameter variation and load torque disturbance. The control is nonlinear and can be applied to a linear or nonlinear plant. In an SMC, the name indicates the drive response is forced to tract or slide along a predefined trajectory or reference model in a phase plane by a switching control algorithm, irrespective of the plant’s parameter variation and load disturbance. The controller detects the deviation of the actual trajectory and correspondingly changes the switching strategy to restore the tracking. In performance, it is somewhat similar to an MRAC, but the design and implementation of an SMC are somewhat simpler. SMCS can be applied to servo drives with dc motors, induction motors, and synchronous motor for application such as robot drives, machine tool control, etc.

One particular approach to robust control controller design is the so-called sliding mode control methodology which is a particular type of Variable Structure Control System (VSCS). Variable Structure Control Systems are characterised by a suite of feedback control laws and a decision rule (termed the switching function) and can be regarded as a combination of subsystems where each subsystem has a fixed control structure and is valid for specified regions of system behavior. The advantage is its ability to combine useful properties of each of the composite structures of the system. Furthermore, the system may be designed to possess new properties not present in any of the composite structures alone.

In sliding mode control, the VSCS is designed to drive and then constrain the system state to lie within a neighborhood of the switching function. Its two main advantages are (1) the dynamic behavior of the system may be tailored by the particular choice of switching function, and (2) the closed-loop response becomes totally insensitive to a particular class of uncertainty. Also, the ability to specify performance directly makes sliding mode control attractive from the design perspective.

Sliding mode control is an efficient tool to control complex high-order dynamic plants operating under uncertainty conditions due to its order reduction property and its low sensitivity to disturbances and plant parameter variations. Its robustness property comes with a price, which is high control activity. The principle of sliding mode control is that; states of the system to be controlled are first taken to a surface (sliding surface) in state space and then kept there with a shifting law based on the system states. Once sliding surface is reached the closed loop system has low sensitivity to matched and bounded disturbances, plant parameter variations.

Sliding mode control can be conveniently used for both non-linear systems and systems with parameter uncertainties due to its discontinuous controller term. That discontinuous control term is used to negate the effects of non-linearities and/or parameter uncertainties.

3.4 BLOCK DIAGRAM OF SLIDING MODE FIELD ORIENTED CONTROL:

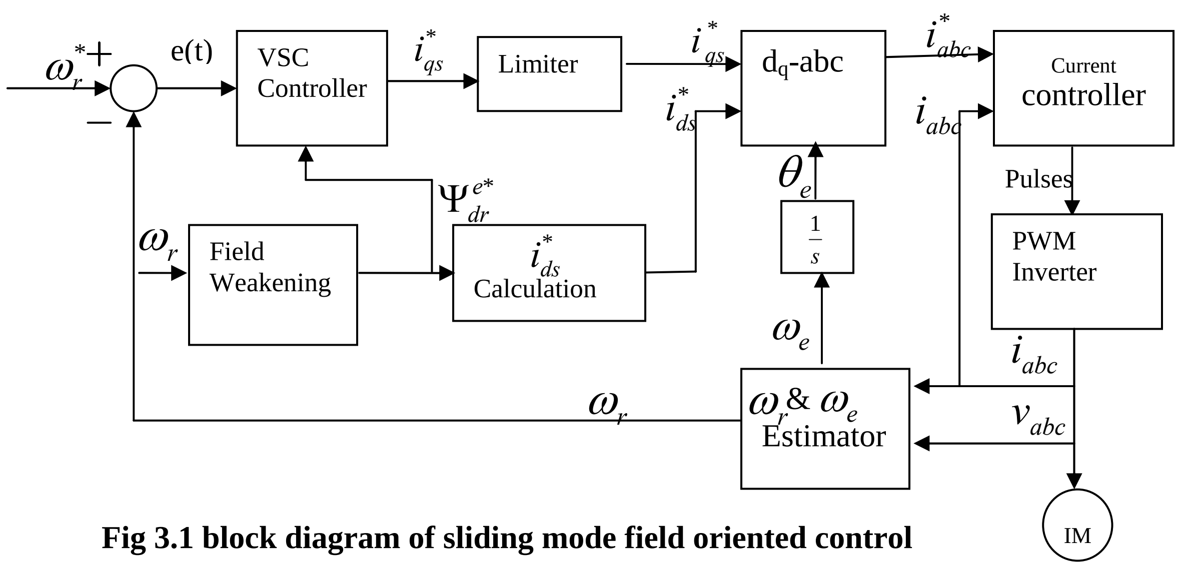

Fig 3.1 block diagram of sliding mode field oriented control

3.5 VARIABLE STRUCTURE ROBUST SPEED CONTROL:

In general, the mechanical equation of an induction motor can be written as:

$$J\dot{w}_m + Bw_m + T_L = T_e \qquad \textbf{(3.1)}$$

Where J and B are the inertia constant and the viscous friction coefficient of the induction motor system, respectively; T_L is the external load; w_m the rotor mechanical speed in angular frequency, which is related to the rotor electrical speed by,

$$\omega_m = \frac{2\omega_r}{p} \qquad \textbf{(3.2)}$$

Where p is the pole numbers and T_e denotes the generated torque of an induction motor, defined as,

$$T_e = \frac{3p}{4}\frac{L_m}{L_r}(\psi^e_{dr} i^e_{qs} - \psi^e_{qr} i^e_{ds}) \qquad \textbf{(3.3)}$$

where ψ^e_{dr} and ψ^e_{qr} are the rotor-flux linkages, with the subscript 'e' denoting that the quantity is referred to the synchronously rotating reference frame; i^e_{qs} and i^e_{ds} are the stator currents.

Then the mechanical equation becomes,

$$\dot{\omega}_m + a\omega_m + f = bi^e_{qs} \qquad \textbf{(3.4)}$$

Where the parameters are defined as,

$$a = \frac{B}{J}, b = \frac{K_T}{J}, f = \frac{T_L}{J} \qquad \textbf{(3.5)}$$

Now, we are going to consider the previous mechanical Equation with uncertainties as follows:

$$\dot{\omega}_m = -(a + \Delta a)\omega_m + (f + \Delta f) + (b + \Delta b)i^e_{qs} \qquad \textbf{(3.6)}$$

Where the terms $\Delta a, \Delta b$ and Δf are the uncertainty of the parameters a, b, and f respectively.

Let us define the tracking speed error as follows:

$$e(t) = \omega_m(t) - \dot{\omega}_m(t) \qquad \textbf{(3.7)}$$

Where, the $\dot{\omega}_m$ is the rotor speed command.

Taking the derivative of the previous equation with respect to time yields:

$$\dot{e}(t) = \dot{\omega}_m - \dot{\omega}_m^{*} = -ae(t) + u(t) + d(t) \qquad \textbf{(3.8)}$$

Where the following terms have been collected in the signal *u* (*t*),

$$u(t) = bi^{e}{}_{qs}(t) - a\omega^{*}{}_{m}(t) - f(t) - \dot{\omega}_m^{*}(t) \qquad \textbf{(3.9)}$$

And, the uncertainty terms have been collected in the signal *d(t),*

$$d(t) = -\Delta a\,\omega_m(t) - \Delta f(t) + \Delta b i^{e}{}_{qs}(t) \qquad \textbf{(3.10)}$$

Now, we are going to define the sliding variable *S* (*t*) with an integral component as:

$$s(t) = e(t) - \int_0^t (k-a)e(\tau)d\tau \qquad \textbf{(3.11)}$$

Where, k is a constant gain.

Then the sliding surface is defined as:

$$s(t) = e(t) - \int_0^t (k-a)e(\tau)d\tau = 0 \qquad \textbf{(3.12)}$$

The variable structure speed controller is designed as:

$$u(t) = ke - \beta\,\mathrm{sgn}(s) \qquad \textbf{(3.13)}$$

In order to obtain the speed trajectory tracking, the following assumption should be formulated,

- The gain k must be chosen so that the term (k a) is strictly negative, therefore $k < 0$.

- The gain β must be chosen so that $\beta \geqslant |d(t)|$ for all time.

If the assumption is verified, the control law 3.13 leads the rotor mechanical speed so that the speed tracking error $e(t) = \omega_m(t) - \omega^*{}_m(t)$ tends to zero if the time tends to infinity. The proof of this theorem will be carried out using the Lyapunov stability theory.

$$v(t) = \frac{1}{2} s(t) s(t) \tag{3.14}$$

Its time derivative is calculated as,

$$\begin{aligned} &\dot{V}(t) \\ &= s(t)\dot{s}(t) = s[\dot{e} - (k-a)e] \\ &= s[(-ae + u + d) - (ke - ae)] = s[u + d - ke] \\ &= s[ke - \beta\,\mathrm{sgn}(s) + d - ke] = s[d - \beta\,\mathrm{sgn}(s)] \\ &\le -[\beta - |d|\, s] \le 0 \end{aligned} \tag{3.15}$$

Using the Lyapunov's direct method, since $V(t)$ is clearly positive-definite, $\dot{V}(t)$ is negative definite and $V(t)$ tends to infinity as $S(t)$ tends to infinity, then the equilibrium at the origin $S(t) = 0$ is globally asymptotically stable. Therefore $S(t)$ tends to zero as the time t tends to infinity. Moreover, all trajectories starting off the sliding surface $S = 0$ must reach it in finite time and then will remain on this surface. This system's behavior once on the sliding surface is usually called sliding mode.

When the sliding mode occurs on the sliding surface then $S(t) = \dot{S}(t) = 0$, and therefore, the dynamic behavior of the tracking problem is equivalently governed by the following equation:

$$\begin{aligned} &\dot{s}(t) = 0 \\ &\Rightarrow \dot{e}(t) = (k-a)e(t) \end{aligned} \tag{3.16}$$

Then, under assumption 1, the tracking error $e(t)$ converges to zero exponentially.

It should be noted that, a typical motion under sliding mode control consists of a reaching phase during which trajectories starting off the sliding surface S = 0 move toward it and reach it in finite time, followed by sliding phase during which the motion will be confined to this surface and the system tracking error will be represented by the reduced-order model, where the tracking error tends to zero.

Finally, the torque current command, $i^{*}_{qs}(t)$, can be obtained directly substituting u (t) in the previous equation,

$$i^{*}_{qs}(t) = \frac{1}{b}[ke - \beta \operatorname{sgn}(s) + a\omega^{*}_{m} + \dot{\omega}_{m}^{*} + f] \qquad \textbf{(3.17)}$$

Therefore, the variable structure speed control resolves the speed tracking problem for the induction motor, with some uncertainties in mechanical parameters and load torque.

3.6 CURRENT CONTROLLER:

The block current controller consists of three hysteresis band current PWM control. it is basically an instantaneous feedback current control method of PWM where the actual current continuously tracks the current command within a hysteresis band.

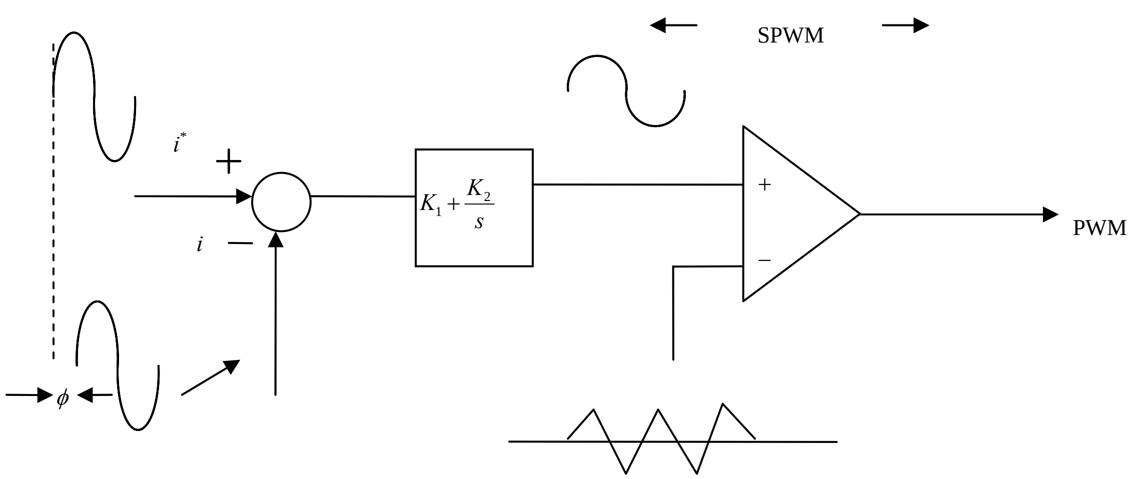

Fig 3.2 hysteresis band current control PWM

The control circuit generates the sin reference current wave of desired magnitude and frequency and it is compared with the actual phase current wave. As the current exceeds the prescribed hysteresis band, the upper switch in the half bridge is turned off and lower is turned on. The output voltage transitions are from +0.5Vd to -0.5Vd, and the current start decay.

As the current crosses the lower band limit, the lower switch is turned off and the upper switch is turned on. A lock out time is provided at each transition to prevent a shoot through fault. The actual current wave is thus forced to track the sine reference wave with in the hysteresis band by back and forth switching of the upper and lower switches. The inverter then essentially becomes a current source with peak to peak current ripple, which is controlled with in the hysteresis band irrespective of Vd fluctuations.

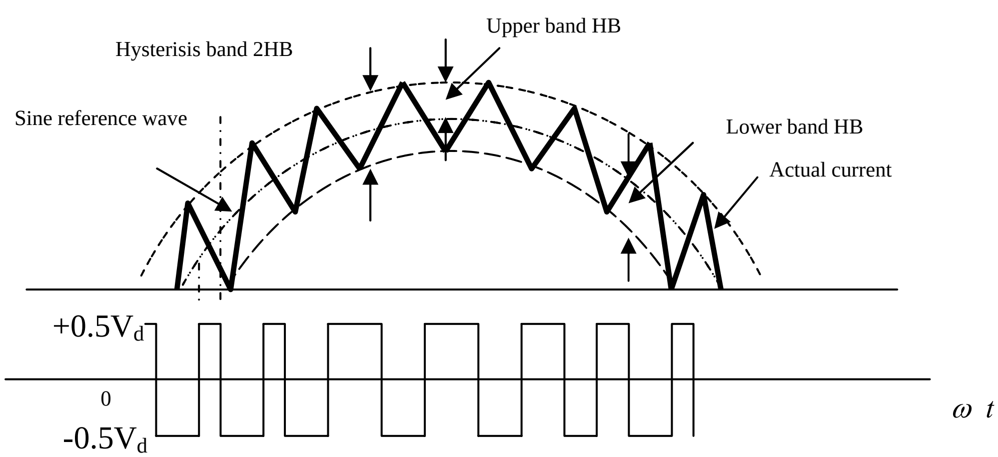

Fig 3.3 principle of hysteresis band control

When the upper switch is closed, the positive slope of the current is given as,

$$\frac{di}{dt} = \frac{0.5V_d - V_{cm}\sin\omega_e t}{L} \qquad \textbf{(3.18)}$$

Where 0.5 V_d is the applied voltage.

$V_{cm}\sin\omega_e t$ = instantaneous value of the opposing load CEMF.

L = effective load inductance.

The corresponding equation when the lower switch is closed is given as,

$$\frac{di}{dt} = \frac{-(0.5V_d - V_{cm}\sin\omega_e t)}{L} \tag{3.19}$$

The peak to peak current ripple and switching frequency are related to the width of the hysteresis band. A smaller band will increase switching frequency and lower the ripple. The hysteresis band PWM can be smoothly transitioned to square wave voltage mode through the quasi PWM region.

When upper switch on $(i^* - i) > HB$

When lower switch on $(i^* - i) < -HB$

The hysteresis band PWM has been very popular because of its simple implementation, fast transient response, direct limiting of device peak current and practical insensitivity of dc link voltage ripple that permits a lower filter capacitor.

3.7 ESTIMATION OF MOTOR SPEED:

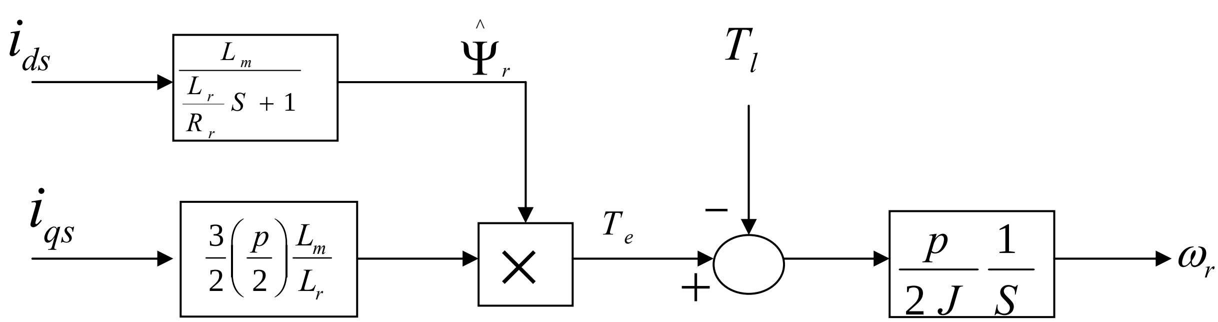

Fig 3.4 transfer function block diagram of vector control drives

From the indirect vector control the stator voltage equation in the stationary frame is given as,

$$\dot{\Psi}_{dr} = \frac{L_r}{L_m}v_{ds} - \frac{L_r}{L_m}(R_s + \sigma L_s\frac{d}{dt})i_{ds} \tag{3.20}$$

$$\dot{\Psi}_{qr} = \frac{L_r}{L_m} v_{qs} - \frac{L_r}{L_m}(R_s + \sigma L_s \frac{d}{dt}) i_{qs} \quad \textbf{(3.21)}$$

The rotor flux equations in the stationary frame are

$$\dot{\Psi}_{dr} = \frac{L_m}{T_r} i_{ds} - \omega_r \Psi_{qr} - \frac{1}{T_r} \Psi_{dr} \quad \textbf{(3.22)}$$

$$\dot{\Psi}_{qr} = \frac{L_m}{T_r} i_{qs} + \omega_r \Psi_{dr} - \frac{1}{T_r} \Psi_{qr} \quad \textbf{(3.23)}$$

Where $T_r = \frac{L_r}{R_r}$

And the angle between rotor flux vectors in relation to the d axis of the stationary frame is,

$$\theta_e = \arctan(\frac{\Psi_{qr}}{\Psi_{dr}}) \quad \textbf{(3.24)}$$

Taking its derivative of the above equation, we get,

$$\dot{\theta}_e = \omega_e = \frac{\Psi_{dr} \dot{\Psi}_{qr} - \Psi_{qr} \dot{\Psi}_{dr}}{\Psi_{dr}^2 + \Psi_{qr}^2} \quad \textbf{(3.25)}$$

Simplifying the above equations,

$$\omega_e = \omega_r - \frac{L_m}{T_r}(\frac{\Psi_{dr} i_{qs} - \Psi_{qr} i_{ds}}{\Psi_{dr}^2 + \Psi_{qr}^2}) \quad \textbf{(3.26)}$$

$$\omega_r = \frac{1}{\Psi_r^2}[\Psi_{dr} \dot{\Psi}_{qr} - \Psi_{qr} \dot{\Psi}_{dr} + \frac{L_m}{T_r}(\Psi_{dr} i_{qs} - \Psi_{qr} i_{ds})] \quad \textbf{(3.27)}$$

Where

$$\Psi_r^2 = \Psi_{dr}^2 + \Psi_{qr}^2 \quad \textbf{(3.28)}$$

Therefore, given a complete knowledge of the motor parameters, the instantaneous speed ω_r can be calculated from the previous equation, where the stator measured current and voltages, and the rotor flux estimated obtained from a rotor flux observer based on Equation 3.5 and 3.6 are employed.

The generated torque in the induction motor is given as,

$$T_e = \frac{3p}{4}\frac{L_m}{L_r}(\Psi^e_{dr} i^e_{qs} - \Psi^e_{qr} i^e_{ds}) \quad \textbf{(3.29)}$$

Using the field-orientation control principle the current component Ids is aligned in the direction of the rotor flux vector ψr, and the current component *Iqs* is aligned in the direction perpendicular to it.

For decoupling control, it is desirable that,

$$\Psi^e_{qr} = 0, \qquad \Psi^e_{dr} = |\Psi_r| \quad \textbf{(3.30)}$$

Taking into account of the previous result the equation of torque is given as,

$$T_e = \frac{3p}{4}\frac{L_m}{L_r}\Psi^e_{dr} i^e_{qs} = K_t i^e_{qs} \quad \textbf{(3.31)}$$

Where, K_t is the torque constant and defined as follows,

$$K_t = \frac{3p}{4}\frac{L_m}{L_r}\Psi^e_{dr} \quad \textbf{(3.32)}$$

3.8 FIELD WEAKENING CONTROLLER:

The block field weakening gives the flux command based on the rotor speed, so that the PWM controller does not saturate.

If $\omega_r < \omega_b, \quad \Psi^*_{dr} = \Psi_{drRated}$

And $\omega_r > \omega_b, \quad \Psi^*_{dr} = \Psi_{drRated} \times \dfrac{\omega_b}{\omega_r}$ **(3.33)**

With the proper mentioned field orientation, the dynamic of the rotor flux is given as:

$$\frac{d\Psi^e_{dr}}{dt} + \frac{R_r}{L_r}\Psi^e_{dr} - \frac{L_m}{L_r}R_r i_{ds} - \omega_{sl}\Psi^e_{qr} = 0 \quad \textbf{(3.34)}$$

For decoupling,

$$\Psi^{e}_{qr} = 0, \frac{d\Psi^{e}_{qr}}{dt} = 0 \tag{3.35}$$

$$\frac{d\Psi^{e}_{dr}}{dt} + \frac{R_r}{L_r}\Psi^{e}_{dr} - \frac{L_m}{L_r}R_r i_{ds} = 0 \tag{3.36}$$

Multiplying the equation by Tr = Lr/Rr called rotor time constant and rearranging we get,

$$\frac{L_r}{R_r}\frac{d\Psi^{e}_{dr}}{dt} + \Psi^{e}_{dr} = L_m i_{ds} \tag{3.37}$$

If the rotor flux Ψ_r is constant then the above equation is reduced as,

$$\Psi_{dr} = L_m i_{ds} \tag{3.38}$$

$$\Psi_{r} = L_m i_{ds} \tag{3.39}$$

In other words the rotor flux is directly proportional to the current I_{ds} in steady state.

Here in this section the analysis of the sliding mode control is given. The torque is calculated by torque component of current. The rotor speed is estimated with out using any speed sensor or any type of transducer. All the mathematical formulation of the entire controller which is used in the block diagram is given in this section.

CHAPTER 4

RESULTS AND DISCUSSION

4.1 OPEN LOOP SIMULATION:

In the first section simulation for the open loop is given. Fig 4.1 is the torque speed characteristics of the induction motor in open loop control.

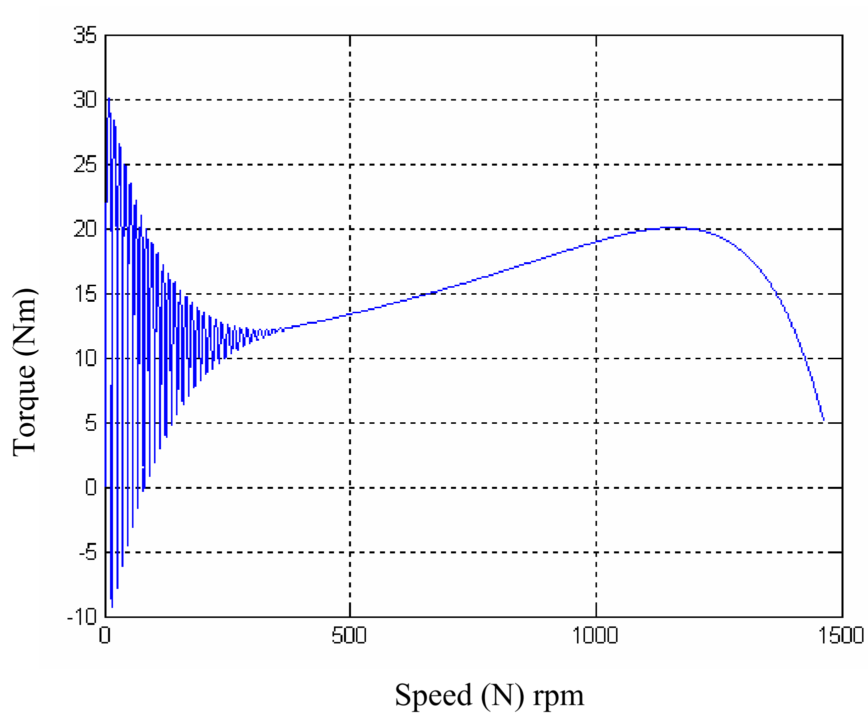

Fig 4.1 Torque speed characteristics

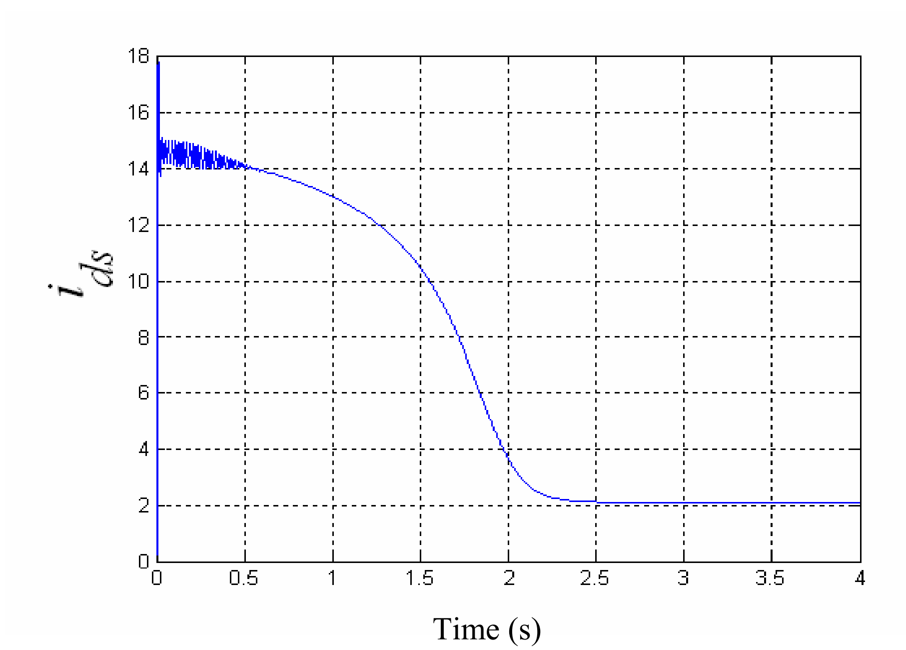

Fig4.2 plot of i_{ds}versus time

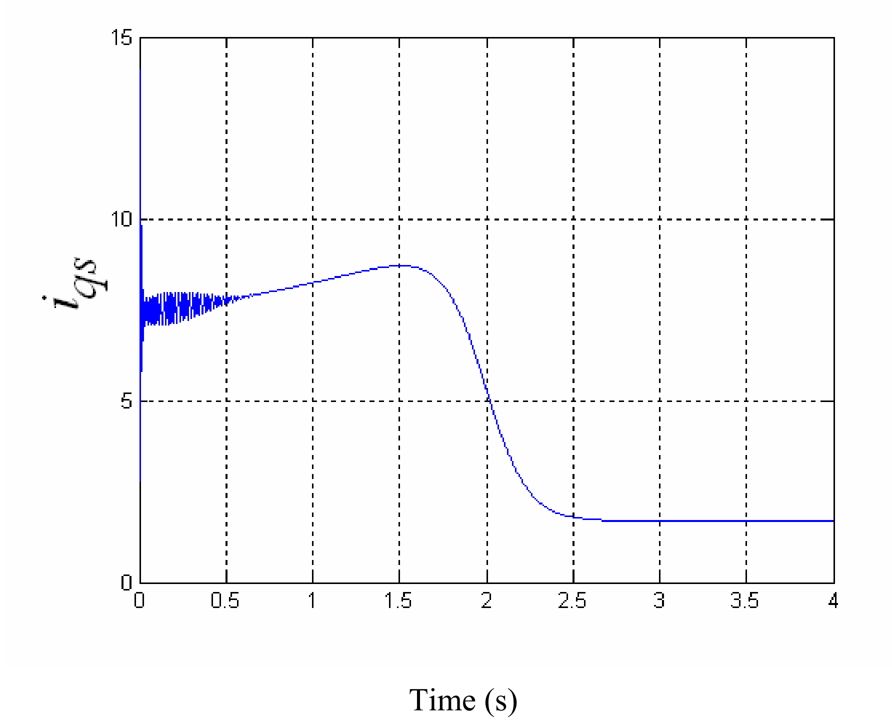

Fig 4.3 plot of i_{qs} versus time

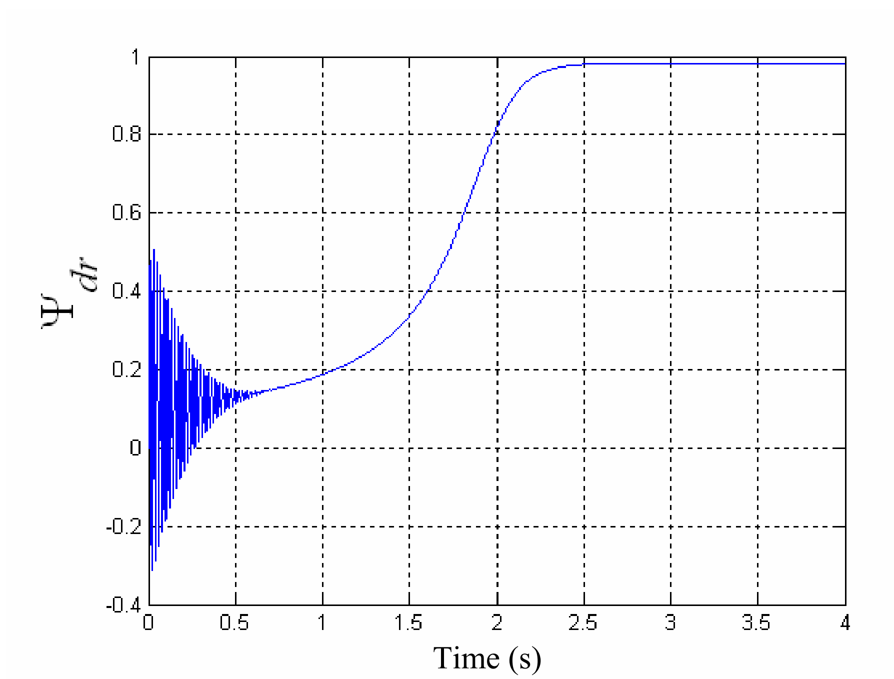

Fig4.4 plot of ψ_{dr} versus time

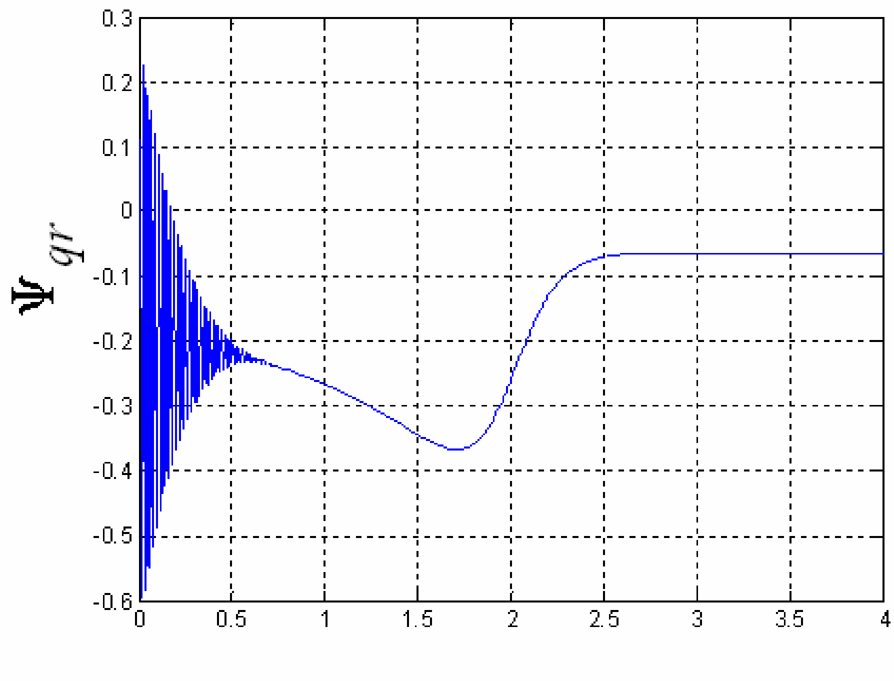

Fig 4.5 plot of ψ_{qr} versus time

In this section fig 4.2 and 4.3 is the plot of i_{ds} and i_{qs} versus time, here we have seen that as the time increases former is decreases very slowly but the later is first increases to a small value and then decreases rapidly. Upto the time 2.0 sec it is the transient period of the system and it comes to the steady state value.

Fig 4.4 and 4.5 is the plot of the rotor flux in d and q axis respectively. Here the rotor flux in the d axis is increases at the steady state where as the flux in the q axis decreases after the transient period, and it comes to negligible value.

The parameters of the induction motor are found out by the no load test and block rotor test, which is given in the appendices.

4.2 SIMULATION RESULTS FOR SLIDING MODE FIELD ORIENTED CONTROL:

The block 'VSC Controller' represents the proposed sliding mode controller and it is implemented by Equations. (3.11) and (3.17). The block limiter limits the current applied to the motor windings so that it remains within the limit value and it is implemented by a saturation function. The block 'dq^e →abc makes the conversion between the synchronously rotating and stationary reference frames and is implemented by Equation. (2.4). The block 'Current Controller' consists of a three hysteresis-band current PWM control, which is basically an instantaneous feedback current control methods of PWM where the actual current (i_{abc}) continually tracks the command current (i^*_{abc}) within a hysteresis band. The block 'PWM Inverter' is a six IGBT-diode bridge inverter with 780 V DC voltage source. The block 'Field Weakening controller' gives the flux command based on rotor speed, so that the PWM controller does not saturate. The block 'i^{e*}_{ds} Calculation' provides the current reference i^{e*}_{ds} from the rotor flux reference through the Equation (3.38).

The block 'ω_r and ω_e Estimator' represent the proposed rotor speed and synchronous speed estimator, and is implemented by equations (3.27) and (3.26) respectively. The block 'IM' represents the induction motor. All the electrical, mechanical and controller parameter are given in the appendices.

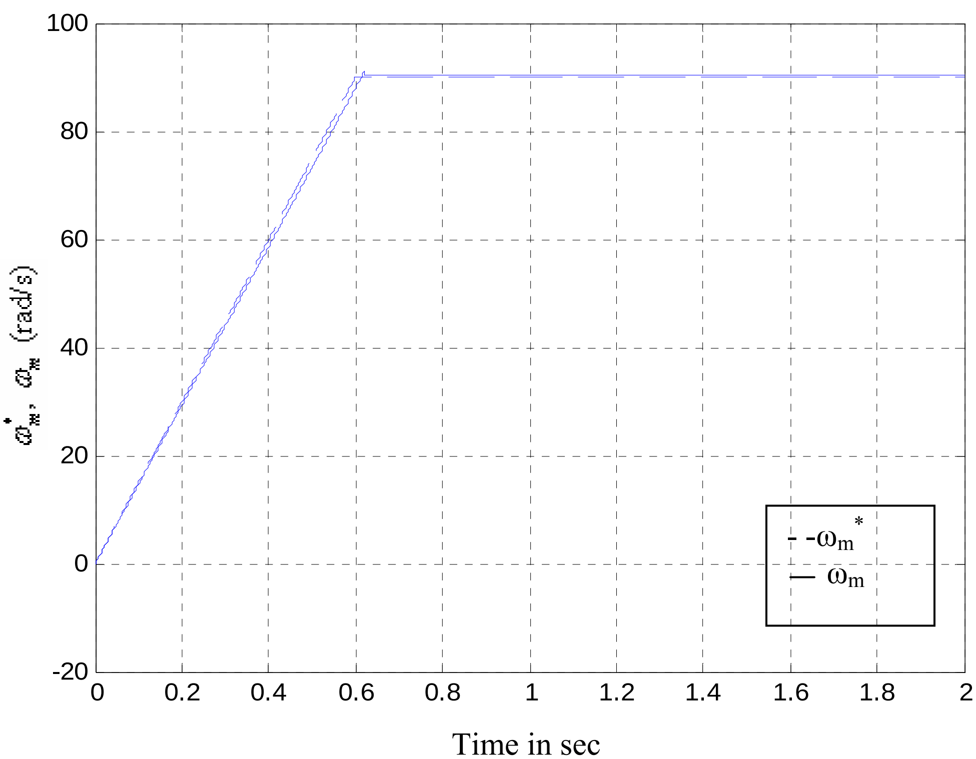

Fig 4.6 Reference and Real rotor speed signal (rad/sec)

The simulation results are shown through various figures. Figure.4.6 shows the desired rotor speed (dashed line) and the real rotor speed (solid line). The motor is allowed to start from a standstill state and the rotor speed to follow the speed command that starts from zero and accelerates until the rotor speed is 90rad/s. The system starts with an initial load torque T_L=50Nm, and at time t=1s the load torque steps to T_L=100Nm. From the figure it is observed that the rotor speed tracks the desired speed in spite of system uncertainties. The speed tracking is not affected by the load torque change at the time t=1 s, because when the sliding surface is reached (sliding mode) the system becomes insensitive to the boundary external disturbances.

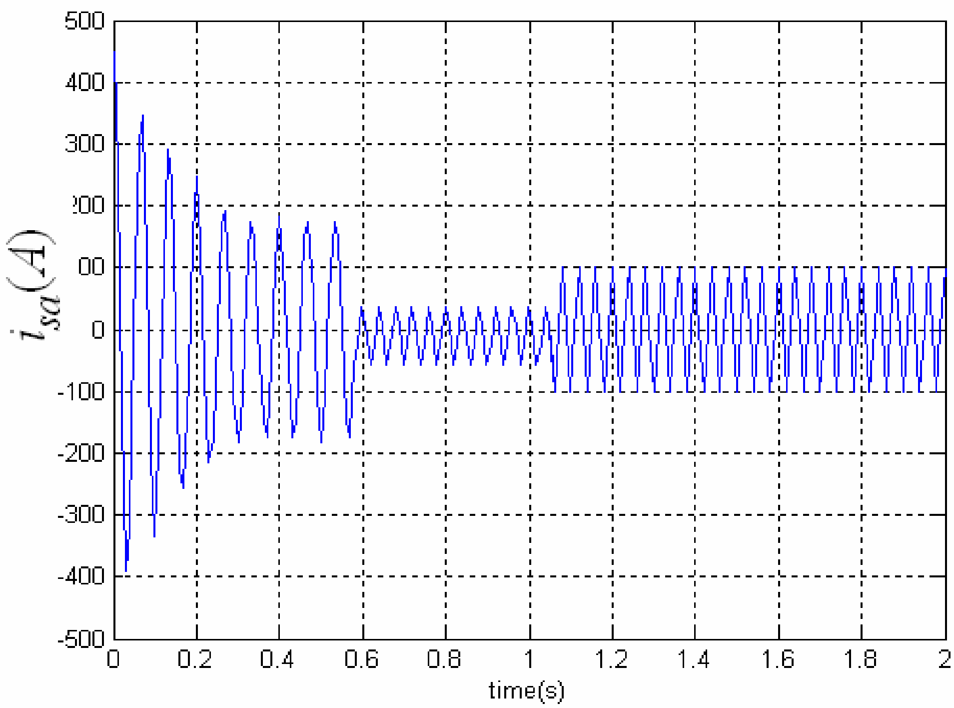

Fig 4.7 Stator current i_{sa} (A)

The fig 4.7 shows the current of one stator winding. This figure shows that in the initial state, the current signal presents a high value because it is necessary a high torque to increment the rotor speed. In the constant speed region, the motor torque only has to compensate the friction and the load torque and so, the current is lower. Finally, at time *t* = 1 s the current increases because the load torque has been increased.

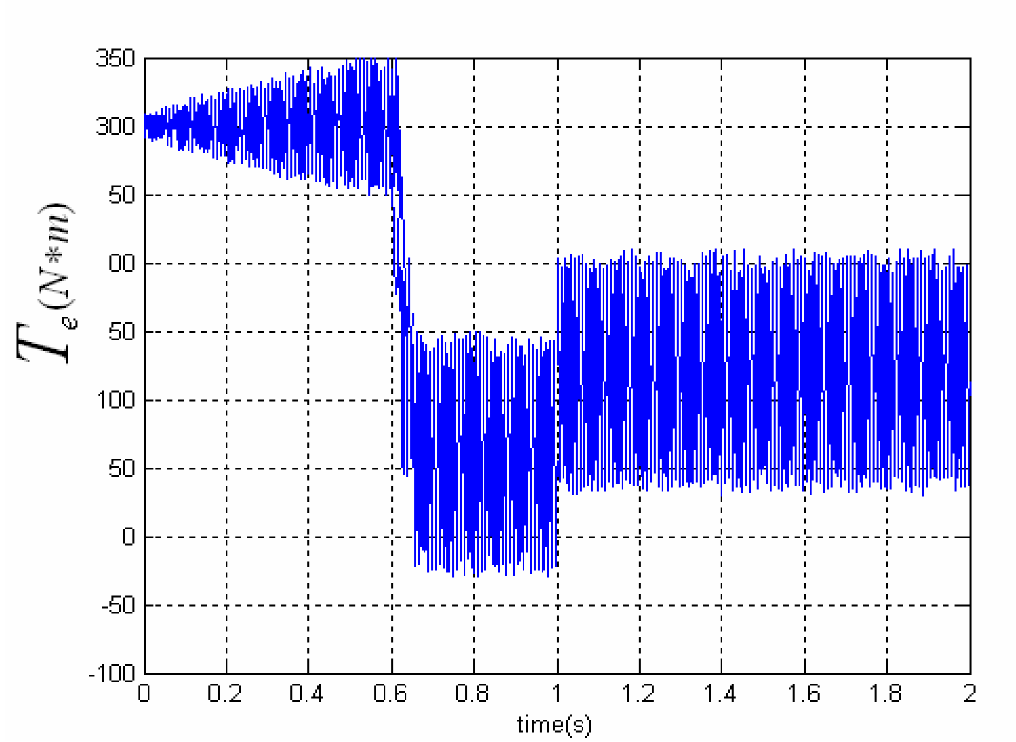

Fig 4.8 Motor torque (Nm)

Fig. 4.8 shows the motor torque. As in the case of the current (Fig. 4.2), the motor torque has a high initial value speed in the acceleration zone, then the value decreases in a constant region and finally increases due to the load torque increment.

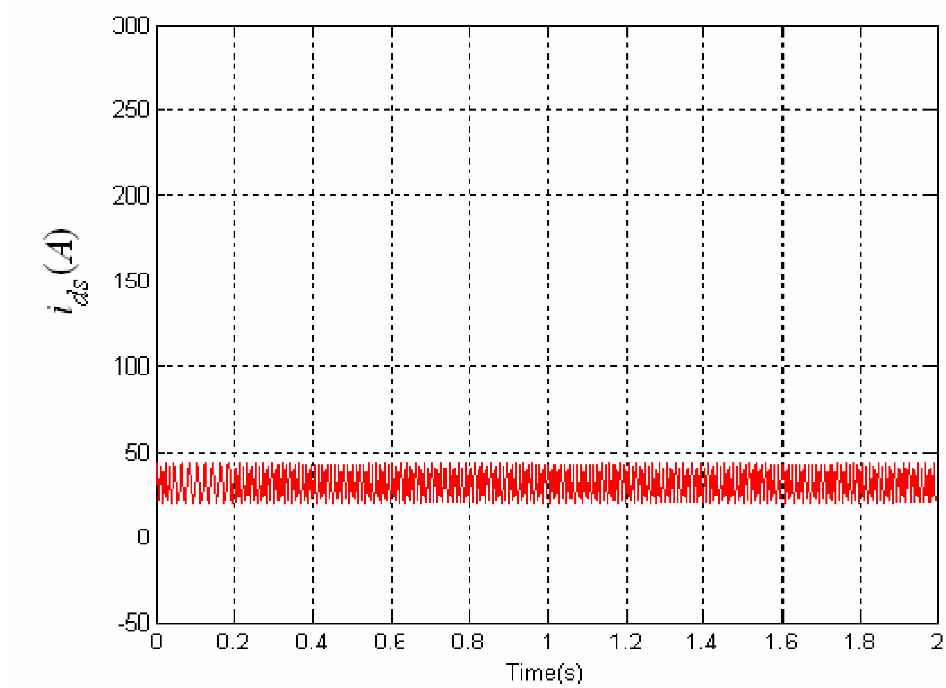

Fig 4.9 Stator current i_{ds} (A)

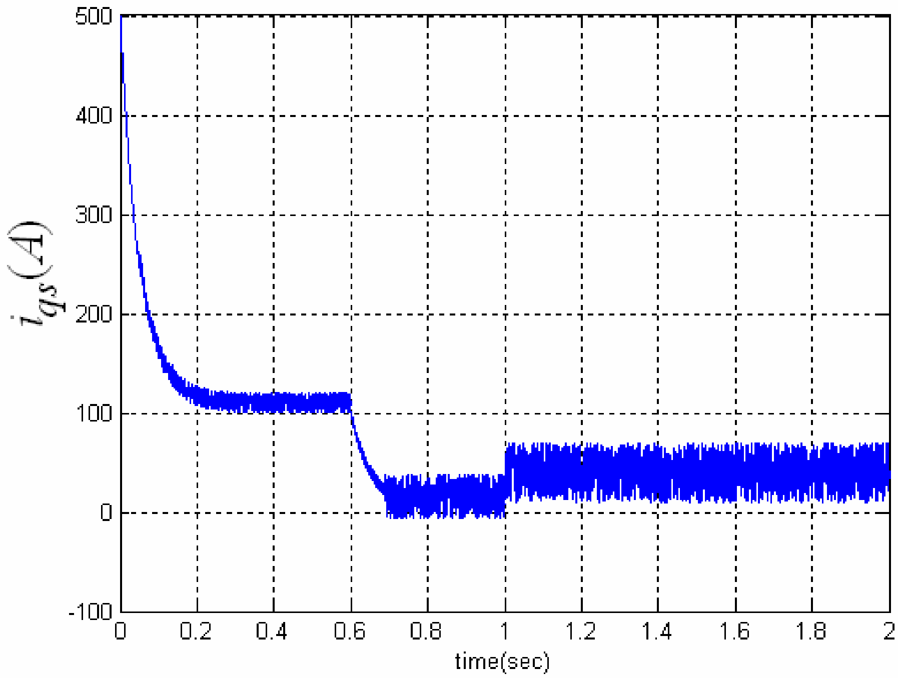

Fig 4.10 Stator current i_{qs}(A)

In this figure 4.8 it may be seen that in the motor torque appears the so-called chattering phenomenon, however, this high frequency changes in the torque will be filtered by the mechanical system inertia.

Figs. 4.9 and 4.10 show the stator currents in the rotating reference frame. As may be observed in the figures, both currents present an initial peak at the beginning, as it is usual in the starting of motors. Then the current, i_{ds}, corresponding to the field component, remains constant. On the other hand, the current i_{qs}, corresponding to the torque component, varies with the torque; that is, presents a high initial value in the acceleration zone, then the value decreases in a constant region and finally increases due to the load torque increment.

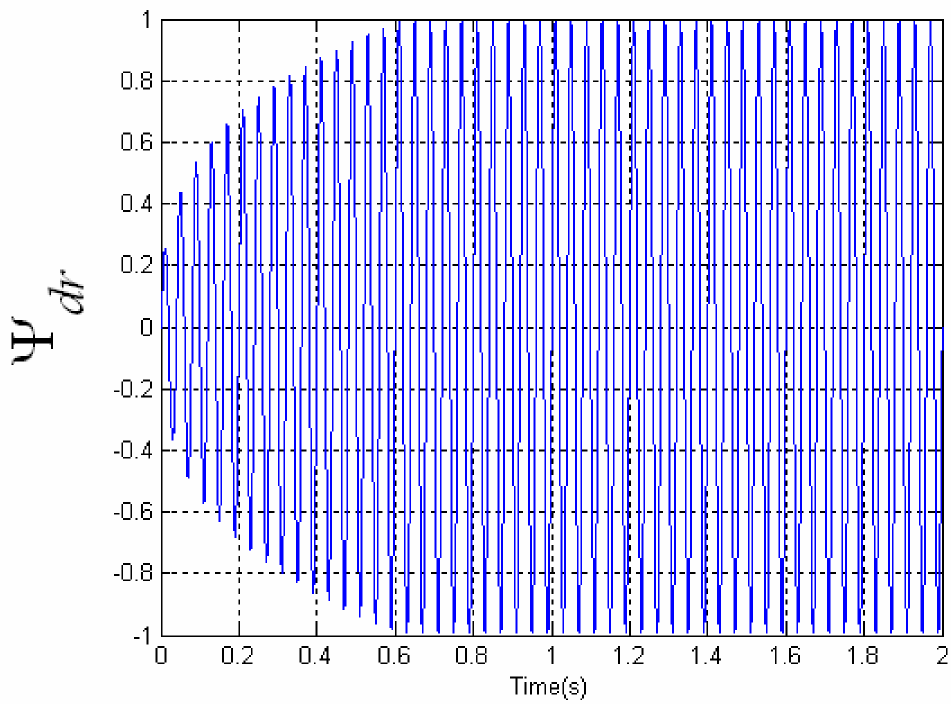

Fig 4.11 Rotor flux ψ_{dr} (Wb)

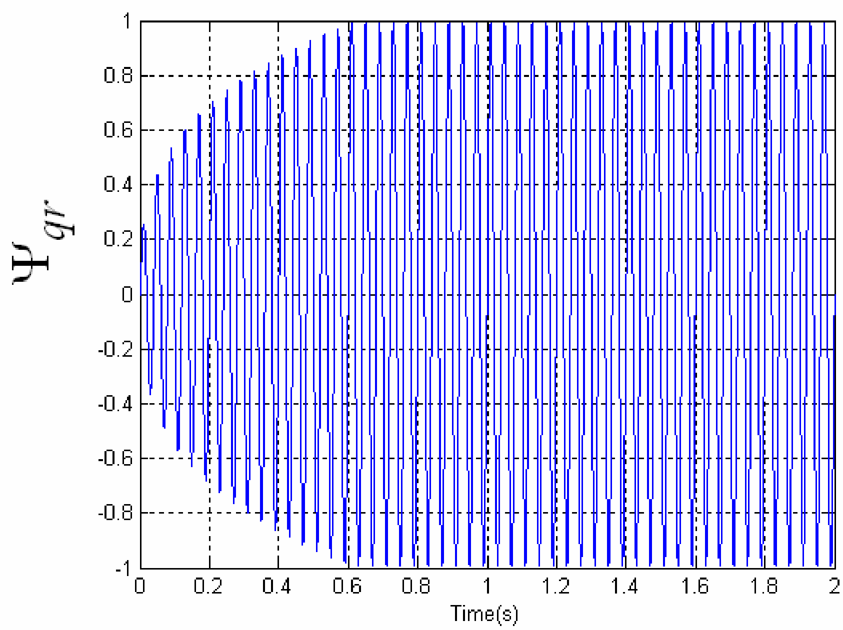

Fig 4.12 Rotor flux ψ_{qr} (Wb)

Figs. 4.11 and 4.12 show the estimated rotor flux in the stationary reference frame. As may be observed the rotor flux starts from zero and increases until the nominal value

4.3 SIMULATION RESULTS UNDER LOAD TORQUE VARIATION:

This figure no 4.13 is the real and reference speeds with a load torque variation from 50 Nm to 100 Nm then at time of 0.3 sec the speed reference changes from 90 rad/s to 120 rad/s. This shows the desired rotor speed with dashed line and the real rotor speed with solid line. When the speed reference steps to 0.3 sec the motor can not follow this reference instantaneously due to the physical limitations of the system. However, after a transitory time in which the motor accelerates until the final speed trajectory tracking is obtained.

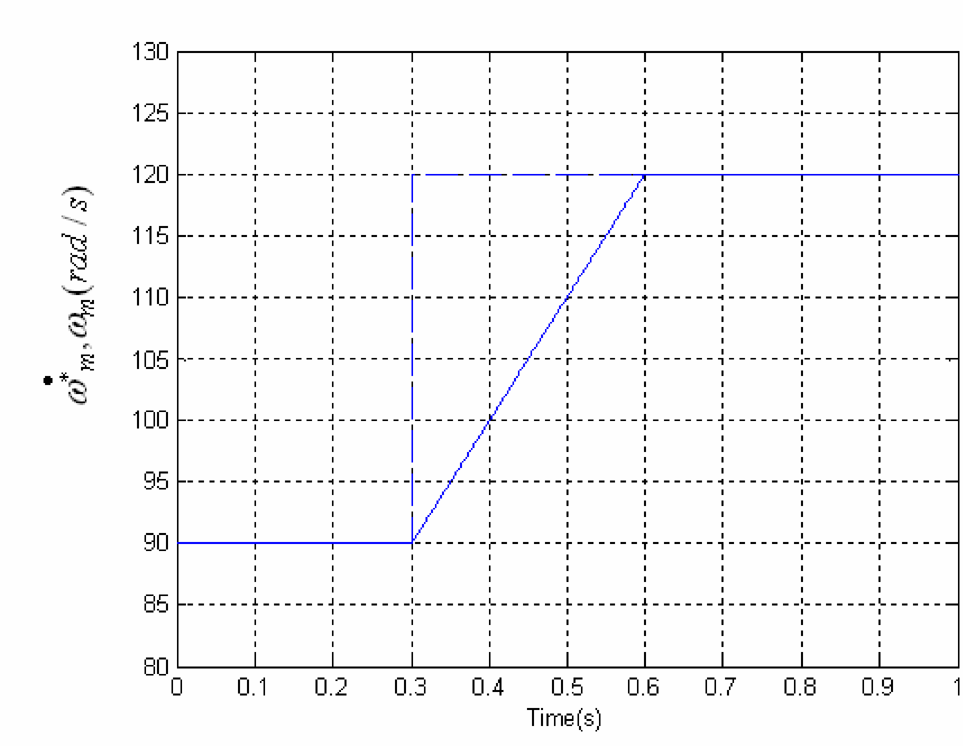

Fig 4.13 reference and real rotor speed signals (rad/s)

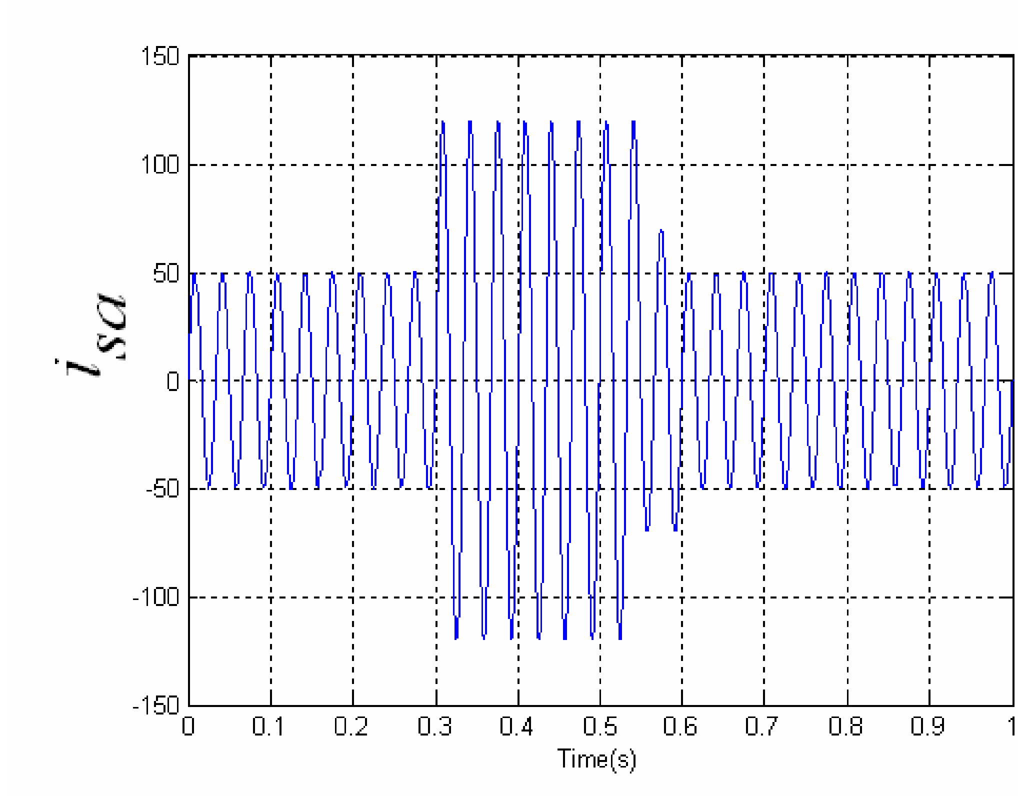

Fig 4.14 Stator current i_{sa}(A)

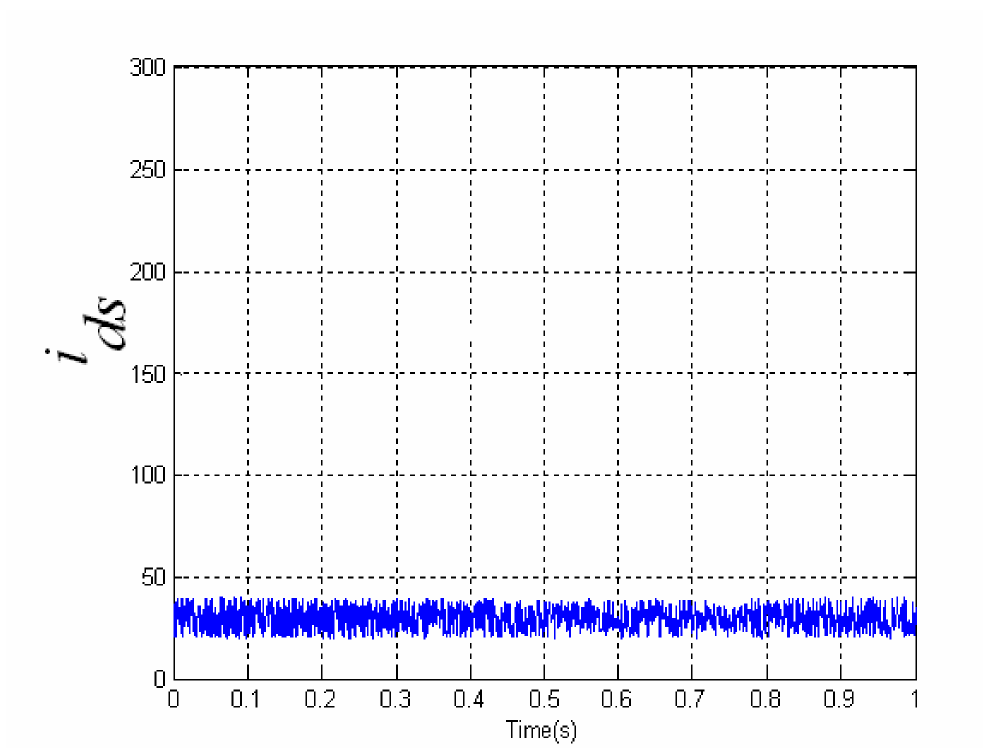

Fig 4.15 Stator current i_{ds} (A)

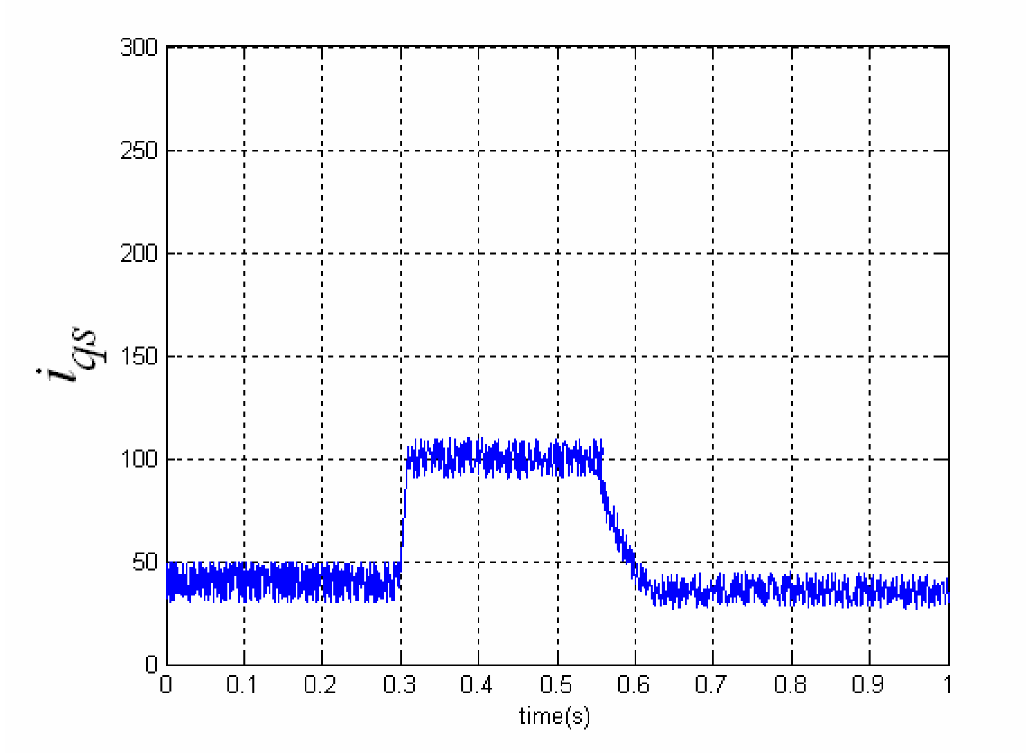

Fig 4.16 Stator current i_{qs} (A)

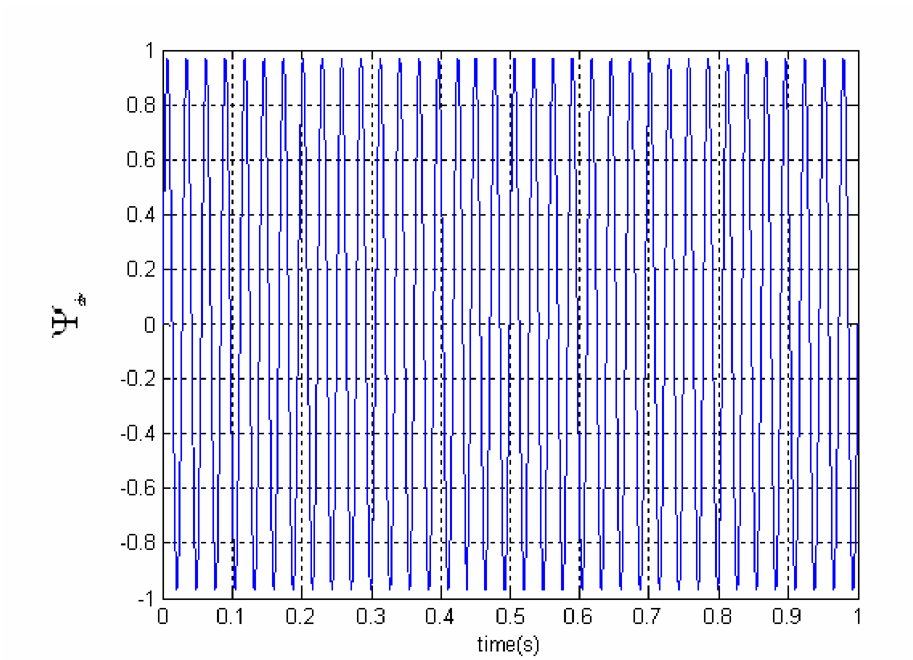

Fig 4.17 Rotor flux ψ_{dr} (Wb)

Figure 4.14 shows the current of one of the stator winding. This figure shows the initial state, the current signal is constant because the speed is constant, then the current increase because the motor is accelerating, and finally remains constant again because the speed is constant. At this point it is noted that the final current value is similar to the initial one.

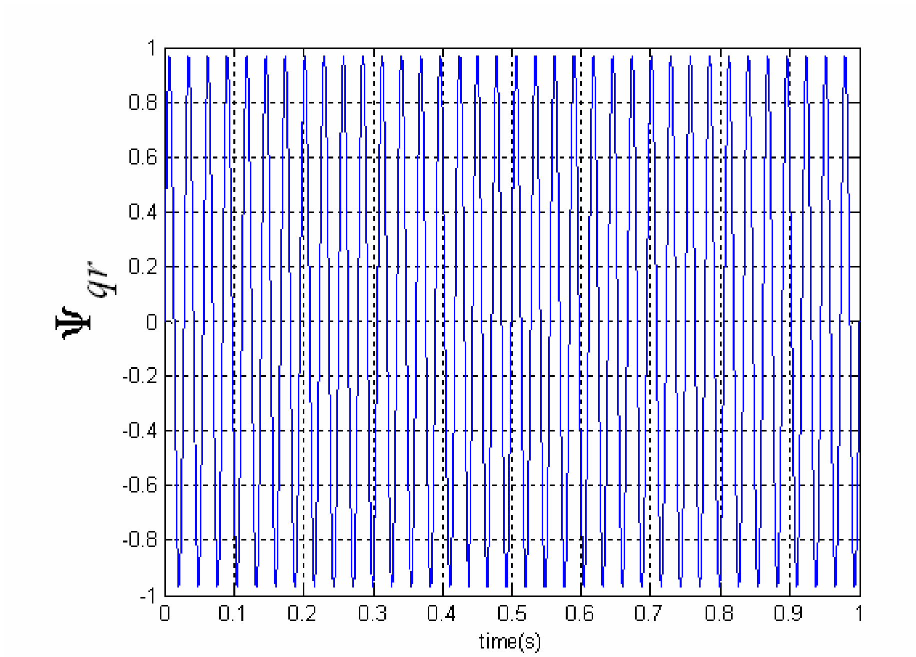

Fig 4.18 Rotor flux ψ_{qr} (Wb)

Fig 4.15 and fig 4.16 shows the stator currents in the rotating reference frame. As may be observed in the figures, the current $i^e{}_{ds}$, corresponding to the field component, remains constant because the flux is maintained constant, and current $i^e{}_{qs}$, corresponding to the torque component, varies with the torque.

Finally figure 4.17 and 4.18 shows the estimated flux in the stationary reference frame. The rotor flux remains constant at the nominal value.

Here in this section the open loop model of an induction motor is studied by the help of the simulation results and the sliding mode vector control of the motor is studied and also the simulation results are presented with load torque variation.

CHAPTER 5

FUTURE WORK

AND

CONCLUSION

5.1 SUMMARY AND CONCLUSIONS:

Sliding Mode Control (SMC) is a robust control scheme based on the concept of changing the structure of the controller in response to the changing state of the system in order to obtain a desired response. The biggest advantage of this system is stabilizing properties are preserved, even in the presence of large disturbance signals. The dynamic behavior of the system may be tailored by the particular choice of switching function and the closed-loop response becomes totally insensitive to a particular class of uncertainty. Also, the ability to specify performance directly makes sliding mode control attractive from the design perspective. One of the problems associated with implementation of SMC is Chattering which is essentially a high frequency switching of the control. Chattering in torque & speed may large, but can be minimized by small computation sampling time higher pwm frequency & minimizing additional delay in feedback signal.

In this work the rotor speed estimator is based on stator voltage equations and rotor flux equations in the stationary reference frame. Here a variable structure control which has an integral sliding surface to relax the requirement of the acceleration signal, that is usual in conventional sliding mode speed control techniques. Due to the nature of the sliding control this control scheme is robust under uncertainties caused by parameter errors or by changes in the load torque. The closed loop stability of the presented design has been proved through Lyapunov stability theory. Finally, by means of simulation examples, it has been shown that the control scheme performs reasonably well in practice, and that the speed tracking objective is achieved under uncertainties in the parameters and load.

5.2 FUTURE STUDY:

Following areas of further study are considered for more and more research:-

- Application of higher order sliding mode control to other non linear systems may be attempted.
- A second order discrete sliding mode control law may be developed.
- Performance of 2-SM algorithms may be studied.

APPENDIX

1. Data for the open loop control:

Sl No	Parameters calculated	Numerical value
1.	Lm (Magnetizing inductance)	0.5 H
2.	Lo (No load stator inductance)	0.53 H
3.	Ls (Stator inductance)	0.53 H
4.	X_{br} (Blocked rotor inductance)	13.921 Ohms
5.	X (Stator inductive reactance)	6.96 Ohms
6.	X (Rotor inductive reactance as ref. to stator)	6.96 Ohms
7.	(R Delta)ac (A.C Stator resistance)	11.66 Ohms
8.	J (Moment of Inertia)	0.174 Kg-meter sq.
9.	B(Frictional constant)	0.033 N-m / rad per sec

2. Data for the sliding mode control:

2.1 Electrical parameters

Sl No	Name of the Parameters	Numerical value
1.	Pole	4 Nos
2.	R_s (Stator resistance)	0.087 Ohms
3.	R_r (Rotor resistance)	0.228 Ohms
4.	L_s(Stator inductance)	35.5 mH
5.	L_r (rotor inductance)	35.5 mH
6.	L_m(Mutual inductance)	34.7 mH

2.2 Mechanical parameters:

Sl No	Name of the Parameters	Numerical value
1.	J (Moment of Inertia)	1.662 kg-m.sq
2.	B(Frictional constant)	0.1 Nms

2.3 Controller parameters:

Sl No	Name of the Parameters	Numerical value
1.	K (Constant gain)	-100
2.	B (switching gain)	30

Bibliography:

[1] Abbondanti, A. and. Brennen, M.B. “Variable speed induction motor drives use electronic slip calculator based on motor voltages and currents.” IEEE Trans. Indus. Applications. IA-11 (1975): pp. 483–488.

[2] Ishida, M. Hayashi, K. and Ueda, M. “A speed detection method of squirrel-cage induction motor utilizing rotor slot-harmonics in the air gap and its application to slip frequency control.” Elect. Eng. Japan, Vol. 99, no. 3, (1979): pp. 74-84.

[3] Ishida, M. and Iwata, K. “A new slip frequency detector of an induction motor utilizing rotor slot harmonics.” IEEE Trans. Industry Applications, Vol. IA-20, (may/June 1984): pp. 575-582.

[4] Ishida, M. and Iwata, K. “steady state characteristics of a torque and speed control system of an induction motor utilization rotor slot harmonics for slip frequency algorithm sensing.” IEEE Trans power Electron, Vol. PE-2, (July 1987): pp. 257-263.

[5] Hammerly, B., Tranner, R. and Zwicky, R. “A rotor speed detector for induction machine utilizing rotor slot harmonics and active three phase injection.” in 2nd Proc. Euro. Con$ Power Electronic. App/. induction motor drive. (EPE), Vol. 2, (1987): pp. 599-604.

[6] Beck, M. and Naunin, D “A conception for a sensorless speed control of the squirrel cage induction motor.” in Proc. Power Electron. Appl. Conf., (1985): pp. 3.51-3.55.

[7] Williams, B. W., Goodfellow, J. K. and Green, T. C. “Sensorless speed measurement of inverter driven squirrel cage induction motors.” in Proc. IEEE 4th Int. Con$ Power Electron. Variable Speed Drives, (1987).

[8] Benchaib, A. Edwards, C. “Nonlinear sliding mode control of an induction motor.” Int. J. Adapt. Control Signal Process. Vol. 14, (2000): pp. 201–221.

[9] Bose, B. K. Modern Power Electronics and AC Drives, Prentice Hall, New Jersey, 2001.

[10] Chern, T. L., Chang, J. Tsai, K. L. “Integral variable structure control based adaptive speed estimator and resistance identifier for an induction motor.” Int. J. Control Vol.69, (1998): pp. 31–47.

[11] Huang, S. J., Huang, C. L. and Lin, Y. S. "Sensorless speed identification of vector controlled induction drives via neural network based estimation." Electric Power Syst. Res. Vol. 48, (1998): pp. 1–10.

[12] Kubota, H. Matsuse, K. and Nakano, T. "DSP-based speed adaptive flux observer of induction motor." IEEE Trans. Indus. Applica. Vol.29, (1993): pp.344-348.

[13] Lehonhard, W. Control of Electrical Drives, Springer, Berlin, 1996.

[14] Lin, F. K. Liaw, C. M. "Control of indirect field-oriented induction motor drives considering the effects of dead-time and parameter variations." IEEE Trans. Indus. Electro. Vol. 40, (1993): pp.486-495.

[15] Marino, R., Peresada, S. and Tomei, P. "Adaptive output feedback control of current-fed induction motors with uncertain rotor resistance and load torque." Automatica, vol.34, (1998): pp.617-624.

[16] Ortega, R., Canudas, C. and Seleme, I. S. "Nonlinear control of induction motors: torque tracking with unknown load disturbances." IEEE Trans. Automat. Contr. Vol. 38, (1993): pp. 1675–1680.

[17] Park, M. H., Kim, K. S."Chattering reduction in the position control of induction motor using the sliding mode," IEEE Trans. Power Electron. Vol.6, (1991): pp. 317–325.

[18] Peng, F. Z., Fukao, T. "Robust speed identification for speed-sensorless vector control of induction motors." IEEE Trans. Indus. Applica. Vol.30, (1994): pp. 1234–1240.

[19] Sabanovic, A. Izosimov, D. B. "Application of sliding modes to induction motor control." IEEE Trans. Indus. Applica. IA-17 (1981): pp. 41–49.

[20] Schauder, C. "Adaptive speed identification for vector control of induction motors without rotational transducers." IEEE Trans. Indus. Applica. vol. 28, (1992): pp. 1054–1061.

[21] Utkin, V. I. "Sliding mode control design principles and applications to electric drives." IEEE Trans. Indus. Electro. Vol. 40, (1993): pp. 26– 36.

[22] Vas, P. Vector Control of AC Machines, Oxford Science Publications, Oxford, 1994.

[23] Zamora, J. L., Garcia-Cerrada, A.,and Zazo, A. “Rotor-speed estimator for induction motors using voltage and current measurements.” Contr. Ing. Pract. Vol. 6, (1998): pp. 369–383.

[24] Utkin, V. I. “Variable structure systems with sliding modes.” IEEE Trans. Automat. Contr., vol. 22, no.2, (1997): pp. 212 -222.

[25] Utkin, V. I. “Sliding Modes in Control Optimization.” Springer-Verlag, Berlin, 1992.

[26] Hung, J. Y., Gao, W. and Hung, J. C. “Variable Structure Control: A survey”, IEEE Trans.Ind. Electron. vol. 40, No.1, (1993): pp.2-22.

[27] Levant, A. “Sliding order and sliding accuracy in sliding mode control”, InternationalJournal of Control, vol.58, No.6, (1993): pp.1247-1263.