Performance comparison of different speed estimation techniques in sensorless vector controlled induction motor drives

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Performance comparison of different speed estimation techniques in sensorless vector controlled induction motor drives

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By

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Supervisor's Certificate

This is to certify that the work presented in the dissertation entitled *Performance comparison of different speed estimation techniques in sensorless vector controlled induction motor drives* submitted by *Vikash Kumar*, Roll Number 711EE2102, is a record of original research carried out by him under our supervision and guidance in partial fulfillment of the requirements of the degree of *B.Tech and M.Tech Dual Degree in Electrical Engineering with specialization in "Power Control and Drives"*. Neither this dissertation nor any part of it has been submitted earlier for any degree or diploma to any institute or university in India or abroad.

Kanungobarada Mohanty
Professor

DEDICATED TO MY PARENTS,
TEACHERS AND FRIENDS

Acknowledgment

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ABSTRACT

Field-oriented control and direct torque control are fast becoming necessities of modern industrial setups for induction motor drive control. Induction motors are considered as the beginning part to create any electrical drive system to be subsequently utilized for several industrial requirements. So now a day due to its high application the need to control the performance of the induction motor is gaining importance. In modern control system, IM is analyzed by different mathematical models mainly depending on its applications. Vector control method is suitably applied to induction machine in 3-phase symmetrical or in 2-phase unsymmetrical version. For vector control IM is realized as DC motor having its characteristics. This dissertation work is aimed to give a detailed idea about the speed control and variations in an induction motor through vector control technique thereby showing its advantage over the conventional scalar method of speed control. It also focusses on the speed estimation techniques for sensorless closed loop speed control of an IM relying on the direct field-oriented control technique. The study is completed through simulations with use of MATLAB/Simulink block sets allowing overall representation of the whole control system arrangement of the Induction motor. The performance of different sensorless schemes and comparison between them on several parameters like at low speed, high speed etc. is also provided emphasizing its advantages and disadvantages. The analysis has been carried out on the results obtained by simulations, where secondary effects introduced by the hardware implementations have not been considered. The simulations and the evaluations of different control techniques are executed using parameters of a 50 HP, 60 Hz induction motor which is fed by an inverter.

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LIST OF SYMBOLS & ABBREVIATION

FOC	Field oriented control
DTC	Direct torque control
AC	Alternating Current
ω_r	Rotor mechanical speed(rad/s)
ω_r^*	Reference or command speed (rad/s)
ω_e	Speed of the reference frame or synchronous speed (rad/s)
ω_{sl}	Electrical slip speed (rad/s)
ψ_{dr}, ψ_{qr}	d – axis and q – axis rotor flux linkages
$\psi_{dr}{}^s, \psi_{qr}{}^s$	Component of rotor flux linkage vector in stationary (α-β) reference frame
$\psi_{ds}{}^s, \psi_{qs}{}^s$	Component of stator flux linkage vector in stationary (α-β) reference frame
I_a*, I_b*, I_c*	Reference current for abc phase of inverter
T_d	Electromagnetic torque developed by motor
T_L	Load torque
v_{ds}, v_{qs}	d and q – axis stator voltages
R_s ,R_r	Resistance per phase of stator and rotor referred to stator
Ls,Lr	Stator and rotor per phase inductance referred to stator
L_m	Magnetizing inductance per phase referred to stator
f	Supply frequency in Hz
J	Polar moment of inertia of the motor
X	State vector
Y	Output vector

CHAPTER 1

INTRODUCTION

1.1 Introduction

Induction machine possesses many significant advantages in comparison to several numerous different types of electrical machines. Induction machine is valued so much because they are inexpensive and rugged. They don't require periodic maintenance and don't possess brushes like Direct control (DC) machines and their packed structure makes then impervious to different environment conditions. Therefore, in the present scenario much importance is provided to controlling methods of induction machines but owing to their complex mathematical model and given non-linearity, an IM requires more sharp control technique as in comparison to DC motors. For several decades, open loop v/f control technique which alter itself to uniform volts-per-Hertz proportion of stator voltage has been implemented. But owing to unsatisfactory dynamic results of this given type of control method resulting in saturation effect along with the variation of electrical parameter with temperature has forced to turn to other control techniques. Now a days brisk and swift switching semiconductor power switches available in power electronics and their improvement in power loss have paved the way for powerful digital signal processors on controller technology which are really fast to be applied for advanced control techniques of IM drives.

In the drive applications it has been the dream to run squirrel cage induction motor as a conventional separately excited dc motor and give the desired performance. Despite the latest researches in Vector control/Field oriented control (FOC) there is high room for improvement as there we have not been quite able to bridge the gap between the dream and reality. Although Direct Torque Control has further simplified the Control circuitry to a greater extent. The availability of cheap microprocessors coupled with advancement in Artificial Intelligence (AI), Genetic Algorithm (GA), Fuzzy Logic (FL) etc. have really influenced the design, monitoring and operation of modern industrial electric drives.

A lot of studies in the last decade and so has been put to test to iron out several diverse possible solutions of IM drives control with sole objectives of quick and accurate torque and flux control along with the reduction of the algorithms complexity involved in FOC etc. This chapter presents brief idea about the Necessity of electric drives, Basic control principles and their classification along with evolution of electric drives. A brief introduction to the FOC has been given in present chapter which is the main focus of the dissertation work.

1.2 Need of electric drive

There are numerous issues which are involved in the driving of the induction machine. Previously motors were designed for only driving specific loads i.e. maximum times the generated torque was greater than the required load torque, which as a result provided inefficient driving system as it led to substantial amount of power loss. As we know the in the induction machine the steady state operating region lies in the range of 80% to 100% of the rated speed because of constant supply frequency and fixed number of poles. During starting, the absence of back emf allows the IM to draw a very high amount of inrush current resulting in high power loss and can occasionally cause insulation failure of burning of the motor parts.

The performance of the other electrical appliances in the same line can similarly be affected because it causes high voltage dip in the supply line. The power factor of induction machine is quite low (as low as 0.1) at light load (i.e. open shaft) condition. The main reason is that at light load IM draws highly inductive current. Although with gradual increase in load power factor improves as it increases thereafter. When induction machine runs at less than unity power factor, the current withdrawn is non- sinusoidal in nature which lowers the supply line power quality and degrades performance of several other utilities connected to the same line. Now a day's distribution company penalizes those customers who draws power at lower PF than specified. Also, an induction motor which is operating at lower PF draws currents which are rich in harmonics causing higher rotor loss affecting motor life and thus may lead to pulsating torque resulting in jerky motion affecting the life of motor bearings and may lead to permanent failure of system.

In applications like hoist, crane etc. it often becomes necessary to stop and reverse quickly. It's evident that quality and productivity of the system could be improved by improving the accuracy of stopping and reversing operation. Earlier mechanical brakes were mainly used for all of the stopping and reversing operation of the IM, which were insufficient as they require continuous maintenance. A lot of the other important application of IM like fan, blower, pump etc. often need to control speed too. Consider the figure 1.1 and 1.2 in these type of loads, the torque and power is related to speed as:

$$Torque \approx k_1 (speed)^2 \quad \text{and} \quad Power \approx k_1 (speed)^3$$

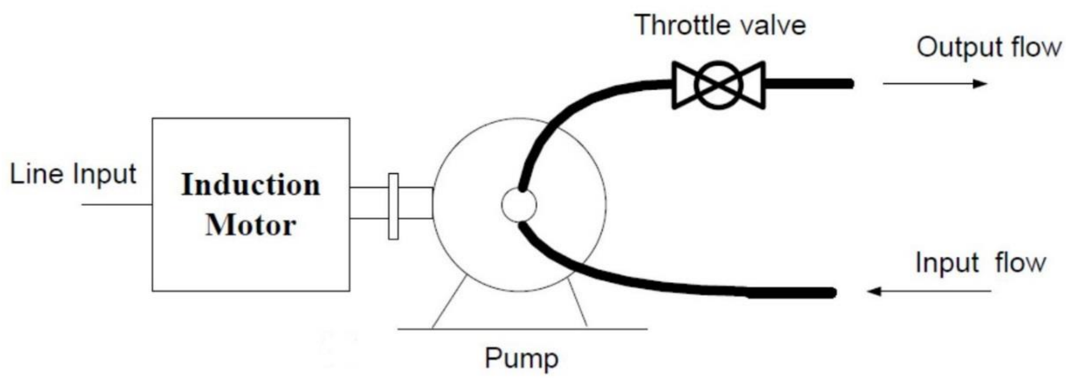

Figure 1-1 Application of induction motor in flow control (IM runs at constant speed and flow is controlled through Throttle)

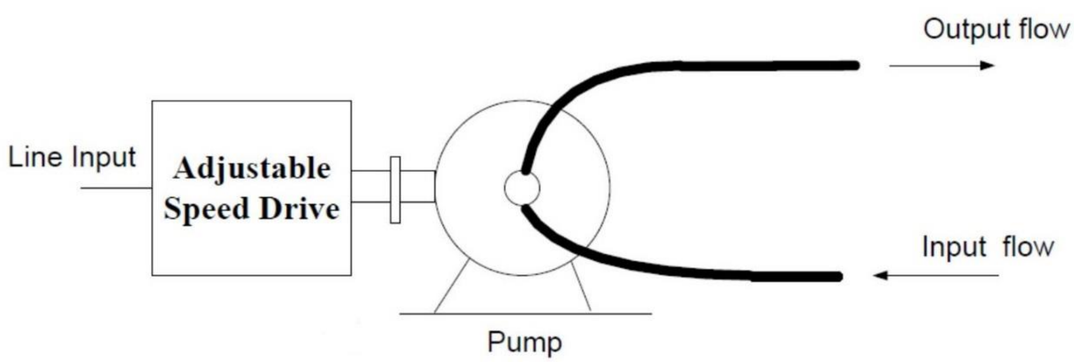

Figure 1-2 Application of induction motor in flow control (IM runs at variable speed to control the flow)

Hence, depending upon the load Variable speed control could provide good amount of energy saving. As can be seen from the above expressions that with reduction of 20% in the operating speed will eventually amount to 50% reduction in the input power to the induction machine. However, it can’t not be possible for those systems in which motor is connected directly to the supply in place of being connected to line through a power converter. Thus Fig 1.2 can be understood as an appreciable power saving scheme.

Thus inferring from the previously mentioned points, the need for intelligent motor control can be understood. Hench with advancement in solid state devices (SSD) technology like IGBT, BJT, SCR, MOSFET etc. and technologies like IC fabrication technology, microcontrollers, microprocessors, AI methods which are capable of executing real-time complex algorithms it is now possible to realize Intelligent drive control.

1.3 Control principle

The speed of Induction machine can be demonstrated in terms of slip, frequency and pole numbers as:

$$\omega_m = \omega_s(1-s) = \frac{4\pi f_s}{P}(1-s)$$

The speed of the IM rotor can be varied or controlled by changing slip, frequency and the number of poles. Changing the pole numbers is done by pole amplitude modulated motors which is mainly achieved with a change in winding connections which requires relays and circuit. The variation of applied voltage or insertion of external resistors in stator or rotor affects the slip speed control. The inverter-driven induction motor with the help of variable frequency operation achieves speed control. To keep the air gap flux constant not only the frequency but the applied voltage also needs to be varied which further doesn’t let it saturate. A number of speed control strategies have been implemented by modern age variable frequency drive (VFD) which remains the main aim of this dissertation and will be discussed in further sections.

1.4 Evolution of drives

The motivation behind the evolution of drives from DC to various form of AC drives has been driven by continuous need of reliability, simplicity, performance, ruggedness, cost and availability. In practice, these needs have often resulted in covering mutually specific goals. New research now a day often bring out techniques with certain advantages and disadvantages. Figure 1.3 shows evolution of basic drives as for assessment purposes this viewpoint is simple and useful.

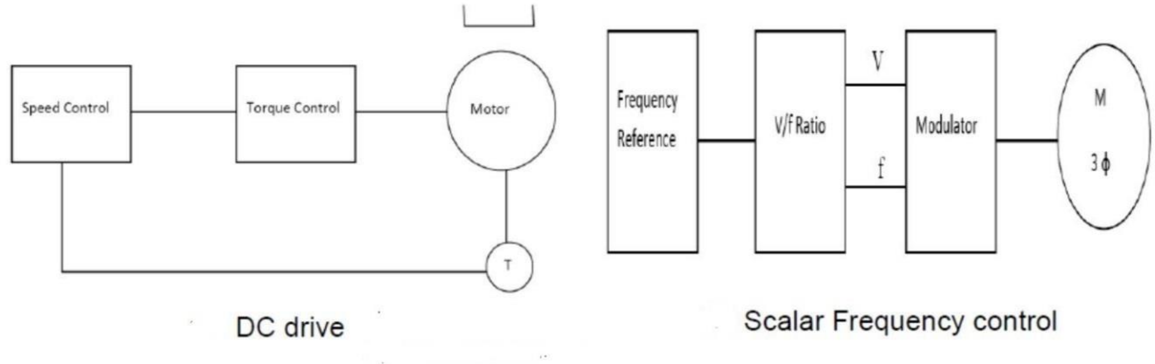

Figure 1-3 Evolution of drives

1.4.1 DC Drive

One of the significant advantage of DC drive is that both direct flux and direct torque control are available. In a dc motor flux is controlled by field control and armature current controls the motor torque. A PWM inverter which is connected at its terminal provides the required flux and voltage to control the torque and speed. The poor accuracy control results due to absence of encoder and field orientation. Even if it's simpler in construction and possesses low cost but it provides very poor performance and less accuracy.

1.4.2 Scalar frequency Control

A block diagram for the scalar control has been shown in figure 1.4. Although, this type of drives uses all the advantages of IM but torque and flux neither directly nor indirectly could be controlled. A PWM invertor connected at its terminal provides the require voltage and flux to control speed and torque. The accuracy of control is very poor due to absence of encoder and field orientation. Although it has simple construction and low cost but it provides very less accuracy and poor performance.

1.4.3 Field Oriented Control (FOC)

The most familiar control scheme used for induction motor drive control is the field oriented control (FOC). This control comprises of stator currents being controlled and represented by a vector. FOC control scheme is mainly based on the principles which transform a 3-phase time as well as speed dependent system into a 2 coordinate system (d and q) which are time invariant system. The major idea behind the vector control of IM is to have an electric drive which possesses superior performance than separately excited DC motor. The squirrel cage induction motor with FOC shows a good level of dynamic performance and stability of the system is a long term phenomenon through the closed loop control of this drive. FOC is separately called ‘Independent or Decoupled control’ wherein flux and torque current vectors are controlled. In the field control stator current are expressed in the rotating coordinate system and are further resolved into two components which produce the torque and flux in the motor and are orthogonal in nature. This arrangement is similar to the DC motor in which flux and torque are controlled independently. To control flux and torque (thereby speed) independently in the induction machine, there is necessity to control the phase and magnitude of three phase stator currents with the help of fast inverter like CC-VSI (current control voltage source inverter). These control algorithms are highly complex and involved and to realize this aim fast acting microcontrollers and other processors are used.

The main disadvantage of Field oriented control is compulsory use of ‘Encoder”. The process of voltage and frequency reference takes good amount of time, thereby limiting the ability to achieve rapid flux and torque. In the FOC for rotor flux regulation, flux position is also required, which is either sensed or estimated**.** Speed and flux estimation are the main problems of the field orientation in the recent years. The induction machine drive without the speed sensors give the advantage of low cost and high reliability. Estimation of magnitude and spatial orientation of flux in the stator or rotor is also necessary for such type of drives.

Rotor flux orientation is mainly categorized in two parts which are the direct field orientation relying on measurement and direct measurement of rotor magnitude and angle and other being the indirect orientation in which slip speed relation is utilized. The scenario of Indirect field orientation is a feedforward approach and is very much sensitive to parameters particularly to the rotor time constant. Due to this several parameter adapting strategies have been developed [8].

Direct field oriented control (DFOC) utilizes the flux angle Θ_e which is obtained or calculated by sensing the air gap flux with the help of flux sensing coils. This arrangement adds to the complexity and the cost of the drive system. So for the last decades many different type of algorithms has been proposed to avoid the use of these flux sensors on the induction machine drive systems to estimate both the rotor flux vector and rotor shaft speed. The most recent trend can be directed toward the use of speed sensors and using algorithms based on the terminal quantities of the machine for the estimation of the fluxes.

One of the preferred flux and speed estimation technique (algorithm) is saliency based with fundamental or high frequency signal injection. The advantage that this saliency based method offers is that the saliency is not sensitive to actual motor parameters. Although this method also suffers of insufficient performance at low and zero speed level. Also when it is subjected to high frequency signal injection this method has shown to cause torque ripples, audible noise and vibration.

1.5 Literature Review

Now a day for industrial applications induction machines associated with high performance drives are used. The history of inductions machines as well as uses have been quite broad. It finds numerous use in commercial, industrial and domestic applications as adjustable speed drives. In present section a literature review has been presented as a means to show the exploration and work of several researchers to make Induction machine very precise, quick and of high performance. A lot of work has been done to develop the technique and to reach the best efficiency of induction motor drive (IMD), many new techniques of control has been developed in the last few years.

Almost 40 years ago, very first paper, in 1971 F. Blaschke [1] presented field oriented control (FOC) of induction motor. A lot of work has been done after that time to develop the technique as now a days FOC drives are an industrial reality and are readily available in the market manufactured by different companies with desired performance and other requirements.

In recent decades much effort has been made to get rid of conventional speed transducer from its use in adjustable induction motor drives. By making the use of voltage and current this aim has been fulfilled. A very good effort was made by Brennen and Abbondanti in 1975 in form of an analog slip calculator which was based on the processing of the motor quantities such as current and voltage [2].In 1979 the same work on sensorless technique was highlighted by Ishida et al [3] who made the use of rotor slot harmonic voltages in slip frequency control.

William L. Erdman and R. G. Hoft in 1990 described airgap and stator field orientation (FOC) methods as an alternative to the familiar rotor orientation process. The advantages of stator and airgap decoupling lie principally in estimating or measuring the corresponding flux. The airgap flux is directly measurable, and the stator flux is closely related to stator terminal quantities. The rotor flux is the most difficult to estimate and for this reason, stator and/or airgap flux implementations may be more robust under parameter and environmental disturbances [4].

Several years later, in 2003 S. Xepapas presented a way for estimation of rotor speed, rotor flux and its angular position as well as the motor torque from the measured terminal currents and voltages. It used nonlinear sliding control technique for control at both low and high speeds [5].

Thereafter Ahmad Razani Haron, Nik Rumzi Nik Idris**,** in 2006 discussed the performance of the rotor flux based MRAS (RF-MRAS) and back e.m.f based MRAS (BEMF-MRAS) for estimating the rotor speed. Both schemes use the stator equation and rotor equation as the reference model and the adjustable model [6].

M. S. Zaky, M. Khater, H. Yasin, and S. S. Shokralla in 2008 discussed about study of the different speed estimation techniques and their corresponding merits and demerits as well as their feasibility for estimating the rotor speed. Many factors remain important to evaluate the effectiveness of the different schemes proposed for speed estimation. Among them are steady state error, dynamic behavior, noise sensitivity, low speed operation, parameter sensitivity, complexity, and computation time [7].

The proposed method in [14] estimates the velocity without the assumption that the speed varies slowly compares to electrical variables studied on non-linear method. So two estimator was constructed: main flux estimator and complementary flux estimator. The main flux estimator did not guarantee convergence for all the operating conditions. So start up complementary estimator is used in such operating conditions. In this method significant sensitivity to parameter uncertainty is observed.

1.6 Objective

The main aim of the project is to compare the performance of different sensorless vector controlled IM drives and provide a detailed study regarding different speed estimation methods and then their corresponding merits and demerits along with their feasibility to estimate speed of the rotor. The simulation study of results forms the basis of the conclusion

1.7 Dissertation Outline

The dissertation has been organized as described below:

Chapter 1 deals with the brief idea about the introduction, need, principle, classification and evolution of the variable frequency electric drives as well as the literature review and a short and brief introduction of the FOC.

Chapter 2 includes the induction machine modelling and dynamical model of machine. This can be implemented to materialize various other equivalent models in difference frame of reference. Vector control technique of IM along with the direct and indirect control technique has been explained.

Chapter 3 is devoted to the various speed estimation methods for sensorless direct field oriented control of the Induction machine. MRAS speed estimator, open loop speed estimator along with adaptive observer has been described.

Chapter 4 demonstrates the simulations of the vector control techniques along with various sensorless speed estimation techniques for analyzing the comparison between there performance.

Chapter 5 shows the performance comparison of these techniques and the preference on these techniques in various conditions. There is include proposal for further research work.

CHAPTER 2

MODELLING & FIELD ORIENTED CONTROL OF INDUCTION MOTOR

2.1 Introduction

Induction machine with rotor cage configuration have always been the widely used machine at a particular fixed speed because of its efficiency, simplicity, reliability, ruggedness, compactness, low cost, and having economical and volume manufacturing advantage. Although in the recent year several developments regarding the field of varying speed drives have opened up possibility of application of variable speed induction motor drives at a larger scale. In comparison to separately excited DC motor Induction machine demands the use of extra complex control techniques because of its structure having nonlinear dynamic structure with strong coupling.

Now any power electronics drive e.g. Induction motor drive, controllers are needed to control the system. This scenario requires mathematical modelling of the drive because of multivariable nature of the systems and its higher order nonlinearity. Thus design and development of the power electronics drive systems could be done by suitable mathematical modelling of the plant. The induction motor modelling is mostly performed by a 3-phase (a-b-c) to synchronously rotating (d-q) transformation with rotor flux linkages and stator current as the stator variables neglecting saturation.

2.2 Induction motor modelling

2.2.1 Steady state modelling

The steady state equivalent circuit of an IM is very much identical to a transformer. With simple mathematical manipulation it is integrated into the circuit despite rotor currents being at slip frequency. An equivalent circuit (stator side referred) of induction motor is shown as in figure 2.1

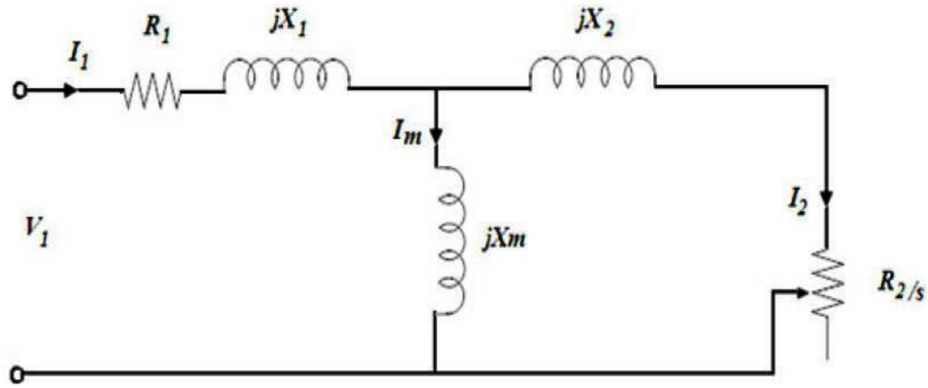

Figure 2-1Equivalent circuit of IM in steady state

Steady state performance equation:

In this scenario the no-load current is viewed as the sum of the core-loss components and the magnetizing component of the current and accordingly is written as:

$$I_o = I_m + I_c \tag{2.1}$$

The magnetizing current in terms of the magnetizing reactance and air gap voltage is written as:

$$I_m = \frac{E_1}{jX_m} \tag{2.2}$$

Here, Xm is the magnetizing reactance and E1 is the airgap voltage

Similarly, core loss component of the stator current is written as:

$$I_c = \frac{E_1}{R_c} \tag{2.3}$$

Where, Rc is the core loss accounting resistance

The rotor phase current is given by:

$$I_r \approx \frac{E_1}{\frac{R_r}{s} + jX_{lr}} \tag{2.4}$$

Where, Ir is phase current of rotor

Then the phase current of stator is:

$$I_{as} = I_r + I_o \tag{2.5}$$

In terms of stator parameters, stator current and induced emf the applied stator voltage is shown as the sum of the induced emf with the stator impedance voltage drop which is given as:

$$V_{as} = E_1 + (R_s + jX_s)I_{as} \tag{2.6}$$

Where Vas is the stator phase voltage

The main variables in the machine comprise of air gap power, torque, mechanical and shaft output. The difference among the total input power with respect to copper losses in stator is expressed as real power transmitted from input to the air gap which is given as:

$$P_a = P_i - 3I_{as}{}^2 R_s \tag{2.7}$$

Neglecting the core losses, we have

$$P_a = 3I_r{}^2 \frac{R_r}{s} \tag{2.8}$$

Which could be written alternatively as

$$P_a = 3I_r{}^2 R_r + 3I_r{}^2 R_r \frac{(1-s)}{s} \tag{2.9}$$

The Power Output Pm (mechanical) is given as:

$$P_m = 3I_r{}^2 R_r \frac{(1-s)}{s} \tag{2.10}$$

Alternatively, it can also be given in form of electromagnetic torque and rotor speed as:

$$P_m = T_e \omega_m \tag{2.11}$$

Where Te is the internal or electromagnetic torque

$$T_e = \frac{3I_r{}^2 R_r (1-s)}{s\omega_m} \tag{2.12}$$

Substituting for the motor speed in terms of the slip and stator frequency,

$$\omega_m = \frac{\omega_r}{P/2} = \frac{\omega_s (1-s)}{P/2} \tag{2.13}$$

The electromagnetic or air gap torque is obtained as:

$$T_e = 3(P/2) \frac{I_r{}^2 R_r}{s\omega_s} \tag{2.14}$$

The net output power of the machine Ps:

$$P_s = P_m - P_{fw} \tag{2.15}$$

The friction and the windage losses are two different losses as they are proportional to the speed and the square of speed respectively. Therefore, for evaluation of variable speed performance of IM they have to be represented as function of speed. There are also other type of losses such as stray load losses which are caused due to stray magnetic fields in the machine.

2.2.2 Dynamic modelling of Induction machines

The above presented steady-state model and the corresponding equivalent circuit are important for studying the steady state performance of the machine. So basically its nota applied for transient operation where speed or torque variation is there. Thus the need is to find out dynamics of the variable speed drives which is fed by converter to evaluate the capability of the converter switches which are applied to a given motor and also their interaction which determines the excursions of the torque and current in motor. The instantaneous effect of the current and voltages along with the torque and frequency variations are considered in dynamic modelling. In the following section the dynamic modelling of the two phase induction motor has been by transforming 3 phase to two phase in direct as well as quadrature axes.

In 3-ϕ IM the voltage equations in synchronously rotating reference frame are:

$$V_{ds} = \frac{d\psi_{ds}}{dt} + R_s i_{ds} - \omega_e \psi_{qs} \tag{2.16}$$

$$V_{qs} = \frac{d\psi_{qs}}{dt} + \omega_e \psi_{ds} + R_s i_{qs} \tag{2.17}$$

$$V_{dr} = \frac{d\psi_{dr}}{dt} - (\omega_e - \omega_r)\psi_{qr} + R_r i_{dr} \tag{2.18}$$

$$V_{qr} = \frac{d\psi_{qr}}{dt} + R_r i_{qr} + (\omega_e - \omega_r)\psi_{dr} \tag{2.19}$$

The electromagnetic torque which develops is given by

$$T_e = \frac{3P}{4}(\psi_{dr} i_{qs} - \psi_{qr} i_{ds}) \tag{2.20}$$

The torque balance equation is written as:

$$T_e = T_l + \beta\omega_r + J\frac{d\omega_r}{dt} \tag{2.21}$$

Where voltage (V) and current (i) are transferred to synchronously rotating reference frame and then corresponding subscripts of d and q axis for the stator as well as rotor are d_s, q_s and d_r, q_r so that R_s is the per phase resistance of stator and R_r is the per phase rotor resistance of the motor whereas ω_e is the reference frame speed , ω_r is the rotor's mechanical speed , ψ is flux linkage. J is the MOI (moment of inertia), P is number of pole pairs and β is viscous friction coefficient. T_e and T_l are the torque developed and the load torque.

The voltage and electromagnetic torque equations can also be represented in matrix form as given below:

$$\begin{pmatrix} V_{ds} \\ V_{qs} \end{pmatrix} = \begin{pmatrix} R_s & 0 \\ 0 & R_s \end{pmatrix}\begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} 0 & -\omega_e \\ \omega_e & 0 \end{pmatrix}\begin{pmatrix} \psi_{ds} \\ \psi_{qs} \end{pmatrix} + \frac{d}{dt}\begin{pmatrix} \psi_{ds} \\ \psi_{qs} \end{pmatrix} \tag{2.22}$$

$$\begin{pmatrix} V_{dr} \\ V_{qr} \end{pmatrix} = \frac{d}{dt}\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} + \begin{pmatrix} R_r & 0 \\ 0 & R_r \end{pmatrix}\begin{pmatrix} i_{dr} \\ i_{qr} \end{pmatrix} + \begin{pmatrix} 0 & -(\omega_e - p\omega_r) \\ (\omega_e - p\omega_r) & 0 \end{pmatrix}\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} \tag{2.23}$$

$$T_e = \frac{3P}{4}[\psi_{dr} \quad \psi_{qr}]\begin{pmatrix} i_{qs} \\ -i_{ds} \end{pmatrix} \tag{2.24}$$

As the rotor windings are short circuited hence the voltages will be $V_{dr} = V_{qr} = 0$

Ignoring the iron loss, flux equations are also displayed in the matrix form

$$\begin{pmatrix} \psi_{ds} \\ \psi_{qs} \end{pmatrix} = \begin{pmatrix} L_s & 0 \\ 0 & L_s \end{pmatrix}\begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} L_m & 0 \\ 0 & L_m \end{pmatrix}\begin{pmatrix} i_{dr} \\ i_{qr} \end{pmatrix} \tag{2.25}$$

$$\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} = \begin{pmatrix} L_m & 0 \\ 0 & L_m \end{pmatrix}\begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} L_r & 0 \\ 0 & L_r \end{pmatrix}\begin{pmatrix} i_{dr} \\ i_{qr} \end{pmatrix} \tag{2.26}$$

Where L_s, L_r are self inductances of stator and rotor respectively, whereas L_m is basically mutual inductance

From equation (2.26) we can write

$$\begin{pmatrix} i_{dr} \\ i_{qr} \end{pmatrix} = \begin{pmatrix} \frac{1}{L_r} & 0 \\ 0 & \frac{1}{L_r} \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} - \begin{pmatrix} \frac{L_m}{L_r} & 0 \\ 0 & \frac{L_m}{L_r} \end{pmatrix} \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} \tag{2.27}$$

Substituting equation (2.27) in (2.25)

$$\begin{pmatrix} \psi_{ds} \\ \psi_{qs} \end{pmatrix} = \begin{pmatrix} \sigma L_s & 0 \\ 0 & \sigma L_s \end{pmatrix} \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} \frac{L_m}{L_r} & 0 \\ 0 & \frac{L_m}{L_r} \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} \tag{2.28}$$

Where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ = leakage coefficient

Taking the rotor voltages zero and substituting (2.27) in (2.23) we get

$$\frac{d}{dt}\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} = \begin{pmatrix} \frac{R_r L_m}{L_r} & 0 \\ 0 & \frac{R_r L_m}{L_r} \end{pmatrix} \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} -\frac{R_r}{L_r} & 0 \\ 0 & -\frac{R_r}{L_r} \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} + \begin{pmatrix} 0 & -(\omega_e - p\omega_r) \\ (\omega_e - p\omega_r) & 0 \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix}$$

$$\frac{d}{dt}\begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} = \begin{pmatrix} a_5 & 0 \\ 0 & a_5 \end{pmatrix} \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} -a_4 & \omega_{sl} \\ \omega_{sl} & -a_4 \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} \tag{2.29}$$

Where $a_5 = \frac{R_r L_m}{L_r}, a_4 = \frac{R_r}{L_r}$ and $\omega_{sl} = -p\omega_r + \omega_e$

Taking equations (2.28), (2.22) and (2.29) we can find

$$\frac{d}{dt}\begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} = \begin{pmatrix} -a_1 & \omega_e \\ -\omega_e & -a_1 \end{pmatrix} \begin{pmatrix} i_{ds} \\ i_{qs} \end{pmatrix} + \begin{pmatrix} a_2 & pa_3\omega_r \\ -pa_3\omega_r & a_2 \end{pmatrix} \begin{pmatrix} \psi_{dr} \\ \psi_{qr} \end{pmatrix} + \begin{pmatrix} c & 0 \\ 0 & c \end{pmatrix} \begin{pmatrix} v_{ds} \\ v_{qs} \end{pmatrix} \tag{2.30}$$

Where

$$a_1 = \frac{1}{\sigma L_s}(R_s + R_r \frac{L_m^2}{L_r^2})$$

$$a_2 = \frac{1}{\sigma L_s} R_r \frac{L_m^2}{L_r^2}$$

$$a_3 = \frac{L_m}{\sigma L_s L_r}, \qquad c = \frac{1}{\sigma L_s}$$

Relating equations (2.29) and (2.30) we get the state space model of induction motor as:

$$\frac{d}{dt}\begin{pmatrix} i_{ds} \\ i_{qs} \\ \psi_{dr} \\ \psi_{qr} \end{pmatrix} = \begin{pmatrix} -a_1 & \omega_e & a_2 & pa_3\omega_r \\ -\omega_e & -a_1 & -pa_3\omega_r & a_2 \\ a_5 & 0 & -a_4 & \omega_{sl} \\ 0 & a_5 & -\omega_{sl} & -a_4 \end{pmatrix}\begin{pmatrix} i_{ds} \\ i_{qs} \\ \psi_{dr} \\ \psi_{qr} \end{pmatrix} + \begin{pmatrix} c & 0 \\ 0 & c \\ 0 & 0 \\ 0 & 0 \end{pmatrix}\begin{pmatrix} v_{ds} \\ v_{qs} \end{pmatrix} \tag{2.31}$$

The electromagnetic torque can similarly be written in state space variables as

$$T_e = \frac{3P}{4}[\psi_{dr} \quad \psi_{qr}]\begin{pmatrix} i_{qs} \\ -i_{ds} \end{pmatrix} \tag{2.32}$$

2.3 Induction machine control

The induction machine drives speed control requires the use of controllers which can be traced to two major types of control: scalar control and vector control. Scalar control includes several way of speed control techniques one of most used being the volts per hertz i.e. v/f control which are of low cost conventionally. In this control, the voltage as well as frequency magnitude are kept in regular proportion. But v/f control performances has been not been found satisfactory because the rate of voltage change as well as frequency is needed to be kept low. Any sudden change in deceleration or acceleration of the frequency or voltage lead to a transient current change, which later leads to severe problems. Although many concrete works and manipulations have been made to control performance of v/f method, still not any of those improvements could provide an effective v/f controlled drive system and so this limitation allowed the DC motor undisputed & permanent choice in case of various variable speed requirements. This scenario started changing when the concept regarding field orientation theory was suggested by Hasse and Blaschke. The complications are much more in field oriented control than the control of DC motor. Even as the most successful class of the controllers which are very popular use vector control method as it regulates magnitude and the phase of AC excitation. So the given particular method results in particular space orientation of field and torque which are orthogonal in nature and are thus called Field Oriented Control (FOC).

2.4 Field Oriented Control

The vector control technique helps induction machine to be driven with dynamic performances of direct control motor. This particular technique decouples the stator current components in which one provides the air gap flux and another one producing torque so that the control characteristics are linearized. Since they are orthogonal the given two fluxes doesn’t show any particular interaction to each other. Thus after adjustment of field current we can control flux of the DC machine and then the machine torque can be regulated independently by armature current adjustments. Since any alternating current machine is complicated due to interaction between rotor and the stator fields and their orientation are not held at 90° and they vary with changing operating conditions. Thus suffice to say DC machine like performance can be obtained by keeping a orthogonal as well as fixed orientation among the fields in any given AC machine by orientating the stator current with regard to rotor flux so as to obtain independently controlled torque and flux. Such a method is called Vector Control or Field Oriented Control.

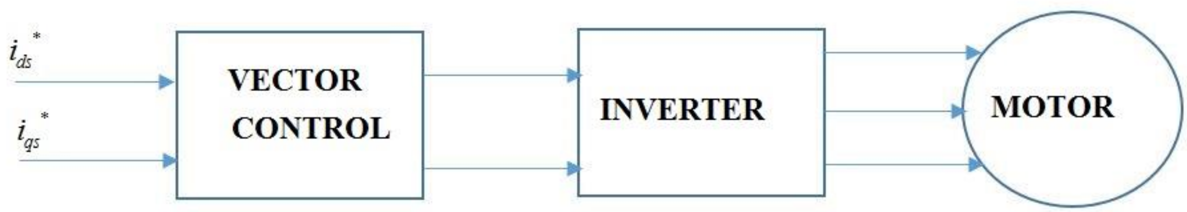

Figure 2-2 Vector Control Analogy

The two components essential for vector control are i_{ds} corresponding to d-axis armature current and i_{qs} corresponding to q-axis field current of a dc motor which is separately excited. The rotor flux linkage is aligned with the d-axis which is the elementary condition for vector oriented control.

At this condition:

$$\left.\begin{array}{l}\psi_{qr} = 0 \\ \psi_{dr} = \psi_r\end{array}\right\} \begin{array}{l}\textit{Rotor flux} \\ \textit{orientation}\end{array} \qquad \left.\begin{array}{l}\psi_{qs} = 0 \\ \psi_{ds} = \psi_r\end{array}\right\} \begin{array}{l}\textit{stator flux} \\ \textit{orientation}\end{array}$$

Considering the above scenario, the block diagram for vector control can be shown as follows:

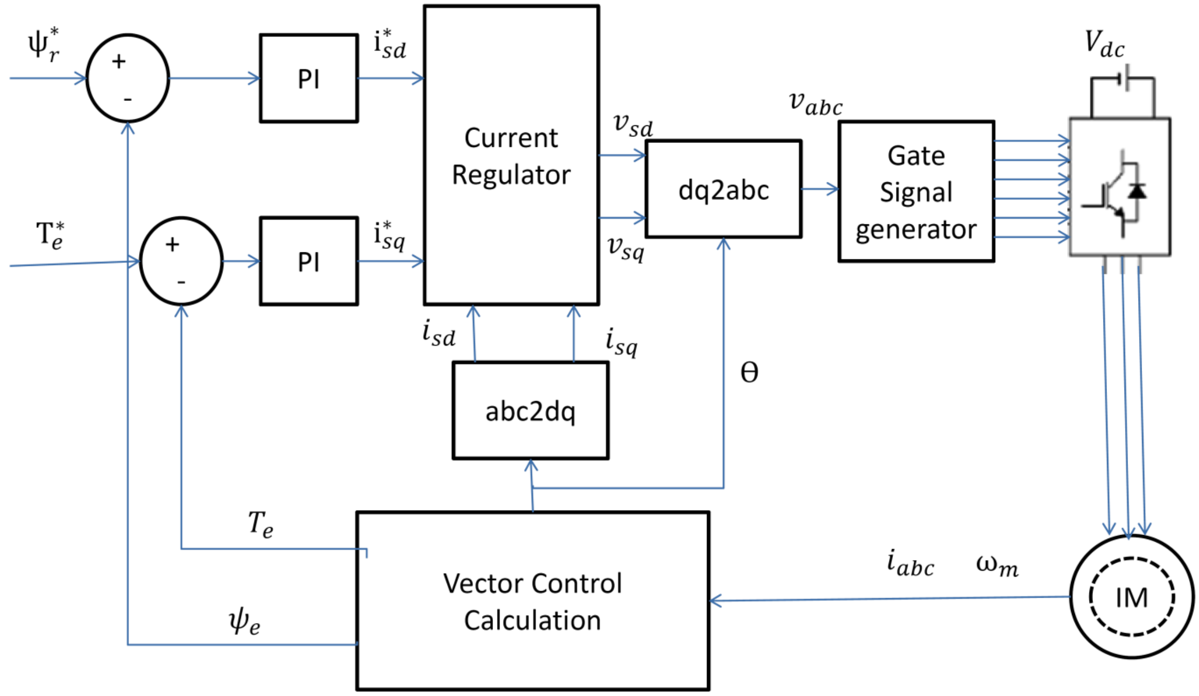

Figure 2-3 Block Diagram of FOC

The angle Ө can be obtained by indirect field orientation control (IFOC) or otherwise also by direct field orientation control (DFOC). Henceforth controllers which are applied in this scenario insures that it can obtain the decoupled torque and flux control and thus are known as field oriented controller.

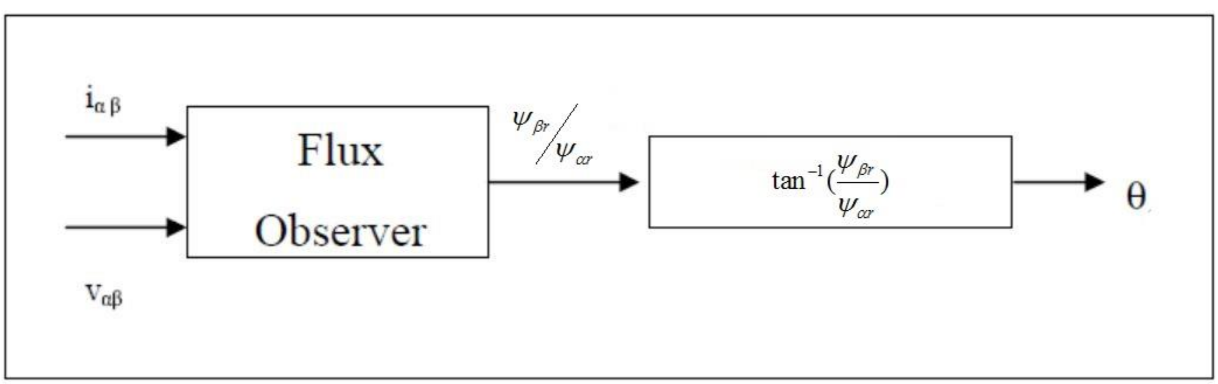

Figure 2-4 Direct field oriented drive system

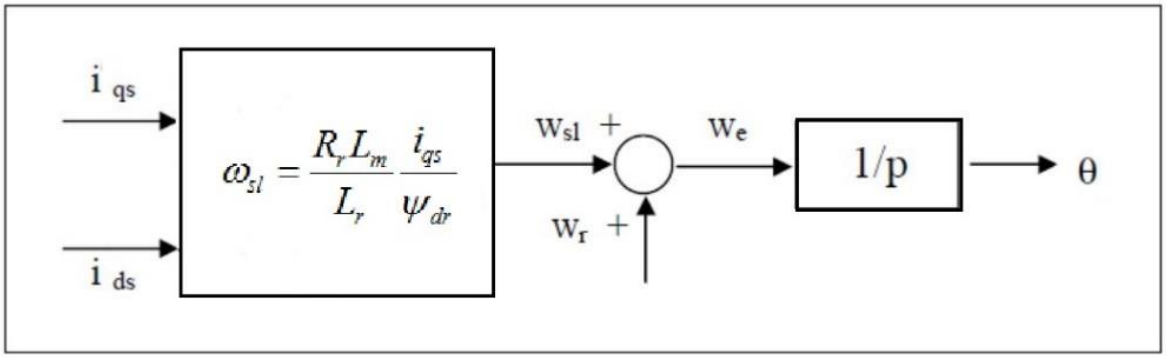

Figure 2-5 Indirect field oriented drive system

Both different techniques of vector control named direct and indirect field oriented control depend upon the basic flux acquisition method.

2.4.1 Direct Vector Control

The use of particular flux coils and sensors can be discontinued by estimation of the rotor flux from the stator terminal quantities which are voltage and currents [9]. Direct method includes the measurement of airgap flux directly incorporating the help of search coils. Sensors or sensors, taped stator windings or it can be done by measurement from machine terminal variables such as stator current, voltage and speed. Now the impossible scenario of direct sensing procedure of rotor flux directs it to be obtained from the directly sensed airgap flux. Rotor flux angle is measured directly through flux measurement and estimation. The very serious drawback of this direct scheme arises is at low speed when the IR drop of the machine is dominating which is difficult to neglect and it also the requirement of integration of signal makes the measurement of airgap flux difficult. This thinking procedure requires the knowledge of stator as well as rotor leakage inductances, magnetizing inductances and there is also need of stator resistance. This procedure is normally referred as Voltage Model Flux Observer (VMFO). Hence in stationary frame the stator flux can be obtained by the equations:

$$\dot{\psi}_{qs}^{\,s} = V_{qs}^{\,s} - r_s i_{qs}^{\,s} \tag{2.33}$$

$$\dot{\psi}_{ds}^{\,s} = V_{ds}^{\,s} - r_s i_{ds}^{\,s} \tag{2.34}$$

Also the rotor flux can be obtained as:

$$\psi_{dr}{}^{s} = \frac{L_r}{L_m}(\psi_{ds}{}^{s} - L_\sigma i_{ds}{}^{s}) \tag{2.35}$$

$$\psi_{qr}{}^{s} = \frac{L_r}{L_m}(\psi_{qs}{}^{s} - L_\sigma i_{qs}{}^{s}) \tag{2.36}$$

Where $L_\sigma = (L_s - L_m{}^2 / L_r)$ is the transient leakage reactance

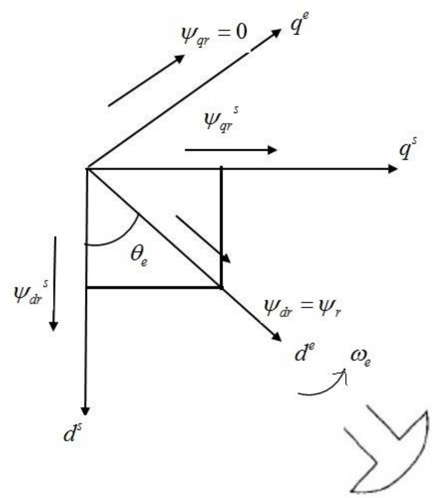

Figure 2-6 Direct vector control phasor

Voltage Model Flux Observer uses the measured current and voltages of the stator and requires a pure integration without the use of any feedback. Thus it is found harder to be implemented for frequencies of low excitation because of offset also and initial condition problems. Thus in actual practice low pass filter is also used often to give stability due to the lack of filter which is necessary for convergence. Accuracy of this model is totally independent to rotor resistance but is found as being most sensitive with respect to stator resistance particularly at low velocities. At high speed the stator resistance (IR) drop sensitivity to speed voltage is low which reduces sensitivity to stator resistance. The sensitivity to the parameters study depicts that leakage inductance very much affects the performance of the system in regard to dynamic response, stability and utilization of inverter and the machine.

To overcome these difficulties caused by leakage inductance changes and also due to stator resistance mainly at low speed as an alternative approach Current Model Flux Observer (CMFO) has been proposed.

This flux model can be measured or estimated as:

$$\dot{\psi}_{dr}{}^{s} = -\frac{1}{T_r}\psi_{dr}{}^{s} - \omega_r \psi_{qr}{}^{s} + \frac{L_m}{T_r} i_{ds}{}^{s} \tag{2.37}$$

$$\dot{\psi}_{qr}{}^{s} = -\frac{1}{T_r}\psi_{qr}{}^{s} - \omega_r \psi_{dr}{}^{s} + \frac{L_m}{T_r} i_{qs}{}^{s} \tag{2.38}$$

This current model uses the stator current which is measured and speed of the rotor. The dependency on speed of the current model is considered as disadvantage as this implies that although use of the estimated flux obviously eliminates the flux sensor, the requirement of position sensor is still there. Furthermore, even at low or zero speed operation rotor flux magnitude response is sensitive to particularly to the rotor resistance, the phase angle is insensitive to all the parameters. Near rated slip, both the quantities are particularly sensitive to magnetizing inductance and resistance of the rotor. In the whole speed range the rotor leakage resistance totally has no effect on accuracy.

2.4.2 Indirect Field Orientation Control

The indirect field orientation of the IM, the rotor vector ψ_{dr} instantaneous speed is same as that of the synchronous speed ω_e and as that of d-q coordinate system where d axis is directly locked on the rotor flux vector as has been case for rotor flux vector orientation. This scenario helps to control the flux by magnetizing current i_{ds} through alignment of total flux with the d axis and also aligning the other torque generating current component with q axis. Henceforth after decoupling of the torque producing and the rotor flux currents the torque can be easily regulated with the current i_{qs}.

The rotor angle is found out with use of these described equations:

$$0 = \frac{R_r}{L_r}\psi_{dr} + \frac{d}{dt}\psi_{dr} - \frac{R_r * L_m}{L_r} i_{ds}$$

$$\psi_{dr} = \frac{L_m}{1 + T_r s} i_{ds}$$

$$0 = R_r i_{qr} + (\omega_e - \omega_r)\psi_{dr} \qquad i_{qr} = -\frac{L_m}{L_r} i_{qs}$$

$$\omega_{sl} = \frac{R_r * L_m}{L_r * \psi_{dr}} i_{qs} \tag{2.39}$$

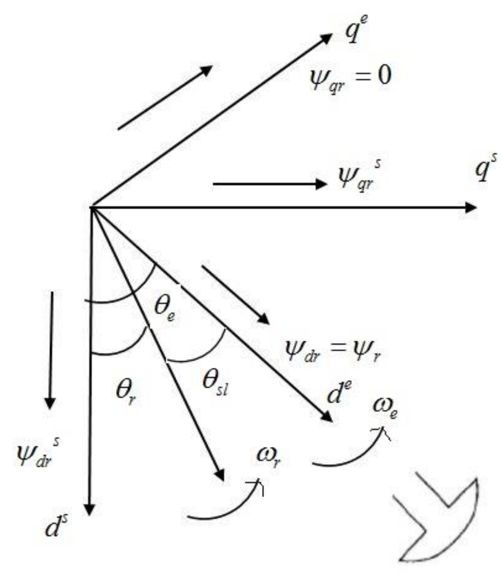

Figure 2-7Indirect vector control phasor

$$\theta_e = \int \omega_e dt = \int (\omega_r + \omega_{sl}) dt = \theta_r + \theta_{sl} \tag{2.40}$$

The Indirect field oriented control is a feedforward control which is basically open loop control and in this method the slip frequency is fed forward which very much guarantees the field orientation. This particular controlling operation is sensitive to rotor open circuit time constant τ_r which has to be known for achieving a decoupled control of flux and torque components by control of the currents. When τ_r is not set correctly, the machine is described as detuned and its performance becomes very sluggish because of decoupled control loss. Hence, the measurement of the time constant of the rotor, its effects on the system performance along with its adaptive tuning with the variations which mainly result during the machine operation have been extensively studied and monitored. The rotor time constant is also affected by the changes in temperature which affects the torque capability of the induction motor. This effect of detuning becomes of high severity mainly in the field weakening region. It also gives results as steady state error and transient oscillation in the rotor torque and flux.

Thus the concept of indirect field orientation was developed and studied by many researchers in last two-three decades. The indirect control is the most original and dominant choice for the rotor flux orientation. Also the IFO control can be simulated or implemented using stator orientation or the air gap flux as well. In air gap orientation the flux as well as slip relations are coupled equations. In this case d-axis current does not control the flux independently as done earlier in orientation along rotor flux. It has been found that in case of orientation along air gap flux, the maximum produced torque is 20% lesser in comparison to the other methods. Particularly in the scenario of orientation along stator flux, the transient reactance is a coupling factor which vary regularly with the operating conditions of the machine [19]. In addition Nasar also showed that in these three types rotor orientation possess linear torque curve and thus is a very obvious choice for indirect oriented control.

2.5 Conclusion

The modelling of IM with the help of mathematical equations has been discussed with equations. Taking stator current as well as rotor flux components as some variables motor differential equations has been expressed in stationary as well as synchronously rotating(d-q) reference frame. Its steady state operations have also been shown in brief. Basic vector control of induction motor which includes direct and indirect vector control have been detailed. The speed measured in these control has been done by sensors which has to be eliminated for the cost and efficiency purpose the idea which next chapter carries forward.

CHAPTER 3

SENSORLESS VECTOR CONTROL OF INDUCTION MOTOR

3.1 Introduction

The main aim of this chapter is select a configuration based on the review of different methods for field orientation of induction motor which is suitable for high performance sensorless drive. As discussed earlier we have two methods of field oriented vector control namely direct and indirect. The use of speed sensor in speed estimation makes the whole arrangement costly and less reliable. The voltage and current sensors can be used as they are very less costly. Verghese did approach speed estimation problem from the view point of parameter identification [13]. The main idea is to take the speed as a constant and unknown parameter, and to find that desired speed which fits the calculated or measured data to the dynamic equations of the motor. However due to the significant impact on the performance of the estimator due to the parameter variations. State resistance variation is also a possibility due to ohmic heating results in performance deterioration. Hence with the idea of having speed estimation without the use of sensors many techniques have been proposed each having its own advantages and disadvantages. Some of these techniques have been discussed here and there efficiency verified with simulation results. Hench in sensorless vector control of induction motor voltage and current sensors are retained but speed or position sensors are eliminated as with voltage and current sensors speed can be sensed although reverse is not possible.

Some of the Induction motor speed estimation techniques are as follows:

(a) Slip calculation

(b) Direct synthesis from rotor flux observer

(c) Model reference adaptive system (MRAS)

(d) Speed adaptive flux observer (Luenberger observer)

(e) Extended Kalman filter

(f) Slot harmonics

3.2 Speed Estimation Schemes

3.2.1 Estimation of synchronous and slip speed

The equations for speed estimation can be written as:

$$\hat{\omega}_r = \hat{\omega}_e - \hat{\omega}_{sl} \tag{3.1}$$

Where $\hat{\omega}_r$ = estimated speed

$\hat{\omega}_e$ = synchronous speed

& $\hat{\omega}_{sl}$ = slip speed

$$V_{ds}{}^{s} = R_s i_{ds}{}^{s} + \dot{\psi}_{ds}^{s} \tag{3.2}$$

$$V_{qs}{}^{s} = R_s i_{qs}{}^{s} + \dot{\psi}_{qs}^{s} \tag{3.3}$$

$$\psi_{ds}{}^{s} = \int (V_{ds}{}^{s} - r_s i_{ds}{}^{s})dt \tag{3.4}$$

$$\psi_{ds}{}^{s} = \int (V_{ds}{}^{s} - r_s i_{ds}{}^{s})dt \tag{3.5}$$

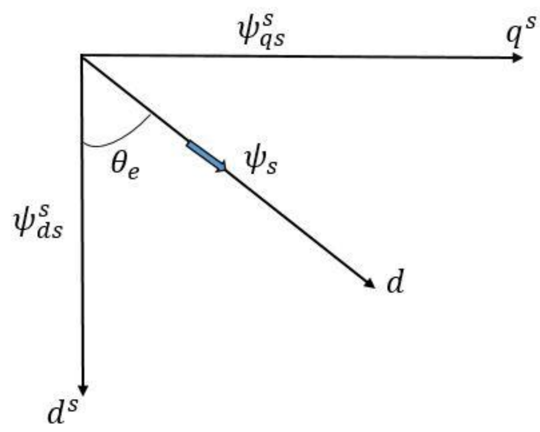

Figure 3-1 Vector control diagram

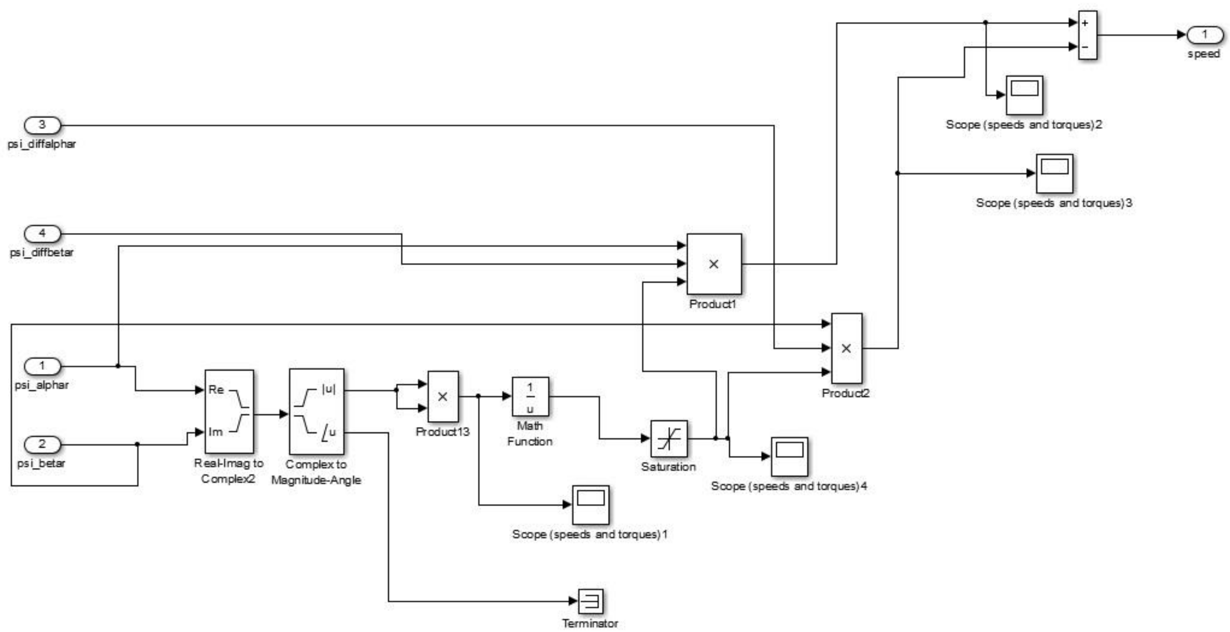

Figure 3-2 Synchronous speed estimation block

The corresponding flux equations for speed estimation can be written as:

$$\hat{\omega}_e = \frac{\psi_{ds}{}^s (v_{qs}{}^s - R_s i_{qs}{}^s) - \psi_{qs}{}^s (v_{ds}{}^s - R_s i_{ds}{}^s)}{\psi_s{}^2} \tag{3.6}$$

$$\hat{\omega}_e = \frac{\psi_{qs}{}^s \dot{\psi}_{ds}{}^s - \dot{\psi}_{ds}{}^s \psi_{qs}{}^s}{\psi_s{}^2} \tag{3.7}$$

Algorithm

1. We sense the phase voltages V_a, V_b, V_c and currents I_a, I_b, I_c .
2. Using the transformation we get $V_{ds}{}^s, V_{qs}{}^s$ and $I_{ds}{}^s, I_{qs}{}^s$.
3. We apply equation 4.2 to 4.5 to get the desired flux and their derivatives.
4. Then the synchronous speed is estimated using the equation 4.6 & 4.7.
5. The motor speed can be calculated from equation 4.1

   Where $\hat{\omega}_{sl}$ is the slip speed whose value is:-

   $$\hat{\omega}_{sl} = \frac{R_r L_m}{L_r} \frac{i_{qs}}{\psi_r} \quad \text{For rotor flux orientation} \tag{3.8}$$

   $$\hat{\omega}_{sl} = \frac{(1 + \sigma s T_r) L_s i_{qs}}{T_r (\psi_{ds} - \sigma L_s i_{ds})} \quad \text{For stator flux orientation} \tag{3.9}$$

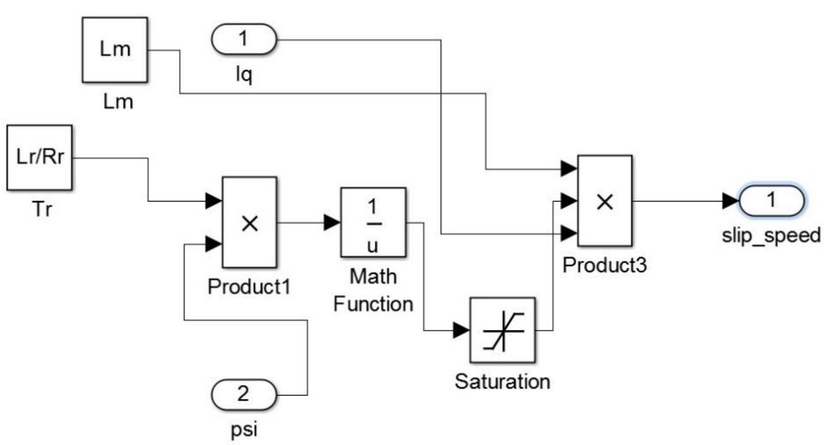

Figure 3-3 Slip speed calculation

This method of speed estimation is one of the simplest method to execute and it also run into some disadvantages. When speed is of less values (e.g. frequency is less) the voltage values $V_{ds}{}^{s}, V_{qs}{}^{s}$ are quite less so then the integration of flux signal derivatives give offset and do not provide desired accurate estimation. Furthermore, at very high speed of the machine (high frequency) slip is very less. So speed estimation depends on the parameter like R_r, L_r. So these values should be accurate otherwise it leads to inaccurate estimation.

3.2.2 Direct synthesis of speed from state equations

The observed stator currents and voltages are thereby used to estimate/measure flux linkages components which are used to find out the speed. The mathematical expression for the rotor speed by the equation $\hat{\omega}_r = \hat{\omega}_e - \hat{\omega}_{sl}$. The corresponding equation for this speed estimation technique are:

$$\dot{\psi}_{dr}{}^{s} = \frac{L_r}{L_m} V_{ds}{}^{s} - \frac{L_r}{L_m}(R_s + \sigma L_s s) i_{ds}{}^{s} \tag{3.10}$$

$$\dot{\psi}_{dr}{}^{s} = \frac{L_r}{L_m} V_{ds}{}^{s} - \frac{L_r}{L_m}(R_s + \sigma L_s s) i_{ds}{}^{s} \tag{3.11}$$

These mentioned equations are state equations. These are used to estimate rotor flux components $\psi_{dr}{}^{s}$ and $\psi_{qr}{}^{s}$ from sensed voltages and currents. Now the current model is used and is written as:

$$\dot{\psi}_{dr}{}^{s} = -\frac{1}{T_r}\psi_{dr}{}^{s} + \frac{L_m}{T_r} i_{ds}{}^{s} - \omega_r \psi_{qr}{}^{s} \tag{3.12}$$

$$\dot{\psi}_{qr}{}^{s} = -\frac{1}{T_r}\psi_{qr}{}^{s} - \omega_r \psi_{dr}{}^{s} + \frac{L_m}{T_r} i_{qs}{}^{s} \tag{3.13}$$

Using these relations, we find out the synchronous speed as:

$$\omega_e = \frac{\psi_{qr}{}^{s}\dot{\psi}_{dr}{}^{s} - \dot{\psi}_{dr}{}^{s}\psi_{qr}{}^{s}}{\psi_r{}^{2}} \tag{3.14}$$

The above equations are used to calculate the speed which is finally given as

$$\hat{\omega}_r = \hat{\omega}_e - \frac{L_m}{Tr}\frac{\psi_{dr}{}^s i_{qs}{}^s - \psi_{qr}{}^s i_{ds}{}^s}{\psi_r{}^2} \tag{3.15}$$

Where $$\hat{\omega}_{sl} = \frac{L_m}{Tr}\frac{\psi_{dr}{}^s i_{qs}{}^s - \psi_{qr}{}^s i_{ds}{}^s}{\psi_r{}^2} \tag{3.16}$$

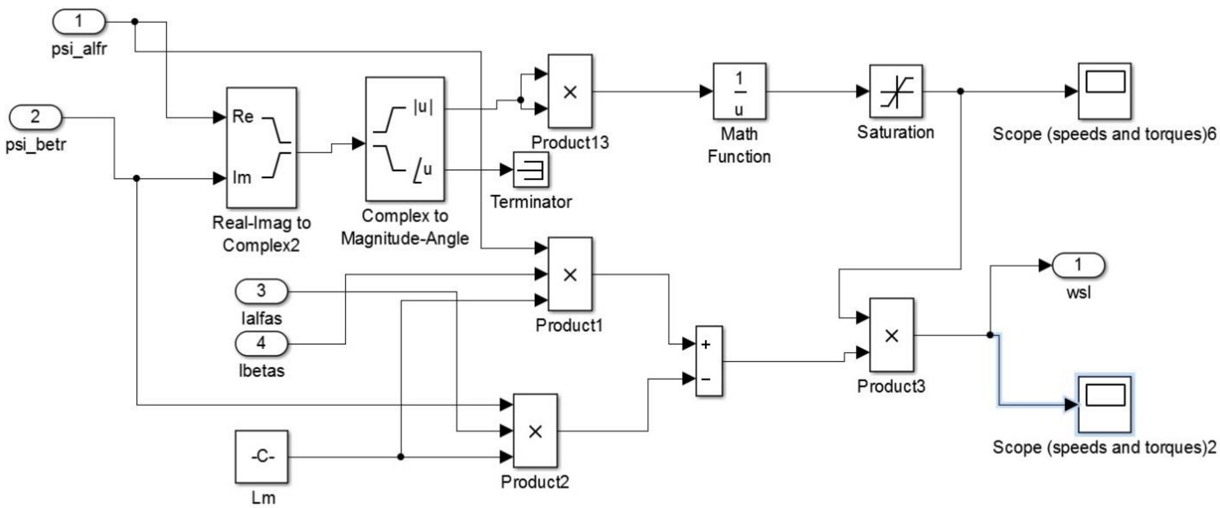

Figure 3-4 Slip speed estimation scheme

This speed estimation scheme follows the same set of steps as mentioned earlier. As evident the hypothesis is highly sensitive to machine parameters that's it tends to provide poor and low accuracy of estimation.

3.2.3 Model Reference Adaptive Systems estimator

The open loop estimator accuracy which is derived from state equations depends strongly on machine parameters. By using closed loop estimator's accuracy can be increased to a good amount by using closed loop estimators. Speed estimator which is based on MRAS is studied in [15]. In this method a comparison in general is usually made among the two estimators output. As a reference model for the induction machine that estimator is chosen which don't contain the quantity to be measured. The other one is referred as adjustable model as this contains the estimated quantity. The input to the adaptation mechanism depends on input from obtained error after comparison between the estimators. In sensorless algorithm control the rotor speed is the element which separate the adjustable model from a reference model. Whenever rotor speed changes in the adjustable model happens in a way that the difference among the two estimators finally converges to zero asymptotically, then actual rotor speed becomes equal to estimated speed of the rotor. In [16] reference model is taken from the voltage model and the current model is taken as the adjustable model and the rotor flux which is estimated is assumed as the reference parameter which is finally compared. Then main differences occurring among these state variables is finally utilized in another adaptation mechanism (PI controller), which then estimates the net desired value of the speed of the rotor keeps on adjusting the adaptive model until a final satisfactory performance is achieved.

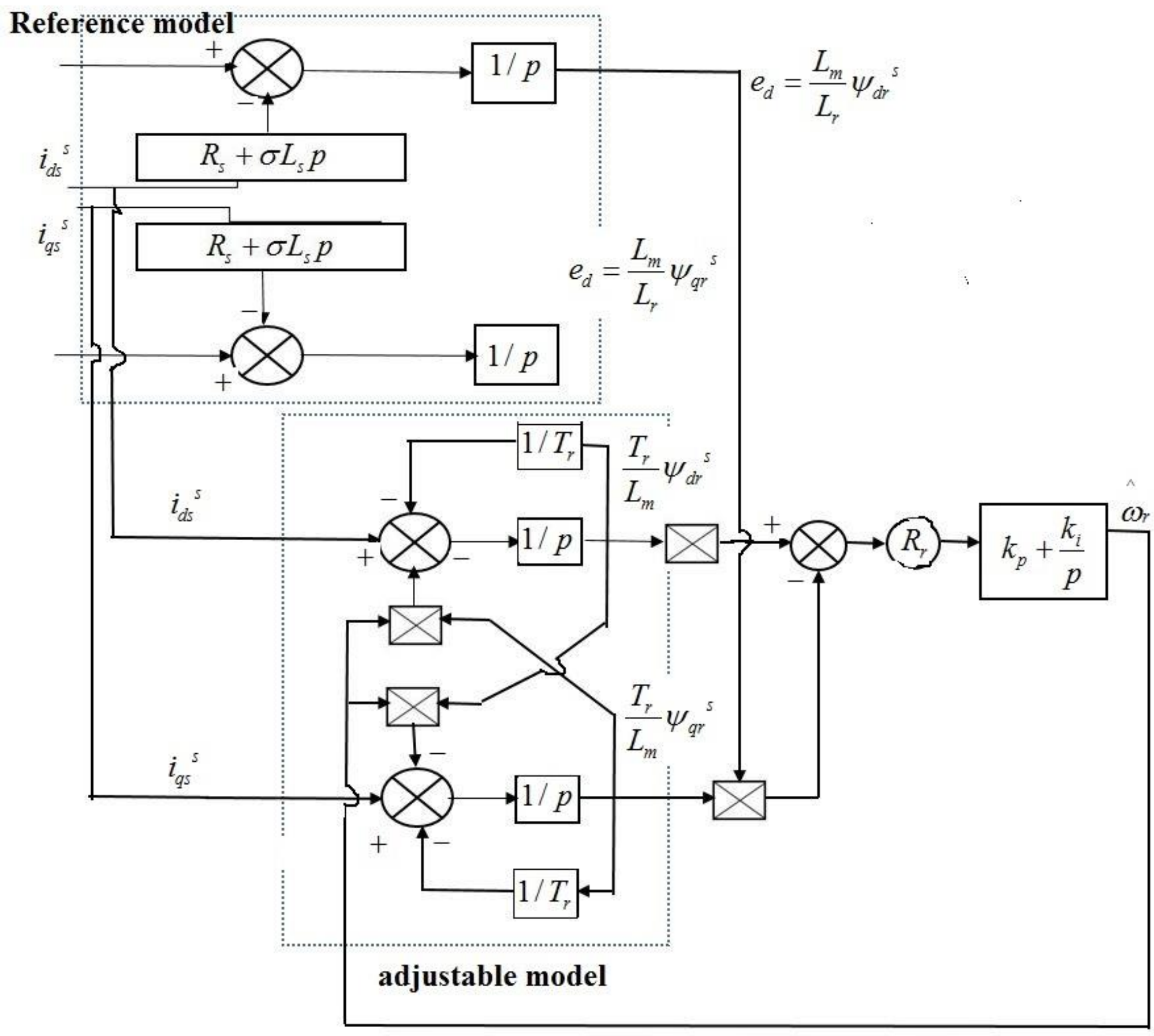

Figure 3-5 MRAS based speed estimator

Reference model makes the use of measured stator voltage and current to finally find out the rotor flux as described:

$$\psi_{dr}{}^{s} = \frac{L_r}{L_m}\frac{1}{s}(V_{ds}{}^{s} - R_s i_{ds}{}^{s} - \sigma L_s \frac{d}{dt} i_{ds}{}^{s}) \tag{3.17}$$

$$\psi_{qr}{}^{s} = \frac{L_r}{L_m}\frac{1}{s}(V_{qs}{}^{s} - R_s i_{qs}{}^{s} - \sigma L_s \frac{d}{dt} i_{qs}{}^{s}) \tag{3.18}$$

The adjustable model calculates/measures the same amount of flux using the estimated speed and the measured current as depicted:

$$\psi_{dr}{}^{s} = \frac{1}{s}(\frac{L_m}{T_r} i_{ds}{}^{s} - \frac{\psi_{dr}{}^{s}}{T_r} - \omega_r \psi_{qr}{}^{s}) \tag{3.19}$$

$$\psi_{qr}{}^{s} = \frac{1}{s}(\frac{L_m}{T_r} i_{qs}{}^{s} - \frac{\psi_{qr}{}^{s}}{T_r} - \omega_r \psi_{dr}{}^{s}) \tag{3.20}$$

In [17] similar type of speed estimators has been taken which is dependent on MRAS and then a secondary variable has been taken as a reference quantity through passage of the rotor flux through a first order delay instead of a pure integration for nullifying the offset. Still as it can be seen that their proposed algorithm produces estimated speed which is inaccurate whenever the excitation frequency is kept underneath a certain level. Apart from these algorithms also undergoes from the uncertainties in parameters of the because of the reference model because parameter variation in the reference can't be modified or corrected. Although [18] suggests an alternative technique of MRAS which is mainly based upon the electromotive force instead of the rotor flux which is considered as reference quantity for estimation of speed and in this method problem of the integration has been successfully overcome to good extent. Further in this a new variable is proposed which basically signifies the instantaneous reactive power for optimizing the magnetizing current. So in this MRAS algorithm the role of stator resistance eventually disappears from major of the equations making the algorithm robust to that particular parameter.

3.2.4 Adaptive Observer

Any state observer is basically a model based state estimator which could be utilized for the state estimation of a non-linear dynamic system. The states are first predicted through the calculations with use of a mathematical model, but these state which are predicted undergoes continuous correction by using a feedback correction scheme. The Stator and rotor equations of the induction machine which are expressed in stator coordinates which are used to obtain a full order speed observer. A full order observer uses the machine electrical model in $d^s - q^s$ frame, where the state variables are stator currents are $i_{ds}{}^s$ and $i_{qs}{}^s$ and the rotor fluxes are $\psi_{dr}{}^s$ and $\psi_{qr}{}^s$.

The Rotor voltage equations can be written as:

$$\dot{\psi}_{dr}{}^s = -\frac{1}{T_r}\psi_{dr}{}^s - \omega_r\psi_{qr}{}^s + \frac{L_m}{T_r}i_{ds}{}^s \tag{3.21}$$

$$\dot{\psi}_{qr}{}^s = -\frac{1}{T_r}\psi_{qr}{}^s - \omega_r\psi_{dr}{}^s + \frac{L_m}{T_r}i_{qs}{}^s \tag{3.22}$$

Equating (4.10) with (4.20) we can get

$$\frac{di_{ds}{}^s}{dt} = \frac{-(R_rL_m{}^2 + R_sL_r{}^2)}{\sigma L_sL_r{}^2}i_{ds}{}^s + \frac{R_rL_m}{\sigma L_sL_r{}^2}\psi_{dr}{}^s + \frac{\omega_rL_m}{\sigma L_sL_r}\psi_{qr}{}^s + \frac{1}{\sigma L_s}v_{ds}{}^s \tag{3.23}$$

$$\frac{di_{qs}{}^s}{dt} = \frac{-(R_rL_m{}^2 + R_sL_r{}^2)}{\sigma L_sL_r{}^2}i_{qs}{}^s + \frac{R_rL_m}{\sigma L_sL_r{}^2}\psi_{dr}{}^s + \frac{\omega_rL_m}{\sigma L_sL_r}\psi_{qr}{}^s + \frac{1}{\sigma L_s}v_{qs}{}^s \tag{3.24}$$

Thus these equations constitute the desired equation which can be written in the form of any general stare equation as

$$\frac{d}{dt}X = AX + BU$$

Where $X = \begin{pmatrix} i_{ds}{}^{s} \\ i_{qs}{}^{s} \\ \psi_{ds}{}^{s} \\ \psi_{qs}{}^{s} \end{pmatrix}$ $U = \begin{pmatrix} v_{ds}{}^{s} \\ v_{qs}{}^{s} \\ 0 \\ 0 \end{pmatrix}$ $Y = \begin{pmatrix} i_{ds}{}^{s} \\ i_{qs}{}^{s} \end{pmatrix}.$ $C = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{pmatrix}$

And $A = \begin{pmatrix} \frac{-(R_r L_m{}^2 + R_s L_r{}^2)}{\sigma L_s L_r{}^2} & 0 & \frac{L_m L_r}{\sigma L_s L_r{}^2} & \frac{L_m \omega_r}{\sigma L_s L_r} \\ 0 & \frac{-(R_r L_m{}^2 + R_s L_r{}^2)}{\sigma L_s L_r{}^2} & -\frac{\omega_r L_r}{\sigma L_s L_r} & \frac{L_m L_r}{\sigma L_s L_r{}^2} \\ \frac{L_m R_r}{L_r} & 0 & -\frac{R_r}{L_r} & -\omega_r \\ 0 & \frac{L_m R_r}{L_r} & \omega_r & -\frac{R_r}{L_r} \end{pmatrix}$ $B = \begin{pmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{pmatrix}$

Thus the input voltage signals are measured from the machine terminal. Now if speed parameter in A matrix is known we can find out the fluxes and the current from the state equations. However, if speed is not correct then there is bound to be deviations between the actual and the estimated value. Thus in the end estimated currents are compared with the actual machine terminal currents and error is injected into the speed estimating algorithms so that this error vanishes.

The speed algorithm used for the speed estimation in this technique is

$$\hat{\omega}_r = k_p (e_{i\alpha s} \hat{\psi}_{\beta r} - e_{i\beta s} \hat{\psi}_{\alpha r}) + k_i \int (e_{i\alpha s} \hat{\psi}_{\beta r} - e_{i\beta s} \hat{\psi}_{\alpha r}) dt \tag{3.25}$$

3.3 Conclusion

In this chapter different Sensorless speed estimation techniques have been described along with their speed estimation ways which has also been described in brief. These techniques utilize the sensed voltage and current values and thus with this the speed is estimated. The experimental verification of these equations have been shown in chap 4 thus establishing the advantages of drive operations without speed sensors.

CHAPTER 4

SIMULATIONS AND RESULTS

4.1 Introduction

The three phase induction motor with suitable rating has been simulated by using Matlab/Simulink model for its speed estimation. The speed tracking performance has been shown with vector control which first includes the use of speed sensor and then without it as required in the sensorless control schemes. A number of speed estimation techniques have been taken into consideration and their performance is verified by simulation results along with its torque variation waveforms. The specification of the IM used for sensorless estimation schemes is 50 HP, 4 pole, 3-phase with parameters: Rs = 0.087 ohm, Rr= 0.228 ohm, Ls= Lr = 0.0355 H, Lm= 0.0347 H, J= 1.662 Kg.m2, B= 0.1 Nm.sec

4.2 Vector Control Simulation Results

Different operating conditions were investigated in order to validate the direct vector control model and to demonstrate the effectiveness of the system modelling and simulation. The speed waveform of the direct vector control shows the reference and actual motor speeds at t =0.5 sec, motor speed is raised from 0 to 70 rad/sec then at t =1 sec it is further decreased to 120 rad/s in reverse direction and so on varied with time to check the effectiveness of the system.

The speed dip at t=6s is due to load torque augmentation. Electromagnetic torque response is also illustrated. Load torque is increased at t=1s. Despite the presence of torque pulsations, the motor follows precisely the load torque value. At t=4s, motor speed is reduced from 60 to -120 rad/s. We notice in both figures that in presence of disturbance the vector control responds adequately and brings back the controlled variable to its desired value. The other two speed control methods are the indirect control method and the stator oriented control methods and these are simulated for same set of reference speed as the direct oriented control. The speed response gets better in indirect control as compared to the direct one although the ripples in speed response are there but are of low magnitudes and thus overall system gives a good performance. The stator oriented control in comparison shows more ripples and speed response is little less good than the indirect method still the response is very much fine.

4.2.1 Direct Vector Control

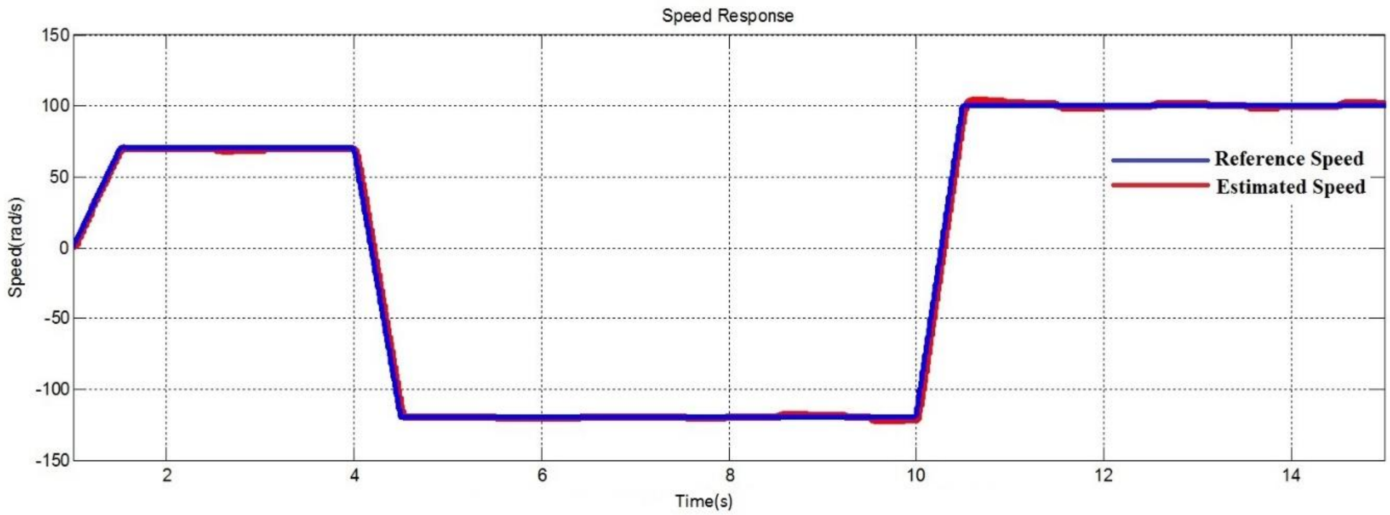

Figure 4-1 Speed Response of Direct Vector Control

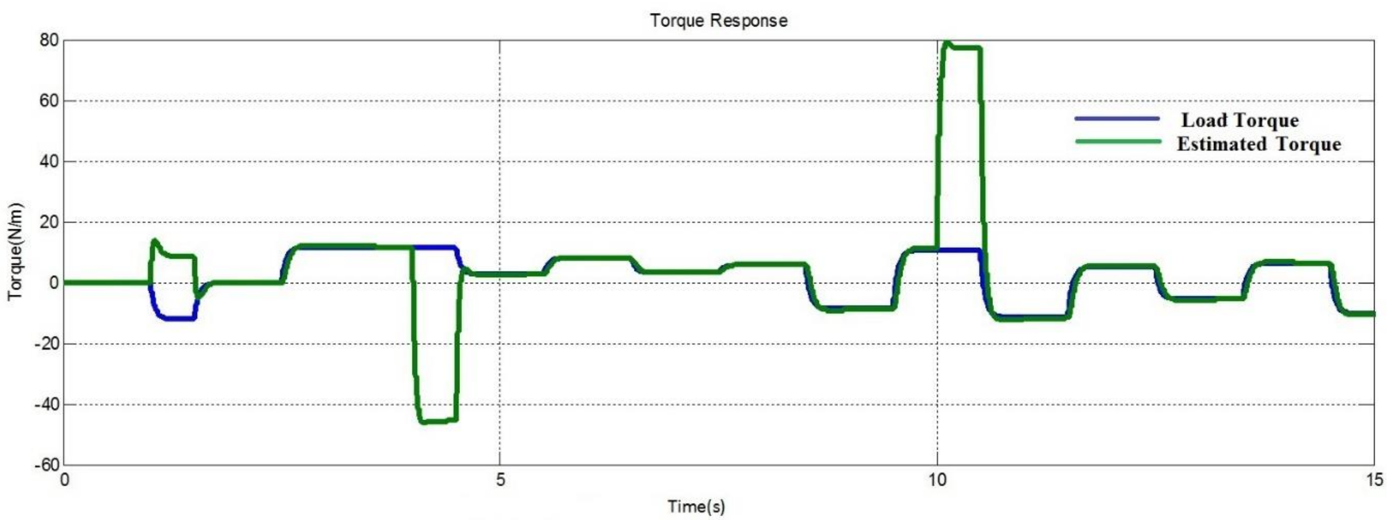

Figure 4-2 Torque Response of Direct Vector Control

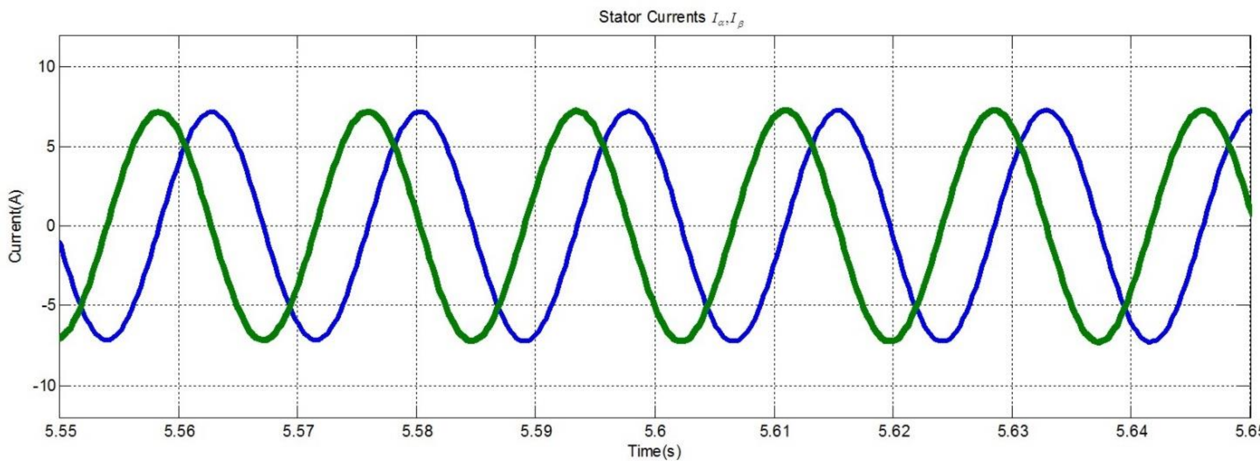

Figure 4-3 Waveform of stator current I_{α}, I_{β}

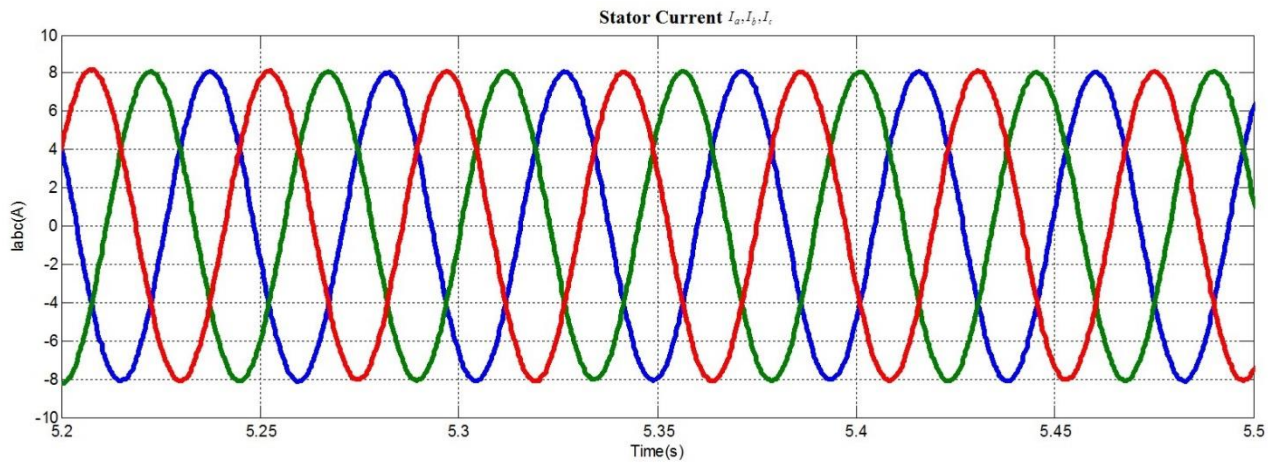

Figure 4-4 Waveform of Stator current I_a, I_b, I_c

4.2.2 Indirect Vector Control

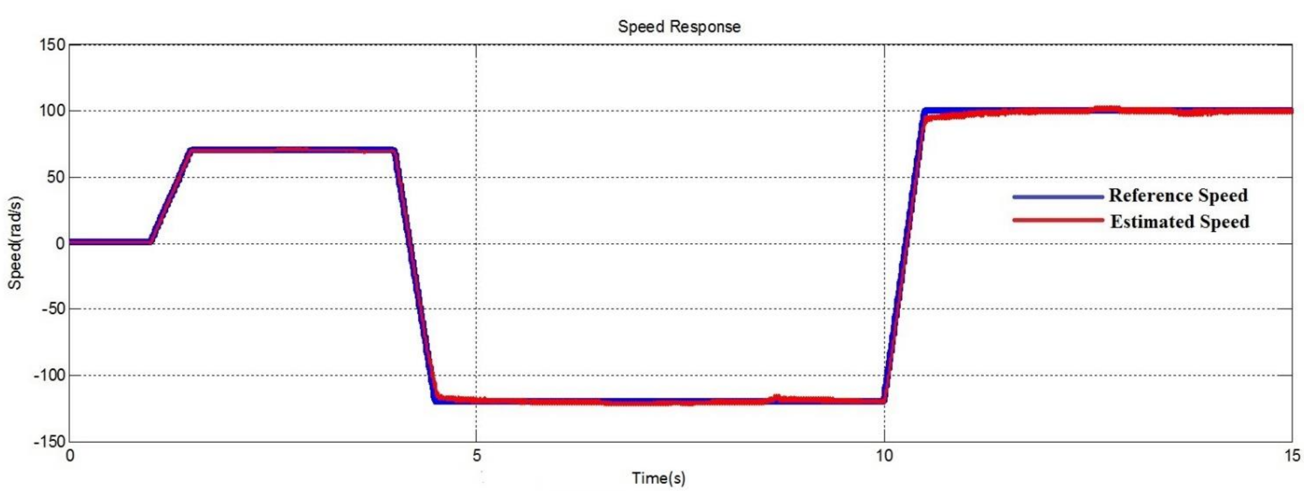

Figure 4-5 Speed Response of Indirect Vector Control

4.2.3 Stator Oriented Control

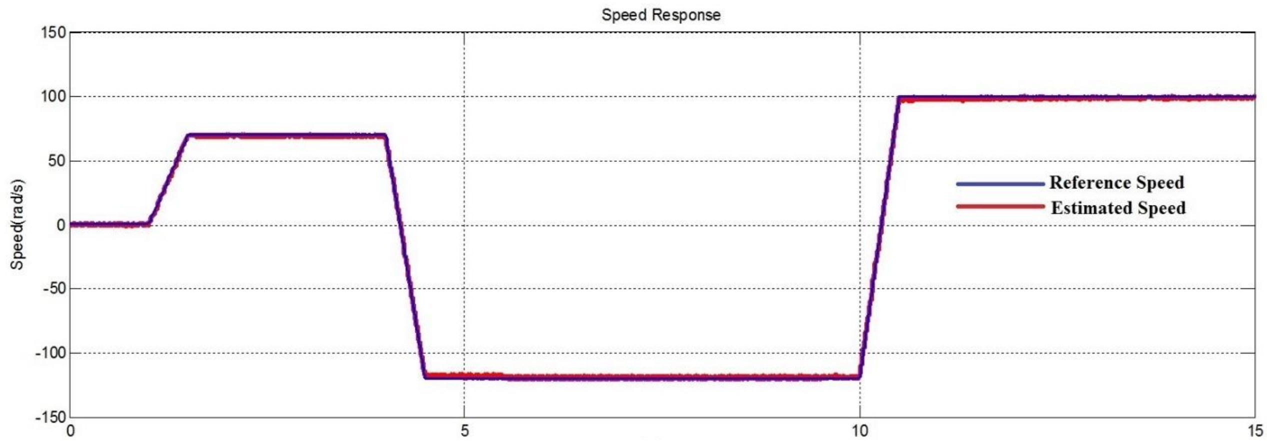

Figure 4-6 Speed Response of Stator Oriented Vector Control

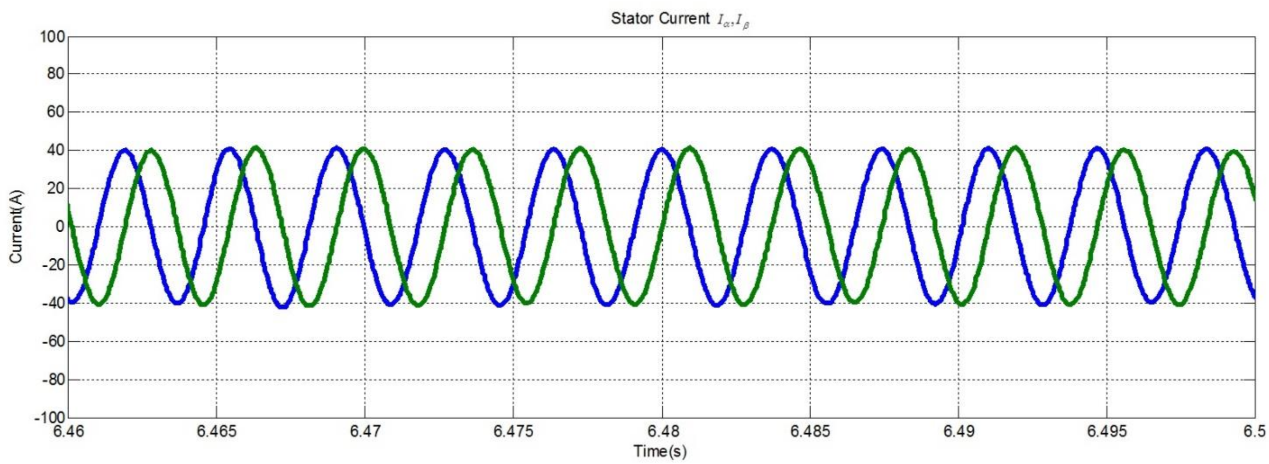

Figure 4-7 Waveform of Stator Currents $I_{\alpha,} I_{\beta}$

4.3 Sensorless Vector Control Simulation Results

4.3.1 Estimation of synchronous and slip speed

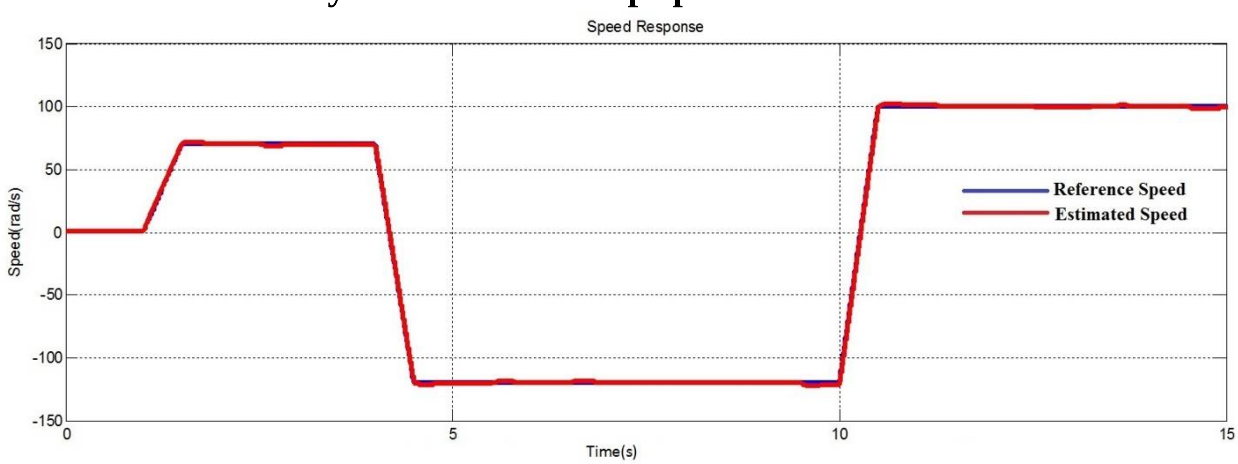

Figure 4-8 Speed Response of sensorless slip estimator scheme

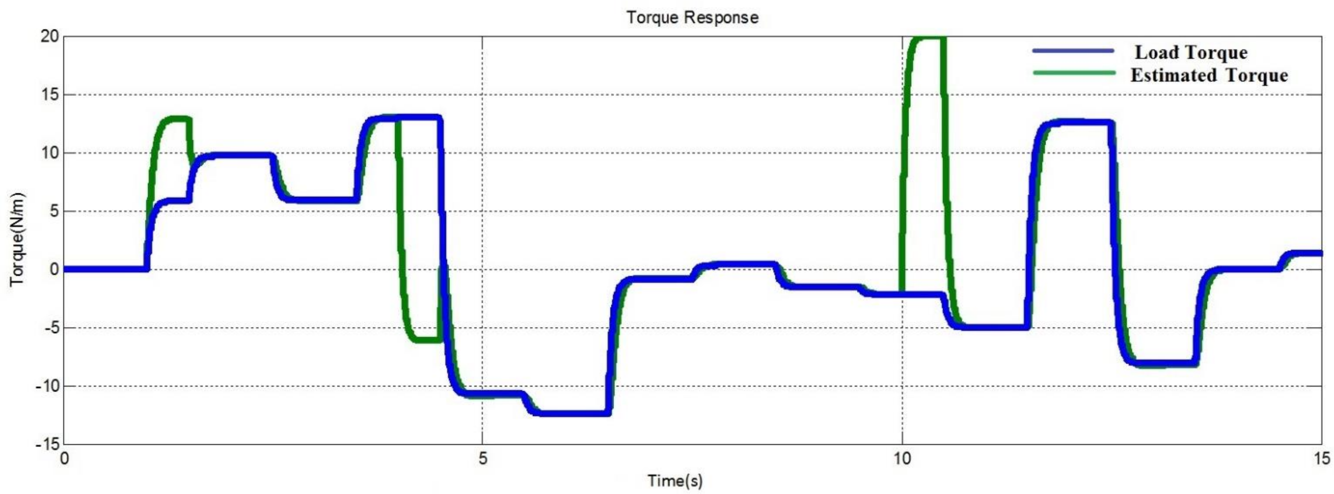

Figure 4-9 Torque response of sensorless speed estimator scheme

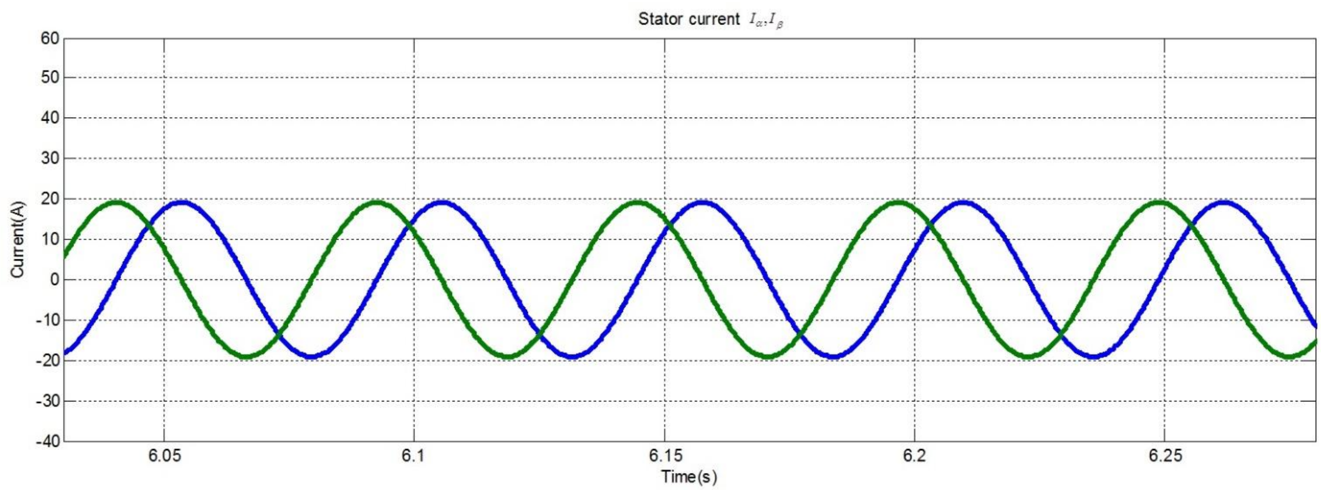

Figure 4-10 Waveform of Stator Currents I_{α}, I_{β}

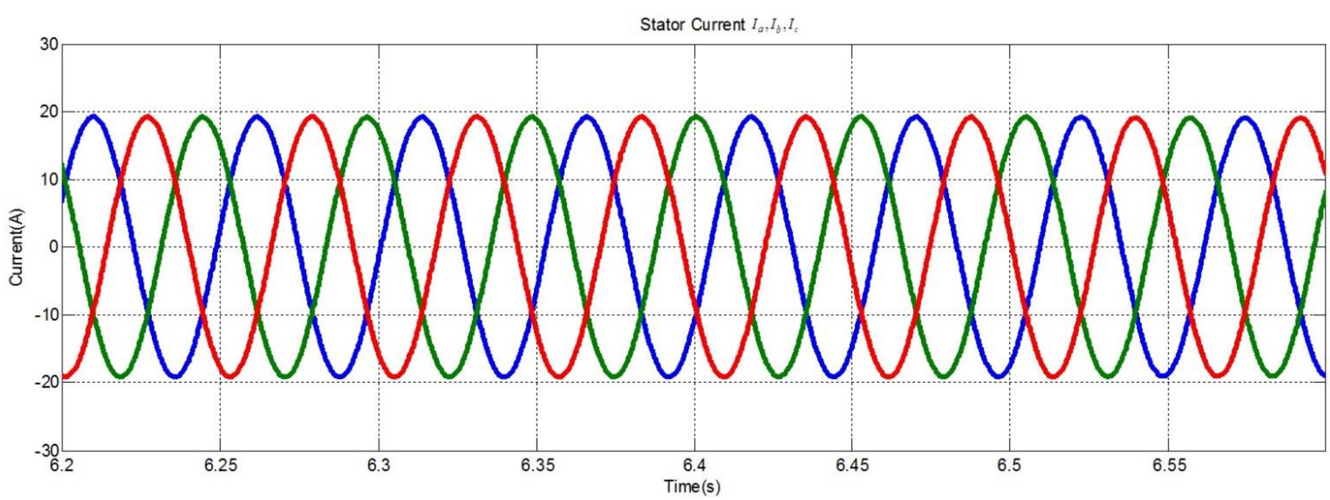

Figure 4-11 Waveform of Stator Currents I_a, I_b, I_c

4.3.2 Direct synthesis of speed from state equations

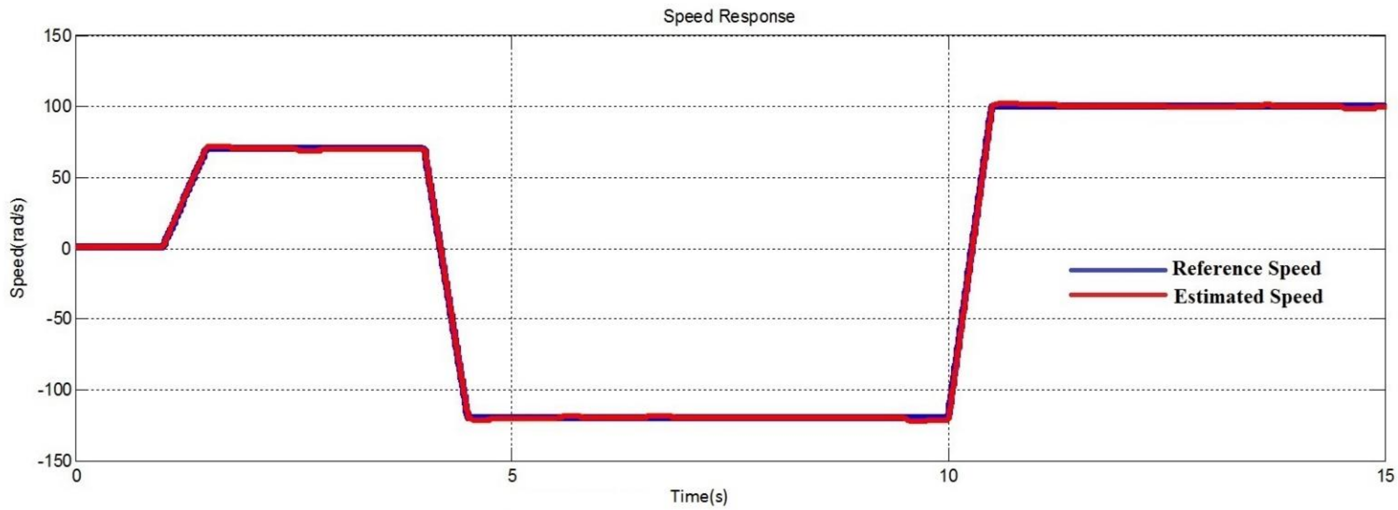

Figure 4-12 Speed response of open loop estimator scheme

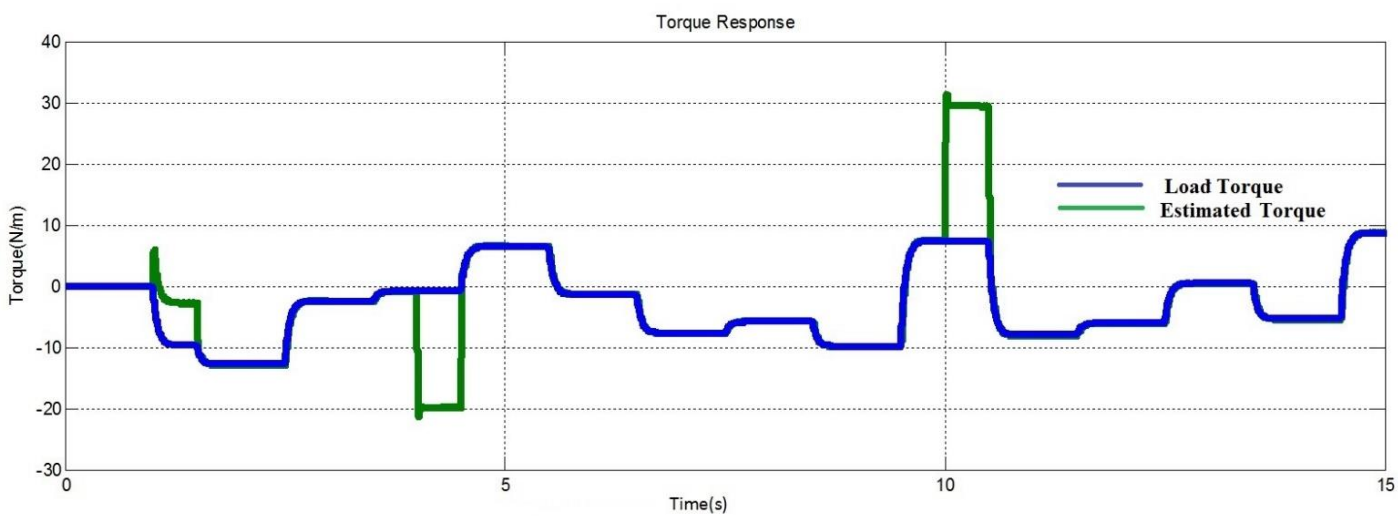

Figure 4-13 Torque response of open loop estimator scheme

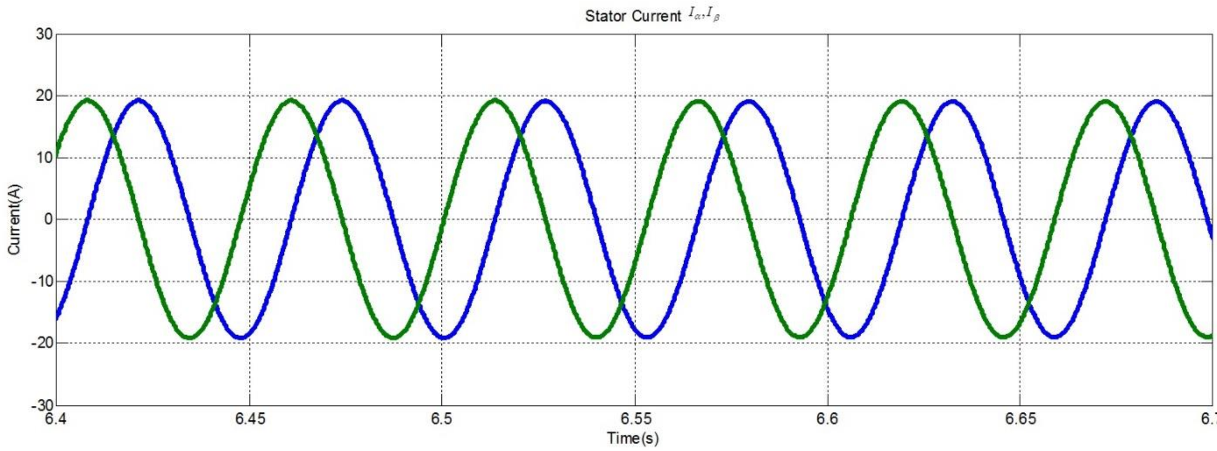

Figure 4-14 Waveform of Stator Current $I_{\alpha,}I_\beta$

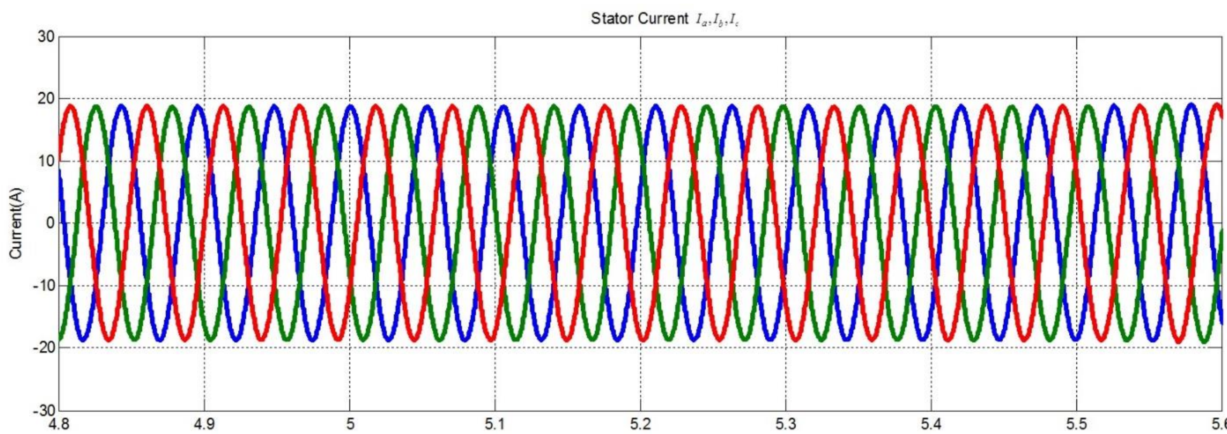

Figure 4-15 Waveform of Stator Current I_a, I_b, I_c

4.3.3 MRAS Estimator

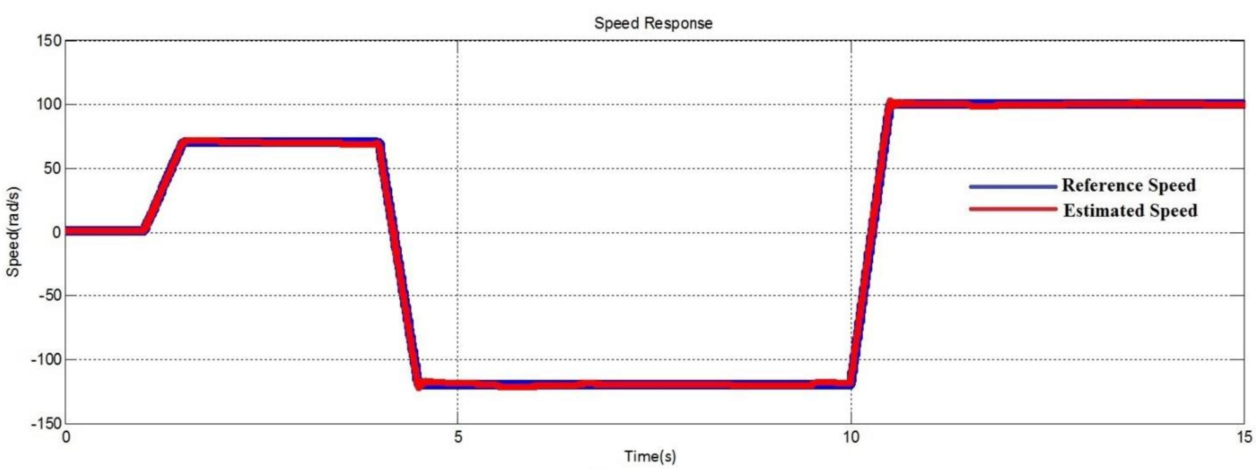

Figure 4-16 Speed response of MRAS

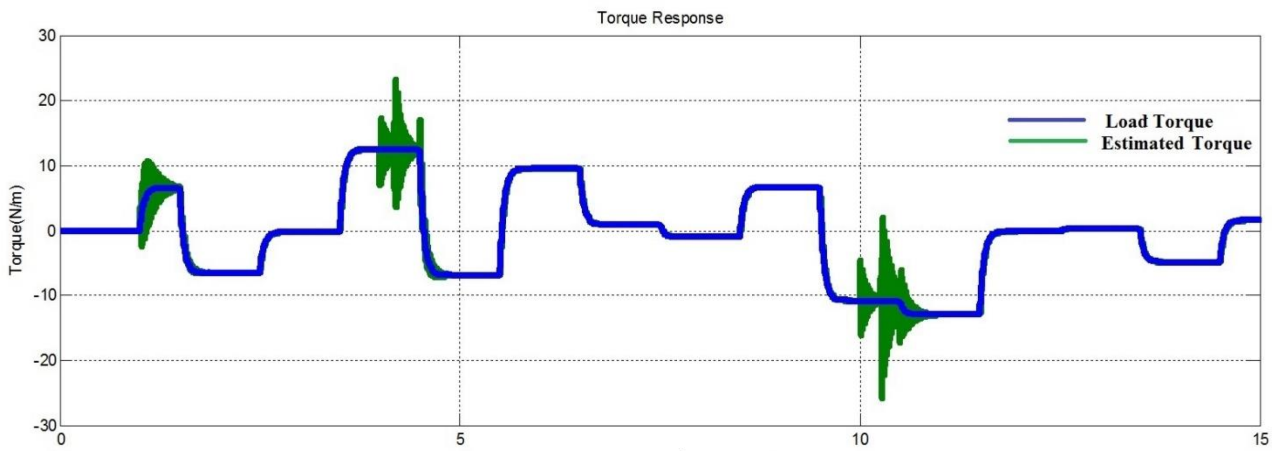

Figure 4-17 Torque response of MRAS

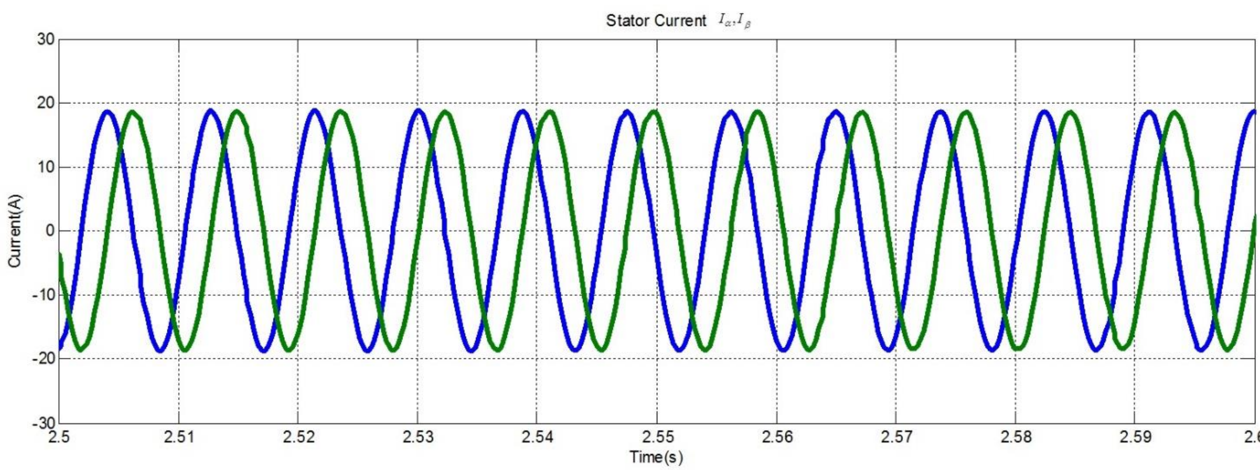

Figure 4-18 Waveform of Stator Current $I_{\alpha,} I_{\beta}$

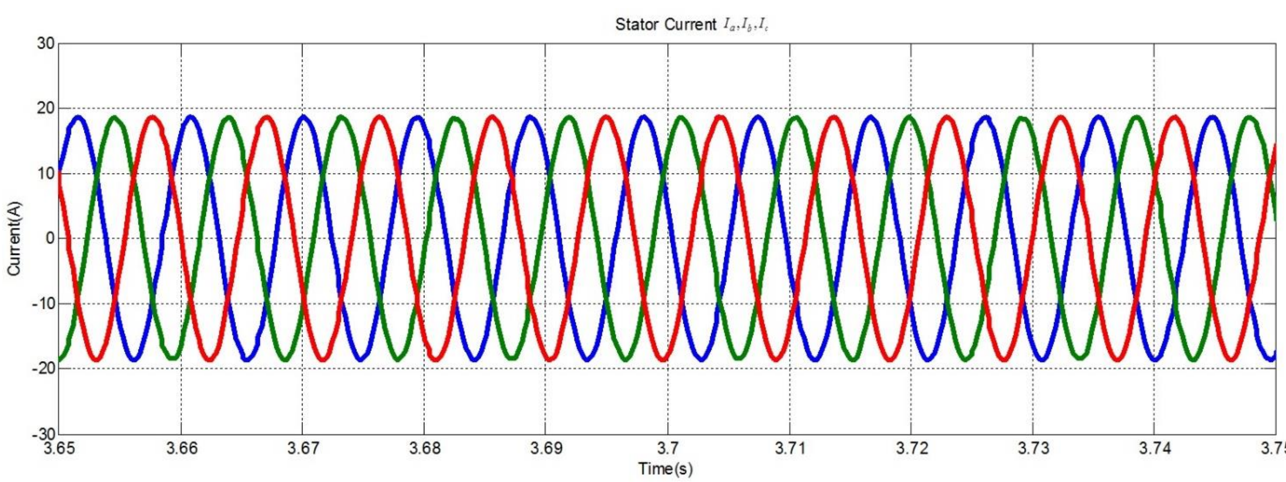

Figure 4-19 Waveform of Stator Current I_a, I_b, I_c

4.3.4 Adaptive Speed Observer

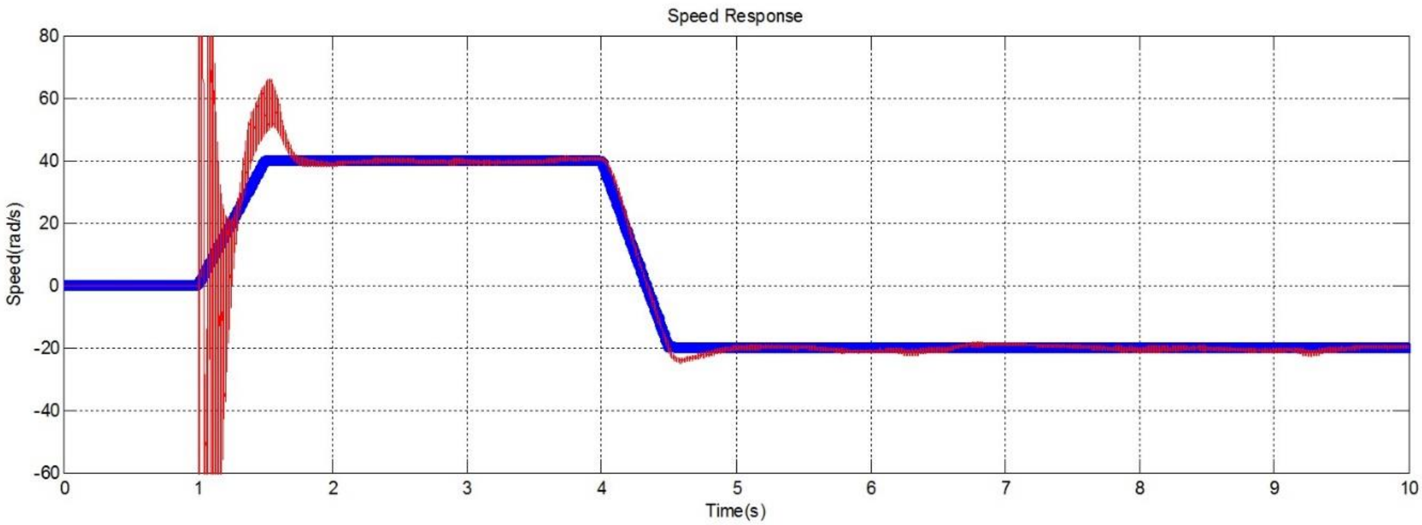

Figure 4-20 Speed Response of Adaptive Observer

All sensorless speed estimation methods have been simulated for rotor flux orientation model. The motor is supplied a three phase supply and the measured stator and rotor currents are fed to the estimator as inputs. Different operating conditions were investigated in order to validate speed estimation model and to demonstrate the effectiveness of the system modelling and simulation. Whenever there is sudden speed change a torque ripple can be seen in the torque waveform. Despite the torque pulsation the speed waveform is found to be following the reference values. The speed is similarly varied and the speed waveform is found to be following the reference value with little error showing the effectiveness of the model.

4.4 Performance comparison of Sensorless speed estimation schemes

The speed estimation becomes difficult particularly at low speed because of the integration of the flux signal derivative gives offset values thus leading to the inaccurate estimation. Thus accuracy especially at low speed values deteriorates due to variation in parameter. Thus for comparison of these speed estimation techniques all these methods have been simulated from Fig 4.12 to 4.15 showing speed variation particularly at low speed.

4.4.1 Estimation of synchronous and slip speed

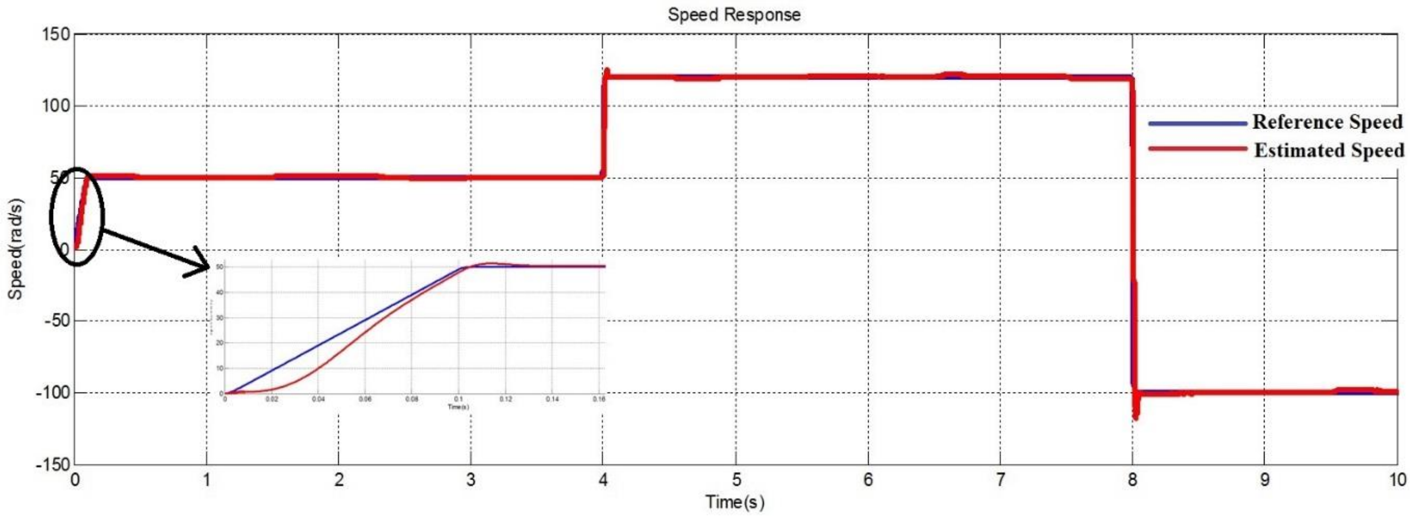

Figure 4-21 Speed Response of Slip Estimation Scheme

4.4.2 Direct synthesis of speed from state equations

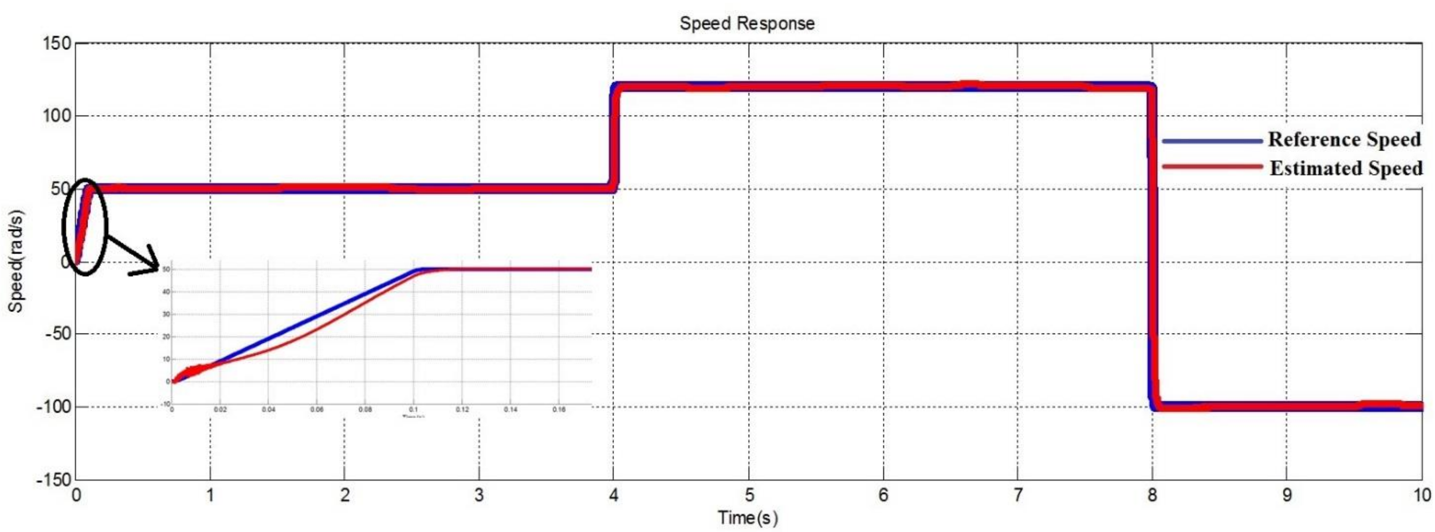

Figure 4-22 Speed Response of Open loop estimator

4.4.3 MRAS Estimator

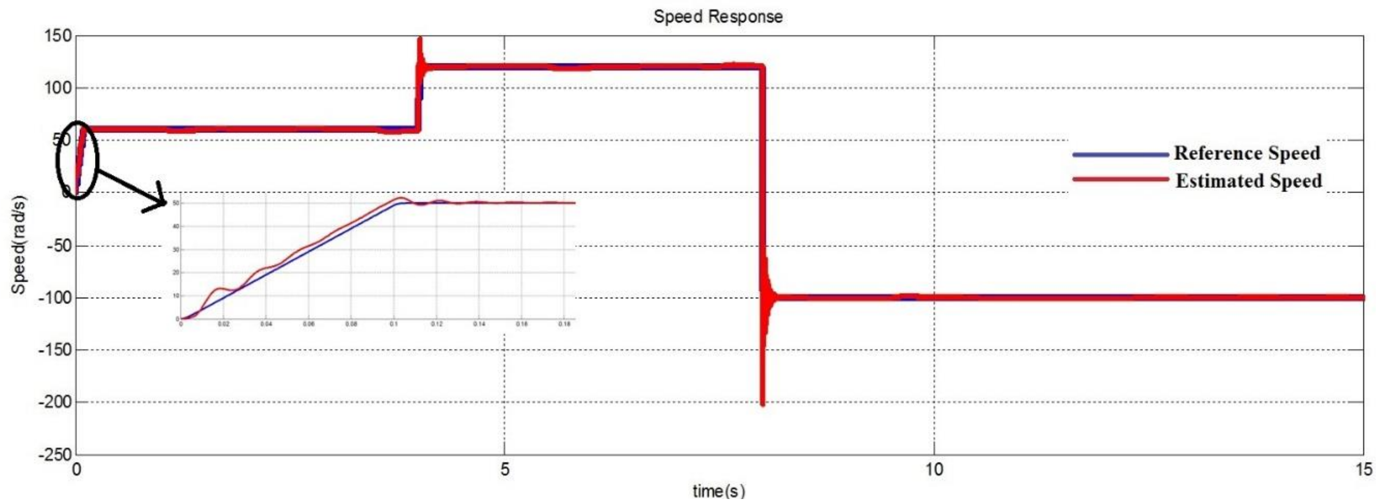

Figure 4-23 Speed Response of MRAS scheme

4.4.4 Adaptive Speed Observer

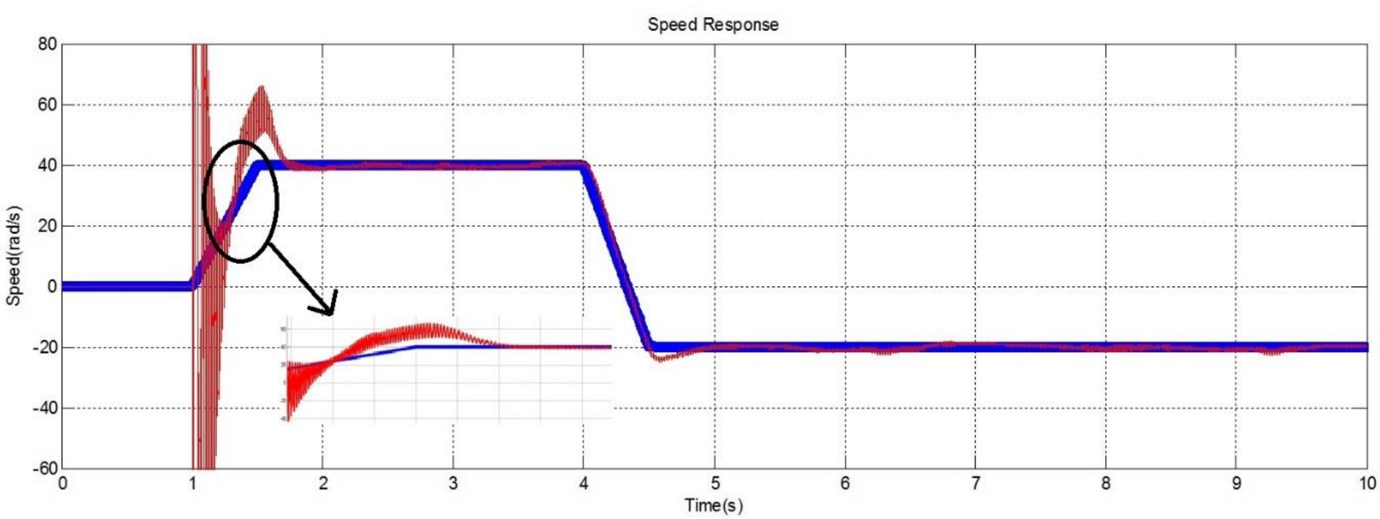

Figure 4-24 Speed Response of Adaptive speed Observer

4.5 Conclusion

In this chapter vector control techniques which uses sensors namely direct, indirect and stator oriented control have been simulated to show the improved performance of the vector control of the induction machine in comparison to the scalar method. The obtained results have been perfect showing the efficiency of this techniques. Thereafter different sensorless speed estimation techniques have been simulated and for their performance comparison at low speed and steady state they have been simulated and compared accordingly.

CHAPTER 5

CONCLUSION AND FUTURE SCOPE

5.1 Conclusion

In this dissertation the main focus has been on finding out the best speed estimation techniques for the sensorless operation of the induction motor. The majority of techniques of speed extraction rely on the induction machine which can also lead to inaccuracies in speed estimation. First mathematical System modelling and various simulations have been shown using Matlab/Simulink. As described rotor flux can be found out by means of rotor flux observer where they are obtained by utilizing voltage and current equations of the IM in stationary reference frame. Similarly, with the help of voltage and current sensors speed is estimated for various sensorless schemes. For high performance drives different flux and speed estimation techniques can be used.

Among there presented schemes MRAS scheme shows best behavior at steady state operation as demonstrated in section 4.4. It can be seen it provides a very good response. The response is as good as direct vector control thus showing its effectiveness. The simulations show there is room for improvement in performance of the system particularly for low speed range. During regular change in loading system stability could be improved to achieve better vector control performance.

5.2 Scope for future work

In this thesis the simulation work shows a great promise for the studied methods. Different estimator's dynamic response and the estimation accuracy can be further improved by re-tuning and also with the help of online parameter estimation techniques will facilitate observation of change of motor parameters during operation especially at low speed. Several other methods can also be applied like Extended Kalman Filter (EKF), neural network based estimators and sliding mode estimators.

References

[1] F. Blaschke, "*A new method for the structural decoupling of A.C. induction machines*," in Conf. Rec. IFAC, Duesseldorf, Germany, Oct. 1971, pp. 1–15.

[2] A. Abbondanti, M. B. Brennen, "*Variable speed induction motor drives use electronic slip calculator based on motor voltages and currents*," IEEE Trans. Industrial Applications, vol.11, 1975, pp. 483-488.

[3] B. K. Bose, *Modern Power Electronics and AC Drives*, Prentice-Hall, New Delhi, 2002.

[4]. W. L. Erdman and R. G. Hoft, *"Induction machine field orientation along air gap and stator flux," IEEE Trans. On Energy Conversion, vol. 5, no. 1, pp.115-121, Mar 1990.*

[5] S. Xepapas, A. Kaletsanos, F. Xepapas, S. Manias, "*Sliding-mode observer for speed sensorless induction motor drives*," *IEE Proc.-Control Theory Application*, vol. 150, no. 6, November 2003, pp.788-795.

[6]. Ahmad Razani Haron, Nik Rumzi Nik Idris**,** *"Simulation of MRAS-based Speed Sensorless Estimation of Induction Motor Drives"* Power and Energy Conference, PECon '06. IEEE International, page 411-415 November 28-29, 2006.

[7] M. Ishida, K. Hayashi, M. Ueda, "*A speed detection method of squirrel cage induction motor utilizing rotor slot-harmonics in the air gap and its application to slip frequency control*," *Electrical Engineering, Japan*, vol. 99, no. 3, 1979, pp. 74-84.

[8] J. Moreira, K. Hung and T. Lipo, "*A simple and robust adaptive controller for the tuning correction in field-oriented induction machines*," IEEE Trans. Ind. Applicat., vol. 28, pp. 1359-1366, Nov./Dec. 1992.

[9] P.L.Jansen and R.D.Lorenz "*Transducerless position and velocity estimation in induction and salient ac machines*", IEEE Tran. IA, vol. 31, no.2, pp.240-247, Mar/April, 1995

[10] B.Robyns, B.Frederique, "*A methodology to determine gains of induction motor flux observers based on a theoretical parameter sensitivity analysis*", IEEE Trans. PE, vol. 15, no.6, pp.983-995, 2000

[11] Takahashi and T. Noguchi, "*A new quick-response and high-efficiency control strategy of an induction motor*," IEEE Trans. Ind. Applicat., vol. 22, pp.820- 827, Sep./Oct. 1986.

[12] P. L. Jansen, C. O. Thompson, and R. D. Lorenz, "*Observer based direct field orientation for both zero and very high speed operation,*" in Proc. PCC, Yokohama, Japan, Apr. 1993, pp.432-437.

[13] K Minami, M. Vélez-Reyes, D. Elten, G. C. Verghese and D. Filbert, "*Multistage speed and parameter estimation for induction machines*," IEEE Power Electronics Specialist's Conference, Boston MA, Jun. 1991.

[14] I. J. Ha and S. H. Lee, "*An on-line identification method for both stator and rotor resistances of induction motors without rotational transducers*," ISIE'96, Warsaw, Poland, Jun. 1996.

[15] C.Schauder "*Adaptive Speed Identification for Vector Control of Induction Motor without Rotational Transducers*" IEEE Tran. IA vol. 28, no. 5, pp. 1054-1061, Oct.1992

[16] H.Tajima, Y.Hori "*Speed Sensorless Field Orientation Control of the Induction Machine*" IEEE Tran. IA vol. 29, no. 1, pp. 175-180, Feb.1993

[17] C.Schauder "*Adaptive Speed Identification for Vector Control of Induction Motor without Rotational Transducers*" IEEE Tran. IA vol. 28, no. 5, pp. 1054-1061, Oct.1992

[18] F.Z.Peng, T.Fukao "*Robust Speed Identification for Speed Sensorless Vector Control of Induction Motors*" IEEE Tran. IA vol. 30, no. 5, pp.1234-1239, Oct.1994

[19] I. Boldea and S. A. Nasar, "*Vector Control of AC Drives*", CRC Press, 1992.

[20] Radu Bojoi, Paolo Guglielmi and Gian-Mario Pellegrino,"*Sensorless direct field-oriented control of three-phase induction motor drives for low-cost applications*," IEEE Trans. Ind. Appl., vol. 44, no. 2, pp. 475-481, Mar. 2008.

[21]Jaison Cherian and Johnson Mathew," *Parameter Independent Sensorless Vector Control of Induction Motor*" *IEEE International Conference on Power Electronics, Drives and Energy Systems December16-19, 2012, Bengaluru, India*

[22] Ahmed A.Z. Diab," *Implementation of Full Order Observer for Speed Sensorless Vector Control of Induction Motor Drive* " 15th International conference on micro/nanotechnologies and electron decices EDM

[23]Renukrishna B. and Shanifa Beevi," *Sensorless Vector Control of Induction Motor Drives using Rotor Flux Observer*" *IEEE International Conference on Power Electronics, Drives and Energy Systems December16-19, 2012, Bengaluru, India*

[24] Djordje Stoji´c, Milan Milinkovi´c, Slavko Veinovi´c," *Improved Stator Flux Estimator for Speed Sensorless Induction Motor Drives*" IEEE Transactions on power electronics , Vol. 30, No. 4, April 2015

APPENDIX

FOC & Open loop simulation of IM:

Power Rating: 3 HP

Stator Voltage: 460 volt

Frequency: 50 Hz

Number of Poles: 4

Stator Resistance: 0.6837 ohm/phase

Stator leakage Inductance: 0.004152 H/phase

Rotor Resistance: 0.451 ohm/phase

Rotor leakage Inductance: 0.004152 H/phase

Mutual Inductance: 0.1486 H

Inertia: 0.05 kg m_2

Sensorless Vector Control:

Power Rating: 50 HP

Stator Voltage: 460 volt

Frequency: 60 Hz

Number of Poles: 4

Stator Resistance: 0.087 ohm/phase

Stator leakage Inductance: 0.0355 H/phase

Rotor Resistance: 0.228 ohm/phase

Rotor leakage Inductance: 0.0355/phase

Mutual Inductance: 0.0347 H

Inertia: 1.662 kg m_2