République Algérienne Démocratique et Populaire

Ministère de l'enseignement Supérieur

et de la Recherche scientifique

Université Echahid Hamma Lakhdar d'El-Oued

Faculté de Technologie

Mémoire de Fin d'Étude

En vue de l'obtention du diplôme de

MASTER ACADEMIQUE

Domaine : Sciences et Technologie

Filière : Electrotechnique

Spécialité : Commande Electrique

Thème

Sensorless speed Control based on Interconnected High Gain Observer for Induction Motor

Réalisé par :

Fethiza Tedjani Tedjani

Gouacem Raid Akram

Encadré par :

Dr. Zoheir TIR

Soutenu en Sebtembre 2019

ACKNOWLEDGMENT

We would like to express our deepest thanks to all those people who influenced our work. We would like to add few heartfelt words for the people who gave their unending support with warm wishes.

We would like to thank our supervisor, **Dr. Tir Zoheir** for all the advices and guidance throughout our project. Without his continued support and interest, the project may not be done as best as it is.

We heartily acknowledge the cooperation and moral support of our families, friends, professors and colleagues who always supported us in the entire course of study and project work.

List of Abbreviations and Symbols

IM: Induction motor

s, r : Index corresponding to stator and rotor

a, b, c : Index corresponding to stator and rotor phases

d, q : Axes corresponding to the frame linked to the rotating field

α, β : Axes corresponding to the frame linked to the stator

[T] : Park matrix

ω : is the stator pulsation

ω_{sl} : Is the pulsation of the sliding

ω_s : Stator pulsation

Ω : Mechanical speed of the rotor

p: Number of pole pairs

v_{sq}, v_{sd} : Stator voltage in the reference frame dq

i_{sq}, i_{sd} : Stator current in the dq reference frame

v_{rq}, v_{rd} : Rotor voltage in the reference frame dq

PMDC : Permanent Magnet Direct Current

PID : Proportionnal Integral-Derivation

PWM : Pulse Width Modulation

i_{rq}, i_{rd} : Rotor current in the dq reference frame

$(u_{s\alpha}, v_{s\alpha})$, $(v_{s\beta,} u_{s\beta})$: Stator voltage in the reference frame $\alpha\beta$

$i_{s\alpha}, i_{s\beta}$: Stator current in the $\alpha\beta$ reference frame

$v_{r\alpha}, v_{r\beta}$: Rotor current in the $\alpha\beta$ reference frame

$\widehat{}$: Refers to observation value

$\widetilde{}$: Refers to estimation value

$i_{r\alpha}, i_{r\beta}$: Rotor current in the $\alpha\beta$ reference frame

Ψ_{sd}, Ψ_{sq}: Stator fluxes in the dq reference frame

Ψ_{rd}, Ψ_{rq} : Rotor fluxes in the dq reference frame

$\Psi_{s\alpha}, \Psi_{s\beta}$: Stator current in the $\alpha\beta$ reference frame

$\Psi_{r\alpha}, \Psi_{r\beta}$: Rotor fluxes in the $\alpha\beta$ reference frame

R_r, R_s : Rotor and stator resistance

L_r, L_s : Own cyclic inductance stator and rotor per phase

M: Mutual inductance between stator and rotor

T_e : Electromagnetic torque

T_L : Load torque

J : Moment of inertia

f_v : Friction coefficient

σ : the coefficient of dispersion.

T_r : Rotor time constant

FOC: Field oriented control

IRFOC: Indirect rotor field oriented control

DFOC: Direct field oriented control

X: State vector

y: output vector

u: Input vector

L : Luenberger observer gain

EKF: The Extended Kalman Filter.

S: The gain of interconnected high gain observer

θ : The constant of interconnected high gain observer

List of Figures

Abstract: This project concerns the observation and control of the induction motor without speed sensor using an interconnected high gain observer, the main goal of this work is to construct, test and evaluate the performance of the mentioned observer associated with the sensorless field oriented control of induction motor. The concerned observer proposed to design the system in two subsystems one as estimator and the second as observer. The results obtained illustrate the good performance in term of robustness, despite the problems of unobservability when the motor runs at low speed.

Résumé: Ce projet concerne l'observation et la commande du moteur à induction sans capteur de vitesse à l’aide d’observateur interconnecté à grand gain. L’objectif principal est de construire, de tester et d'évaluer les performances de l’observateur mentioné, associés au commande sans capteur du moteur à induction. L’observateur concerné basé sur la décomposition du système en deux sous-systèmes, l’un fonctione comme un estimateur et l’autre comme un observateur. Les résultats obtenus illustrent la bonne performance en terme de robustesse, malgré les problèmes d’inobservabilité lorsque le moteur tourne à basse vitesse.

ملخص: يتعلق هذا المشروع بالمراقبة والتحكم في المحرك اللاتزامني بدون لاقط السرعة باستخدام مراقب ذو المعامل الكبير، والهدف الرئيسي هو إنشاء، اختباروتقييم أداء المراقب المذكور، والذي يرتبط بالتحكم بدون لاقط للمحرك اللاتزامني.المراقب المقترح يعتمد أساسا على نظام ينطوي على نظاميين فرعيين، أحدهما يعمل كمقدر والآخر يعمل كمراقب، توضح النتائج التي تم الحصول عليها الأداء جيد من حيث المتانة. على الرغم من مشاكل عدم القدرة على الملاحظة عند تشغيل المحرك بسرعة منخفضة.

GENERAL INTRODUCTION:

Sensorless field oriented control of induction motors has become popular due to reliability and maintanenence concerns. Precise speed and torque for an induction motor can be monitored based on dynamic characteristics of the induction motor by using the field oriented control theory and high performance digital signal processor. The indirect field oriented control of an induction motor requires precise speed information; therefore a speed sensor is usually used on the shaft of the motor for measuring the motor speed. However, a speed sensor increases the cost and requires a connection line between the control system and the motor, which may lead to unstable operation of the control system due to interference from the signal line. In addition, an exact servo motor control performance is needed in a system in some cases where the attachment of a speed sensor is impossible. The sensorless field oriented control that can precisely control an induction motor without a speed sensor has been taken great interest, and results of studies have given various speed estimation algorithms and sensorless control methods. The role of an observer is to monitor the dynamics of the state as basic information in the system to increase the robustness of the control and improve its performance.

This master project is organized as follows:

- The first chapter is concerned to give an overview on sensorless speed control where we will present some well known types of observers as well as some sensorless speed control techniques.
- In the first part of the second chapter, we will present the mathematical model of induction motor in different frames using park and Clark transformations to make it possible developing a field oriented control associated with the observer for IM then validating the results using MATLAB SIMULINK. While in the second part we will show the principal of field oriented control for IM and its main advantages then we will finish this section by applying some tests to check the reliability of the control system.

- In the third chapter we will present the interconnected high gain observer for sensorless speed control of IM. Then using MATLAB SIMULINK we will test how the observer works and how much it is robust to motor parameters' variations and then we will finish this section by testing the whole sensorless speed control for IM.

Chapter I
Overview on Sensorless Control

I-1 Introduction:

The implementation of control system is often based on the assumption that all states are available for online measurement; however, in practice, some of them may not be measurable due to a lack of appropriate estimating devices or the high price of sensors [1]. As a consequence, measuring the missing states or variables is expensive and time consuming due to the significant technical standard requirements and the high cost of installation of these devices. For these reasons, devices called observers have been developed to reconstruct the state vector in order to estimate the missing variables and, at the same time, to reduce the usage of high-priced sensors. Luenberger [2,3,4] and Kalman [5] introduced the basic concepts of state observers and Kalman Filter based observers in the 1960s. However, over the years, research in the design of observers has become popular but challenging due to the requirements of high accuracy, low cost and good prediction performances. In fact, many observers today are simply modifications and extended versions of the classical Luenberger observer and Kalman filter. In recent years, various types of observers have been developed to accurately estimate state variables in linear and nonlinear processes. They have been widely used both theoretically and practically through simulations and real plant testing.

In this chapter we will present the history of state observers through looking over their principle development stages.

I-2 A survey of observers for nonlinear systems:

I-2-1 Definitions:

a. **Estimator:** The estimators, used in open loop, are based on the use of a copy of the model of a representation of the motor, the dynamics of an estimator depend on clean modes of the motor. Such an approach leads to the implementation of simple and fast algorithms, but sensitive to modeling errors and parametric variations during operation. Indeed, there is no feedback with real quantities to take into account these errors or disturbances.

b. Observer: An observer is a closed-loop estimator with independent system dynamics. It provides an estimate of an internal physical quantity of a given system, based solely on information concerning the inputs and outputs of the physical system with the input feedback of the error between the estimated outputs and the actual outputs, using a gain matrix to thus adjust the convergence dynamics of the error.

I-2-2 Luenberger observer:

The most obvious approach to estimate the state of a known system

$$\begin{cases} \dot{x}(t) = Ax(t) + Bu(t) \\ y(t) = Cx(t) \end{cases} \quad \text{(I-1)}$$

Is to create a copy:

$$\begin{cases} \dot{\hat{x}}(t) = A\hat{x}(t) + Bu(t) \\ \hat{y}(t) = C\hat{x}(t) \end{cases} \quad \text{(I-2)}$$

Whose state provides an estimate $\hat{x}(t)$ of the original system's state x(t). (We know the input u(t), so we can apply it to the copy as well as to the original system.). The simplicity of this method has a downside, however, if the initial state of the copy matches that of the original system exactly, i.e $\hat{x}(0)$ = x(0), the copy will provide the exact value of the state of the initial system.

Luenberger Observer possesses a relatively simple design that makes it an attractive general design technique.

Observer error is defined as

$$e = x - \hat{x} \quad \text{(I-3)}$$

Differentiating equation (I-5)

$$\dot{e} = \dot{x} - \dot{\hat{x}} \quad \text{(I-4)}$$

and substituting equations (I-1), (I-2) into (I-4) we get

$$\dot{e} = Ax + Bu - [A\hat{x} + Bu + L(y - \hat{y})] \quad \text{(I-5)}$$

or

$\dot{e} = A(x - \hat{x}) - LC(x - \hat{x})]$ (I-6)

$\dot{e} = (A - LC)e$ (I-7)

Solution of equation (I-7) is given by

$$e(t) = e^{A-LC}\, e(0)$$

The eigen-values of the matrix A−LC can be made arbitrary by appropriate choice of the observer gain L, when the pair [A,C] is observable (i.e. observability condition holds). So the observer error $e \to 0$ when $t \to \infty$.

Note:

The given system is observable if and only if the n × n matrix, $[C^T\ A^T\, C^T\ \ldots (A^T)^{(n-1)}\, C^T]$ is of rank n. This matrix is called the observability matrix.

I-2-3 The Kalman filter:

Within the significant toolbox of mathematical tools that can be used for stochastic estimation from noisy sensor measurements, one of the most well-known and often-used tools is what is known as the Kalman filter. The Kalman filter is named after Rudolph E. Kalman, who in 1960 published his famous paper describing a recursive solution to the discrete-data linear filtering problem [5].

The Kalman filter is a set of mathematical equations that provides an efficient computational means to estimate the state of a process, in a way that minimizes the mean of the squared error.The filter is very powerful in several aspects: it supports estimations of past, present, and even future states, and it can do so even when the precise nature of the modeled system is unknown.

I-2-4 The Extended Kalman Filter (EKF):

The Kalman filter addresses the general problem of trying to estimate the state of a discrete-time controlled process that is governed by a linear stochastic difference equation. But what happens if the process to be estimated and (or) the measurement relationship to the process is non-linear? Some of the most interesting and successful applications of Kalman filtering have been such situations. A Kalman filter that linearizes about the current mean and covariance is referred to as an extended Kalman filter or EKF [5].The basic approach for an EKF is to linearize the original nonlinear system and then implement a Kalman filter setup on the linearized system. Owing to the inherent flaws of linearization (basically taking only the first term of the Taylor series), EKFs are not optimal estimators, unlike Kalman filter.

That being said, EKF is one of the most widely used and practical estimator for nonlinear systems.

I-2-5 High gain observer:

A special role is played by the high-gain observers in which the error trajectory has an exponential decay rate that can be imposed arbitrarily fast by acting on a design parameter, appearing in the observer structure, typically known as high-gain parameter [6].

Historical overview:

The use of high-gain observers in feedback control appeared first in the context of linear feedback as a tool for robust observer design. In their celebrated work on loop transfer recovery [7], Doyle and Stein used high-gain observers to recover, with observers, frequency-domain loop properties achieved by state feedback. The investigation of high-gain observers in the context of robust linear control continued in the 1980s [8]. The use of high-gain observers in nonlinear feedback control started to appear in the late 1980s [9],Soon afterwards the technique was developed independently by two schools of researchers a French school lead by Gauthier, Hammouri [10], and others and a U.S. school lead by Khalil at Michigan State University. The work of Gauthier and others in the French school focused on deriving global results under global Lipschitz conditions. The work of Khalil and co-workers took a different turn after it was demonstrated by Esfandiari and Khalil [11].

I-3 Sensorless speed control:

Accurate speed information is always necessary to realize high performance and high-precision speed control of induction motor drives. Conventionally, a direct speed sensor, such as an encoder, is usually mounted to the motor shaft to measure its speed. The use of such direct speed sensors implies additional electronics, extra wiring, extra space, frequent maintenance and careful mounting which detracts from the inherent robustness and reliability of the drive. Also, it adds an extra cost and the drive system becomes expensive. For these reasons, the development of alternative indirect methods becomes an important research topic [12]. Therefore, there is a great interest in the research community to develop a high performance induction motor drive that does not require a direct speed sensor for its operation; in other words, to develop a sensorless speed induction motor drive. Many advantages are expected from sensorless speed induction motor drives such as reduced hardware complexity, low cost, reduced size, elimination of direct sensor wiring, better noise immunity, increased reliability, and less maintenance requirements. Speed-sensorless motor drives are also preferred in hostile environments and high-speed applications [13]. The positive features of speed sensorless systems introduces them as a preferable choice for the next generation of commercial motor drives, not only for induction machines but also include other electrical machines, such as switched reluctance motors (SRM) and permanent-magnet synchronous motors (PMSM) [14].

I-4 Speed estimation schemes of sensorless induction motor drives:

Recently, serious attempts to eliminate direct speed sensor of induction motor drives are reported. All these attempts employ motor terminal variables and its parameters in some way or another to estimate the speed. The question always arises is to which extent the method is successful without deteriorating the dynamic performance of the drive. Several methods have been recently, proposed for speed estimation of high performance induction motor drives.

I-4-1 Non Adaptive (Open Loop) Method:

a. **Direct Calculation method**:

In direct calculating method, value of rotor angle and speed are calculated directly this is done by converting the three phase varying quantities into two phase (d-q) reference frame. The voltage and current values are then calculated along both axis (i.e, d and q axis). This method is direct, simple and easy having quick dynamic response, because of this reason no observer is required. Although due to measurement noise, any stator current deviation will leads to calculation error in rotor angle and speed values. The massive problem of this method is parameter uncertainty of motor.

b. Method based on calculation of stator inductance:

The methods based on inductance calculation utilize the value of stator inductance for the mechanical values estimation like rotor position and its speed. This method can only be used for the motors whose salient values are large such as switched reluctance and interior motors. Stator inductance depends on rotor position and therefore position of rotor can be known by tracking the inductance value and rate of change of current [15]. A B Kulkarni described a method in which rotor position of motor is calculated by stator phase current and voltage [16]. This scheme is very useful for high saliency motor even at starting i.e. null speed where the motional EMF becomes zero. It is an open loop method, cannot be guaranteed about estimation correctness [17].

c. Back EMF integration:

The back EMF induced in motor is directly proportional to speed of rotor or its angle and therefore the back EMF can effectively estimates the rotor values. [18] In ref. [19], authors propose the determination of back EMF which is independent from voltage probes that makes the method cheaper with added advantage of increased systems reliability. Reference values are preferred over measured voltage. Due to the integrator drift problem calculation of back EMF is not done by the total flux linkage integration of the stator phase circuits. As this method is direct, it is very fast, straightforward and hence doesn’t require complex observer which makes it simple, robust, cost effective and accurate at higher frequencies. At low speed such as starting, the method gives poor accuracy, one of its major problems is the back EMF calculation as it is not easy to obtain.

I-4-2 Adaptive (Close Loop) Methods:

Close loop method is also called adaptive method which adapts the varying situation and behaves according to it. They do this by comparing the outputs of two different systems, real and mathematically modelled system. The main objective of these methods is to minimize the error between both systems. Some close loop methods are presented below:

a. Extended Kalman Filter:

The Kalman and Extended Kalman Filters (EKF) are stochastic type filters in which Kalman filter is used for linear system and EKF for nonlinear systems. These filters are works on least squarse variance estimation to calculate the state of dynamics nonlinear system. EKF is different from Luenberger observer on the basis of the selection of the gain matrix. The evaluation of EKF is very much dependent to the errors of flux linkage. However, EKF application has a major drawback, which is to be solved yet and that is the poor performance at lower speeds (-5 HZ). This method is insensitive to measurement noises and parameter inaccuracy. Computational time is also reduced by this method.

b. Model Reference Adaptive System (MRAS):

Unlike EKF, MRAS method compares a reference model value with an adjustable model value such as to reduce the error up to zero level. The Model Reference Adaptive System is based on a criterion known as Popov’s stability, is a very favourable technique, involved in calculation of speed and rotor resistance [20]. As the name suggests this system adapts the situation and adjusts the controlling parameters. An error vector represents the difference between the reference and adaptive model. Dynamic error will be expressed as the distinction between real and estimated value. As a consequence, the evaluated parameter vector will finally settle to its true value. High speed of adaptation is advantage of this system [21].

c. Sliding Mode Observer (SMO):

When two or more functions or control laws are chosen such that they move towards a common boundary and then slides on that boundary or surface known as sliding surface. The combined effect of both functions makes the system slide on the sliding surface. Sliding mode controller (SMC) can be designed by following steps, a surface has to be defined first.Then, control laws or functions need to be defined so as to have convergence of the systems towards the surface defined in above step. These functions make the system slide on that surface hence the name the sliding mode. Most of the discussed techniques suffer due to the variation occurred in motor parameters. Also, it is known that in a nonlinear system any variations of the operating point will lead to high errors in conventional linear estimators as they are not adaptive. SMO provides a better solution for estimating rotor position and its speed information. This observer has good robustness and can be easily implemented. In SMO, a sliding hyper-plane is chosen with the help of an error vector [22]-[23]. This hyper-plane is then utilized by making the system slide over it until it reaches minimum error. This technique assures zero error for the estimations of rotor values due to its robustness. In this method reconstructed back EMF is used for the estimation purpose. SMO is a demanding and popular technique due to its robustness property towards parameter variation. But the drawback of this technique is the poor performance at standstill or low speeds, because the value of back EMF is very low at low speeds. Another problem in this method is conflict between good steady state result and fast convergence rate. To ensure accurate speed and rotor angle values, the convergence rate will be compromised. The gain coefficient should be chosen such that to ensure good speed estimation results with better convergence [24].

I-4-3 Saliency and Signal Injection based Method:

Rotor voltage or back-EMF voltage is dependent on rotor speed so, the performance of rotor angle estimation is critically dependent on this. At lower speeds the value of this voltage becomes negligible which creates a problem in the estimation of rotor speed and its position. To overcome these problems non ideal property methods can be used [25]. To improve the performances better estimation of rotor value in the low speed range, different type of methods and techniques are discussed for sensorless speed control.

a. High Frequency (HF) injection method:

In this method high frequency signal is given as an excitation to the motor with a fundamental frequency signal. The magnitude of stator voltage and stator current are kept low in comparison to the fundamental values to reduce the extra losses introduced due to this signal. By injecting HF signal the current derivatives becomes large and hence the variation in inductance becomes prominent even at low speeds or at standstill. This method utilizes the saliency of the rotor and a magnetic pole [26] the advantage of this method is that it is insensitive to parameter inaccuracy and rotor position estimation error.

b. Low Frequency (LF) injection methods:

This method is also used for evaluating the rotor parameters in low speed range. This method does not have much effect of saliency of the rotor and hence can be used for non-salient motors, which is an added advantage. In LF injection method, a low frequency signal is superimposed on the reference d axis. This added signal will introduce ripples in back-EMF, and these ripples are utilized to estimate rotor speed and position. Originally this method was used for the speed estimation at low speeds for SMPM motor. Unlike HF injection method, in this method saliency is not necessary condition for estimation of rotor speed. The estimation process depends on the motor inertia (J). Another drawback of this estimator is slow dynamic response [27].

I-5 Speed estimation at low speed:

The key problem in sensorless field oriented control of alternating current (AC) drives is the accurate dynamic estimation of the stator flux vector during wide speed range operation using only terminal variables (currents and voltages). The main difficulty consists on state estimation at very low speeds where the fundamental excitation is low and the observer performances tend to be poor. The reasons are the observer sensitivity to model parameter variations, not modeled nonlinearities and disturbances, limited accuracy of acquisition signals, drifts and DC offsets. The main sources of poor speed estimation at low speed depends on three main problems [28 - 29]:

I-5-1 Data Acquisition Errors:

Data acquisition errors become significant at low speeds. That is because current sensors convert the machine currents to voltage signals which are subsequently digitized by A/D converters. Parasitic DC offset components are superimposed to analogous signals and appear as AC- components of fundamental frequency after their transformation to synchronous coordinates. They act as disturbances to the current control system, thus generating a torque ripple. The effect of unbalanced gains of the current acquisition channels is also another problem [30].

I-5-2 Voltage Distortion due to the pulse width modulation (PWM) Inverter:

The task of the power inverter is to produce the desired voltage on the stator winding using PWM controlled switches. Since the switching times of the existing transistors are not infinitely short, the necessary blanking time to avoid short circuiting the dc link during commutations must be introduced, which is also known as dead time. This small time delay is the most important cause of the inverter nonlinearity and introduces a magnitude and phase error in the output voltage vector. In addition to the dead time, there is also the finite voltage drop across the switch during the ON state, which introduces an additional error in the magnitude of the output voltage. Taking the inverter model into consideration enables a more accurate estimation of the stator flux linkage vector [30 - 31].

I-5-3 Stator Resistance Drop:

In the upper speed range, the resistive voltage drop is small as compared with the stator voltage; hence the stator flux vector and speed estimation can be made with good accuracy. At low speeds the stator frequency is also low. The stator voltage reduces almost in direct proportion, while the resistive voltage drop maintains its order of magnitude and becomes significant at low speed. The resistive voltage drop greatly influence the estimation accuracy of the stator flux vector and hence the speed estimation. On the other hand, considerable variations of the stator resistance are encountered when the motor temperature changes at varying load. These variations need to be tracked to maintain stability of flux estimation at low speed [29, 30]. Several methods to improve the low-frequency performance of the voltage model have been proposed. For example, the stator voltage and current can be measured more accurately, the voltage drop of the inverter can be compensated, and the stator resistance can be identified with an adaptive scheme, or the integrator itself can be modified to a low-pass filter [29 - 31].

I-6 Conclusion:

The invention of state observer has made a real revolution in the world of control system by leading to minimize the use of mechanical sensors in industry, this made designing and developing observers a very interesting approach for researchers. In this chapter we have outlined the history of state observer as well as the most popular sensorless speed control techniques. In the next chapter we are going to present the modeling and control of IM based on field oriented technique.

Chapter II

Modeling and Field Oriented Control of Induction motor

II-1 Introduction:

The induction motor is used today in many applications, including transportation (metro, trains, propulsion of ships), industry (machine tools), and home appliances. This large choice is due to the main advantage which lies in the absence of sliding electrical contacts, which leads to a simple and robust structure easy to build. It is used in a range of application power from few Watts to several Mega Watts [32].

The control of AC motors is one of the most classical and challenging topics of electrical engineering. In the present scenario with highly development of renewable wind energy utilization and electric vehicles, this topic continuously gains highest interest for industrial and academic communities. In addition, the fast improving calculation capability of modern microprocessors offer the possibility to consider more intelligent and powerful control strategies. In the field of high performance speed control techniques, two methods have been dominating the market since the last decades. One of them is field oriented control (FOC). FOC is a linear strategy, which uses linear controllers and a pulse width modulator (PWM) to generate the voltages applied to the motors. Due to its high performance at both steady and transient states, FOC has been widely applied in power electronics. However, it has cascaded structure with normally external speed proportional-integral (PI) controller and inner current PI controllers containing many parameters to be tuned.

In this chapter we are going to represent the induction motor mathematically in the form of different state models according to the choice of the frame and then simulate the FOC of IM and discuss the results.

II-2 Induction motor:

Modeling:

The induction motor comprises of a stator and a rotor, consisting of a stack of silicon steel sheets and having notches in which the windings are placed.

The stator is fixed, there are windings connected to the source, the objective being to obtain a distribution of magneto-motive forces and the most sinusoidal flux possible in the gap. The rotor is mounted on an axis of rotation. Depending on whether the rotor windings are accessible from the outside or are permanently closed on themselves, are defined two types of rotor: wound or squirrel cage, however, the last structure is often taken during the modelization as electrically equivalent to that of a wound rotor whose windings are short-circuited.

II-2-1 Simplifying assumptions:

Electrical machines are very complex systems, to take into account in their mathematical modeling all the physical phenomena that they contain. It is then essential to admit some simplifying conventional assumptions, which all the same do not alter the authenticity of the model of the motor on the framework of this project.

The induction motor considered will have the following assumptions:

- Symmetrical two poles, three phases windings.
- The slotting effects are neglected.
- The permeability of the iron parts is infinite.
- The flux density is radial in the air gap.
- Iron losses are neglected.
- The stator and the rotor windings are simplified as a single, multi-turn full pitch coil situated on the two sides of the air gap.

II-2-2 Mathematical model of induction motor:

a. Three-phase equations:

The stator consists of three windings distributed in space, and separated by an electrical angle of 120 °, the same for the rotor that is squirrel cage or formed of three coils.

The figure I.1 illustrates the arrangement of the stator and rotor windings:

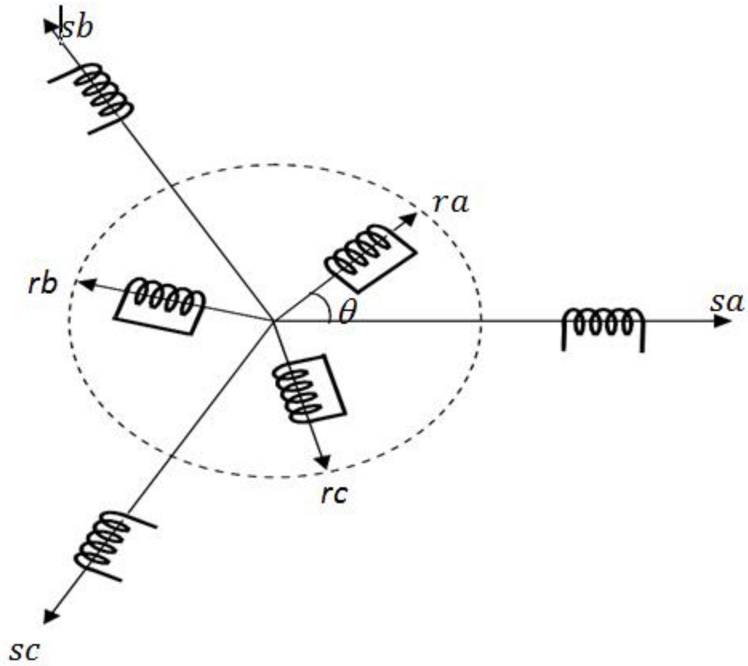

Fig.II-1: Spatial representation of IM windings

The stator voltages will be formulated in this section from the motor natural frame, which is the stationary reference frame fixed to the stator. In a similar way, the rotor voltages will be formulated to the rotating frame fixed to the rotor.

In the stationary reference frame, the equations can be expressed as follows:

$$[V_{sabc}]=[R_s][I_{sabc}]+\frac{d[\Psi_{sabc}]}{dt} \tag{II-1}$$

$$[V_{sabc}]=\begin{bmatrix}V_{sa}\\V_{sb}\\V_{sc}\end{bmatrix}$$

Similar expressions can be obtained for the rotor:

$$[V_{rabc}]=[R_r][I_{rabc}]+\frac{d[\Psi_{rabc}]}{dt} \tag{II-2}$$

$$[V_{rabc}]=\begin{bmatrix}V_{ra}\\V_{rb}\\V_{rc}\end{bmatrix}=\begin{bmatrix}0\\0\\0\end{bmatrix}$$

The rotor voltage equals 0 in the case of a squirrel cage.

The instantaneous stator and rotor fluxes linkage values per phase can be expressed as:

$$\begin{cases} [\Psi_{sabc}] = [L_{ss}][I_{sabc}] + [M_{sr}][I_{rabc}] \\ [\Psi_{rabc}] = [L_{rr}][I_{rabc}] + [M_{rs}][I_{sabc}] \end{cases} \qquad \text{(II-3)}$$

The magnitudes,$[I_{sabc}]$,$[I_{rabc}]$, $[\Psi_{sabc}]$,$[\Psi_{rabc}]$, are vectors of dimension 3x1 defined as follows:

$$[I_{sabc}] = \begin{bmatrix} I_{sa} \\ I_{sb} \\ I_{sc} \end{bmatrix} \qquad [I_{rabc}] = \begin{bmatrix} I_{ra} \\ I_{rb} \\ I_{rc} \end{bmatrix}$$

$$[\Psi_{sabc}] \begin{bmatrix} \Psi_{sa} \\ \Psi_{sb} \\ \Psi_{sc} \end{bmatrix} [\Psi_{rabc}] = \begin{bmatrix} \Psi_{ra} \\ \Psi_{rb} \\ \Psi_{rc} \end{bmatrix} \qquad [R_s] = R_s \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \qquad [R_r] = R_r \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

$$[L_{ss}] = \begin{bmatrix} ls & ms & ms \\ ms & ls & ms \\ ms & ms & ls \end{bmatrix} \qquad [L_{rr}] = \begin{bmatrix} lr & mr & mr \\ mr & lr & mr \\ mr & mr & lr \end{bmatrix}$$

$$[M_{sr}] = \mathrm{M} \begin{bmatrix} \cos(\theta) & \cos(\theta - \frac{4\pi}{3}) & \cos(\theta - \frac{2\pi}{3}) \\ \cos(\theta + \frac{2\pi}{3}) & \cos(\theta) & \cos(\theta - \frac{2\pi}{3}) \\ \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) & \cos(\theta) \end{bmatrix} \quad ; \quad [M_{rs}] = [M_{sr}]^{t}$$

b. Two-phase equations:

The two-phase model of IM is a transformation of the three-phase reference into a two-phase reference, which is in fact only a basic change in physical quantities (voltages, fluxes and currents), it leads to relations independent of the angle θ and the order reduction of the equations of the machine. The most well-known transformation for electrical engineers is Park (1929). [33]

Figure II-2 shows the direct axis d and the quadrature axis q of the Park referencial

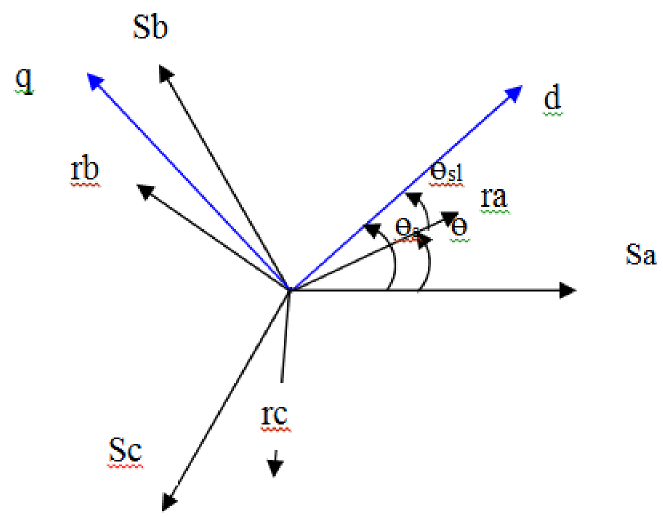

Fig.II-2: dq axes with respect to abc axes

$$\theta_s = \theta_{sl} + \theta \qquad \text{(II-4)}$$

The transform matrix of Park [T] is defined as follows:

$$[T] = \frac{2}{3}\begin{bmatrix} 1/2 & 1/2 & 1/2 \\ \cos(\theta i) & \cos(\theta i - \frac{2\pi}{3}) & \cos(\theta i - \frac{4\pi}{3}) \\ -\sin(\theta i) & -\sin(\theta i - \frac{2\pi}{3}) & -\sin(\theta i) \end{bmatrix}$$

We have chosen (2/3), for the unchanged values of amplitudes of the voltage, currents and fluxes

θi is the angle between the axis d and the reference axis in the three-phase system.

The direct transformation is then:

$$\begin{bmatrix} x_0 \\ x_d \\ x_q \end{bmatrix} = \frac{2}{3}\begin{bmatrix} 1/2 & 1/2 & 1/2 \\ \cos(\theta i) & \cos(\theta i - \frac{2\pi}{3}) & \cos(\theta i - \frac{4\pi}{3}) \\ -\sin(\theta i) & -\sin(\theta i - \frac{2\pi}{3}) & -\sin(\theta i) \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \qquad \text{(II-5)}$$

Where x represents the considered variables of the machine that are voltages, currents or fluxes.

The variable x_0 represents the homopolar component, added to make the transformation reversible; it is zero when the neutral is not connected. The inverse Park transform is necessary in order to return to the three-phase quantities, it is defined by:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & \cos(\theta i) & -\sin(\theta i) \\ 1 & \cos(\theta i - \frac{2\pi}{3}) & -\sin(\theta i - \frac{2\pi}{3}) \\ 1 & \cos(\theta i - \frac{4\pi}{3}) & -\sin(\theta i - \frac{4\pi}{3}) \end{bmatrix} \begin{bmatrix} x_0 \\ x_d \\ x_q \end{bmatrix} \quad \text{(II-6)}$$

Symbolized by the vector stator flux, the rotating field is the field created by the stator winding and which turns, in steady state, with the speed of synchronism. If we choose to fix the dq mark to the rotating field then we have:

$$\frac{d\theta_s}{dt} = \omega_s \Rightarrow \omega_{sl} = \omega_s - \omega = \omega_s - p\Omega \quad \text{(II-7)}$$

Where: ω_s : is the stator pulsation

ω : is the rotor pulsation

ω_{sl} :is the pulsation of the sliding

Ω: the mechanical speed, it is connected to the rotor pulsation by:

$\Omega = \omega / p$

The electrical equations of the IM in a Park reference linked to the rotating field are:

At the stator:

$$\begin{cases} v_{sd} = R_s i_{sd} + \frac{d\Psi_{sd}}{dt} - w_s \Psi_{sq} \\ v_{sq} = R_s i_{sq} + \frac{d\Psi_{sq}}{dt} + w_s \Psi_{sd} \end{cases} \quad \text{(II-8)}$$

At the rotor:

$$\begin{cases} v_{rd} = 0 = R_r i_{rd} + \frac{d\Psi_{rd}}{dt} - (\omega_s - p\Omega)\, \Psi_{rq} \\ v_{rq} = 0 = R_r i_{rq} + \frac{d\Psi_{rq}}{dt} + (\omega_s - p\Omega)\Psi_{rd} \end{cases} \quad \text{(II-9)}$$

With the fluxes equations as below:

For the stator:

$$\begin{cases} \Psi_{sd} = L_s i_{sd} + M i_{rd} \\ \Psi_{sq} = L_s i_{sq} + M i_{rq} \end{cases} \quad \text{(II-10)}$$

For the rotor:

$$\begin{cases} \Psi_{rd} = L_r i_{rd} + M i_{sd} \\ \Psi_{rq} = L_r i_{rq} + M i_{sq} \end{cases} \quad \text{(II-11)}$$

With

L_s = stator cyclic inductance

L_r = rotor cyclic inductance

M = Cyclic mutual inductance between stator and rotor.

The electromagnetic torque Te can take several forms:

$$\begin{cases} T_e = \frac{3}{2} p \left(\Psi_{sd} i_{sq} - \Psi_{sq} i_{sd}\right) \\ T_e = \frac{3}{2} p \left(\Psi_{rq} i_{rd} - \Psi_{rd} i_{rq}\right) \\ T_e = \frac{3}{2} L_m \left(\Psi_{rd} i_{sq} - \Psi_{rd} i_{sq}\right) \\ T_e = K_t \left(\Psi_{rd} i_{rq} - \Psi_{rq} i_{sq}\right) \end{cases} \quad \text{(II-12)}$$

where

$$K_t = p \frac{3}{2} \frac{L_m}{L_r}$$

p: The number of pairs of poles.

The mechanical rotation speed is deduced from the fundamental law of general mechanics (the sum of the torques to the shaft is equivalent to the inertial torque), so it is written:

$$J \frac{d\Omega}{dt} = T_e - T_L - f_v \, \Omega \quad \text{(II-13)}$$

Where:

J: Moment of inertia

T_L: Load torque

f_v: Friction coefficient

Now we will rewrite the equations of the motor in state model, taking the stator currents isd, isq and the rotor fluxes Ψ_{rd} , Ψ_{rq} as state variables.

We replace the rotor currents and stator fluxes from the equations II-10 and II-11

The rotor currents: $\begin{cases} i_{rd} = \frac{1}{L_r}\Psi_{rd} - \frac{M}{L_r} i_{sd} \\ i_{rq} = \frac{1}{L_r}\Psi_{rq} - \frac{M}{L_r} i_{sq} \end{cases}$ (II-14)

The stator fluxes: $\begin{cases} \Psi_{sd} = \left(L_{s-}\frac{M^2}{L_r}\right) i_{sd} + \frac{M}{L_r}\Psi_{rd} \\ \Psi_{sd} = \left(L_{s-}\frac{M^2}{L_r}\right) i_{sq} + \frac{M}{L_r}\Psi_{rq} \end{cases}$ (II-15)

$$\begin{cases} \frac{di_{sd}}{dt} = -\frac{1}{\sigma L_s}\left(R_s + \frac{R_r M^2}{L_r^2}\right) i_{sd} + \omega_s i_{sq} + \frac{1}{\sigma L_s}\left(\frac{R_r M}{L_r^2}\right)\Psi_{rd} + \frac{1}{\sigma L_s}\left(\frac{M}{L_r}\right)\omega\,\Psi_{rq} + \frac{1}{\sigma L_s} \\ \frac{di_{sq}}{dt} = -\omega_s i_{sd} - \frac{1}{\sigma L_s}\left(R_s + \frac{R_r M^2}{L_r^2}\right) i_{sq} - \frac{1}{\sigma L_s}\left(\frac{M}{L_r}\right)\omega\,\Psi_{rd} + \frac{1}{\sigma L_s}\left(\frac{R_r M}{L_r^2}\right)\Psi_{rq} + \frac{1}{\sigma L} \\ \frac{d\Psi_{rd}}{dt} = \frac{R_r M}{L_r} i_{sd} - \frac{R_r}{L_r}\Psi_{rd} + \omega_{sl}\Psi_{rq} \\ \frac{d\Psi_{rq}}{dt} = \frac{R_r M}{L_r} i_{sq} - \omega_{sl}\Psi_{rd} - \frac{R_r}{L_r}\Psi_{rq} \end{cases} \quad (II-16)$$

$\sigma = 1 - \frac{L_m^2}{L_r L_s}$ Is the dispersion coefficient.

This transformation is also called Clarke transformation, which is in fact a special case of Park transformation, when θs = 0, dq axis are now called **αβ** axis**.**

- The direct transformation is then as follows:

$$\begin{bmatrix} x_0 \\ x_\alpha \\ x_\beta \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/2 & 1/2 & 1/2 \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (II-17)$$

Where, x represents the voltages, currents or fluxes.

- The inverse transform is of the form:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/2 & 1 & 0 \\ 1/2 & -1/2 & \sqrt{3}/2 \\ 1/2 & -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} x_0 \\ x_\alpha \\ x_\beta \end{bmatrix} \quad (II-18)$$

From the previous definitions, the electrical equations of the motor are rewritten as follows:

For the stator: $$\begin{cases} v_{s\alpha} = R_s i_{s\alpha} + \frac{d\Psi_{s\alpha}}{dt} \\ v_{s\beta} = R_s i_{s\beta} + \frac{d\Psi_{s\beta}}{dt} \end{cases} \qquad \text{(II-19)}$$

$$\begin{cases} \Psi_{s\alpha} = L_s i_{s\alpha} + M i_{r\alpha} \\ \Psi_{s\beta} = L_s i_{s\beta} + M i_{r\beta} \end{cases} \qquad \text{(II-20)}$$

For the rotor: $$\begin{cases} v_{r\alpha} = 0 = R_r i_{\alpha} + \frac{d\Psi_{r\alpha}}{dt} + \omega\,\Psi_{r\alpha} \\ v_{r\beta} = R_r i_{r\beta} + \frac{d\Psi_{r\beta}}{dt} - \omega\,\Psi_{r\beta} \end{cases} \qquad \text{(II-21)}$$

$$\begin{cases} \Psi_{r\alpha} = L_r i_{r\alpha} + M i_{s\alpha} \\ \Psi_{r\beta} = L_r i_{r\beta} + M i_{s\beta} \end{cases}$$

Now we will rewrite the equations of the motor in state model, taking the stator currents $i_{s\alpha}$, $i_{s\beta}$ and the rotor fluxes $\Psi_{r\alpha}$, $\Psi_{r\beta}$ as state variables.

$$\begin{cases} \frac{di_{s\alpha}}{dt} = -\frac{1}{\sigma\, L_s}\left(R_s + \frac{1}{T_r}\frac{M^2}{L_r}\right) i_{s\alpha} + \frac{1}{\sigma\, L_s}\left(\frac{M}{L_r}\right)\frac{1}{T_r}\,\Psi_{r\alpha} + \frac{1}{\sigma\, L_s}\,(M)\omega\,\Psi_{r\beta} + \frac{1}{\sigma\, L_s}\, v_{s\alpha} \\ \frac{di_{sq}}{dt} = -\frac{1}{\sigma\, L_s}\left(R_s + \frac{1}{T_r}\frac{M^2}{L_r}\right) i_{s\beta} - \frac{1}{\sigma\, L_s}\left(\frac{M}{L_r}\right)\omega\,\Psi_{r\alpha} + \frac{1}{\sigma\, L_s}\left(\frac{M}{L_r}\right)\frac{1}{T_r}\,\Psi_{r\beta} + \frac{1}{\sigma\, L_s}\, v_{s\beta} \\ \frac{d\Psi_{r\alpha}}{dt} = \frac{R_r M}{L_r}\, i_{sd} - \frac{R_r}{L_r}\,\Psi_{rd} + \omega_{sl}\Psi_{rq} \qquad \text{(II} - 22) \\ \frac{d\Psi_{r\beta}}{dt} = \frac{R_r M}{L_r}\, i_{sq} - \omega_{sl}\,\Psi_{rd} - \frac{R_r}{L_r}\Psi_{rq} \end{cases}$$

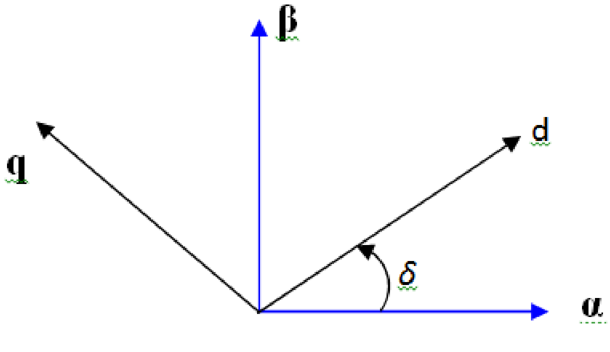

Fig.II-3 position of the landmarks

The transformation is then as follows:

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = [\boldsymbol{P}(\delta)] \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \qquad \text{(II-24)}$$

And conversely

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = [\boldsymbol{P}(-\delta)] \begin{bmatrix} x_d \\ x_q \end{bmatrix} \quad \text{(II-25)}$$

With: $[\boldsymbol{P}(\delta)] = \begin{bmatrix} cos(\delta) & sin(\delta) \\ -sin(\delta) & cos(\delta) \end{bmatrix}$ (II-26)

II-2-3 Simulation results:

Simulation of the IM model is performed using Matlab/Simulink.

The simulation will be done in the frame (d, q) for a nominal load test after a no-load start. Supply voltages are assumed to be perfectly sinusoidal with equal and constant amplitudes.

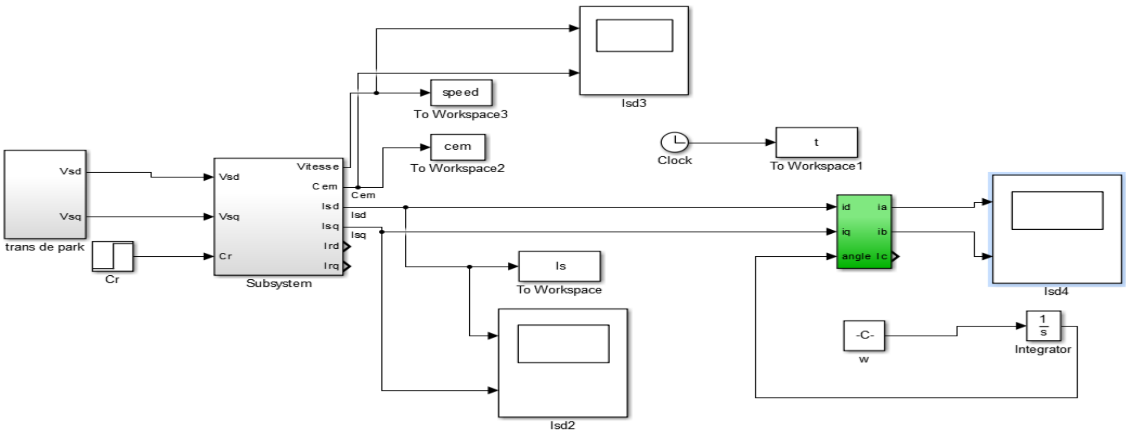

Fig.II-4 induction motor simulation diagram

The IM motor parameters:

F=50 Hz, Rs=4.85 ohm, Rr=3.805 ohm, Ls=0.274 H, Lr=0.274 H

p=2, M=0.258 H, f=0.008 N.s/rad, j=0.031 Kgm^2

a. No-load test

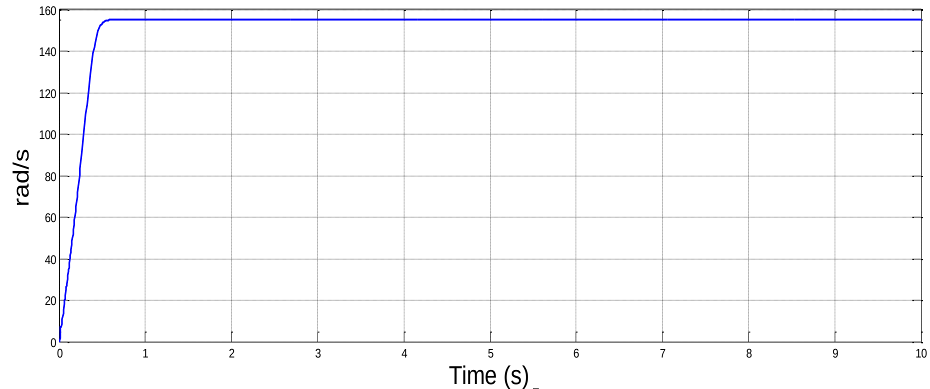

Fig.II-5 speed rotor

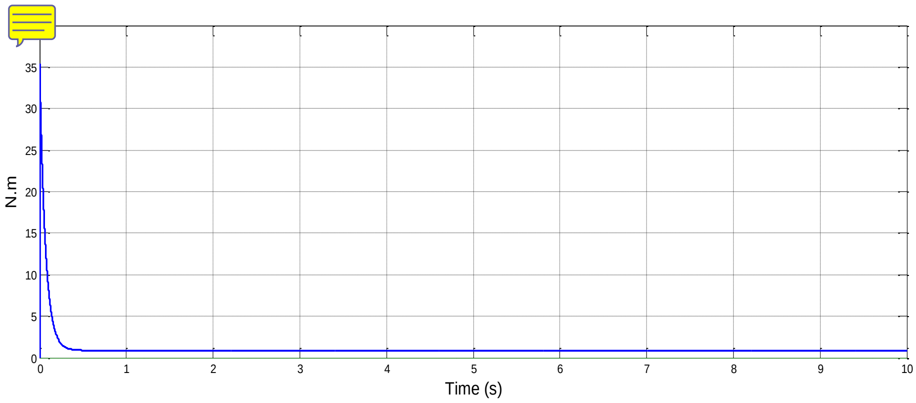

Fig.II-6 electromagnetic torque

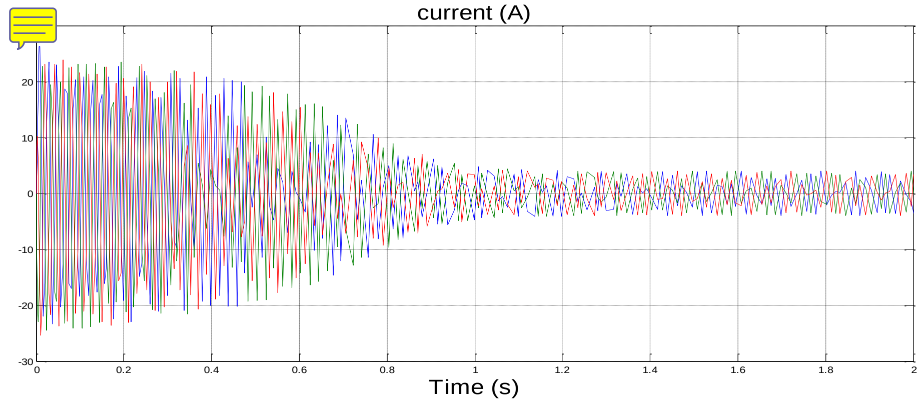

Fig.II-7 satatoric currents (Is(a,b,c))

- Fig.II-5 shows that the speed increases linearly then stabilizes at a value almost identical to the speed of synchronism.
- Fig.II-6 shows that the electromagnetic torque reaches a maximum value at the startup, and then it decreases to its final value in the case of no-load.
- Fig.II-7 shows that the current reaches a maximum value at the startup (start peak), and then decreases to its nominal value.

b. Load test: load applied at time t = 1.5s. / TL = 5 N.m

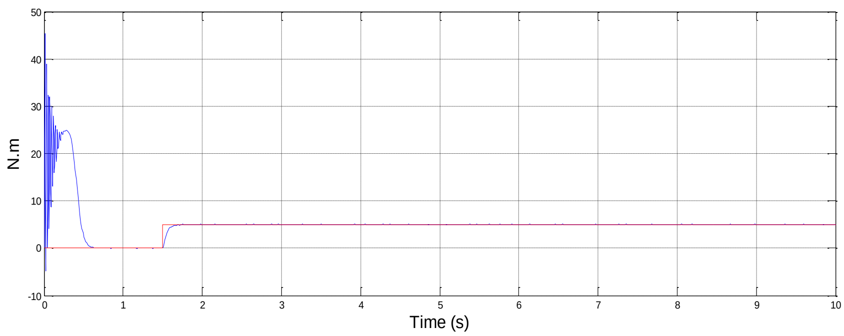

Fig.II-8 electromagnetic torque (blue) and load torque (red)

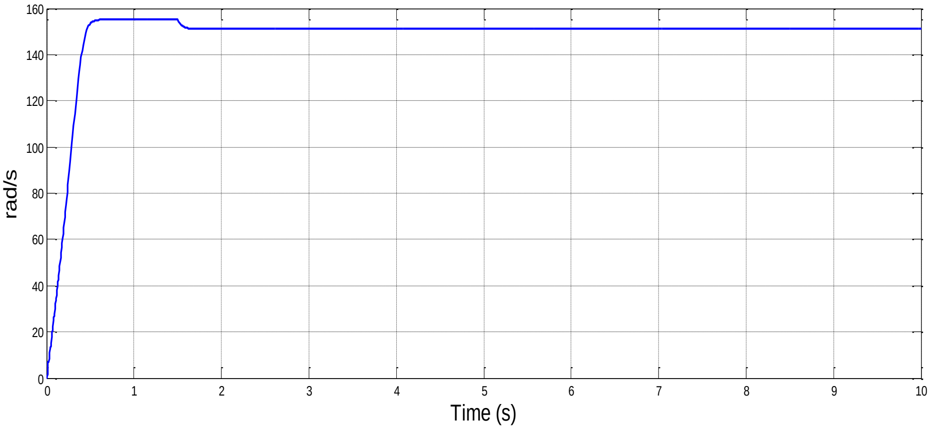

Fig.II-9 speed rotor

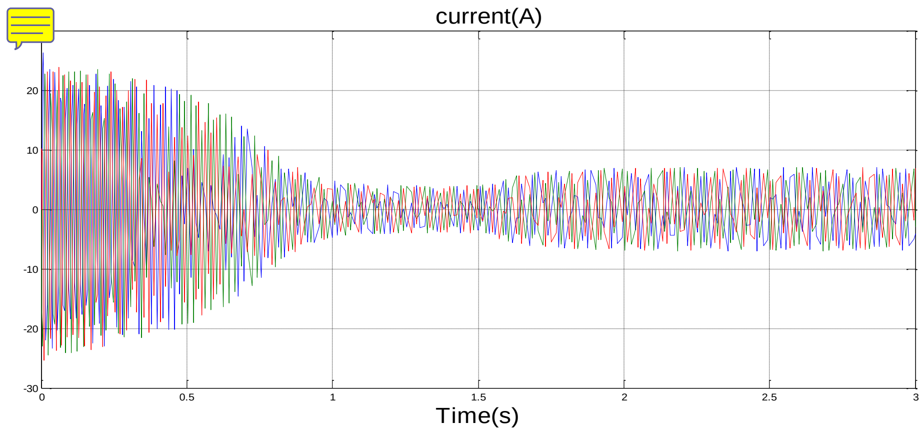

Fig.II-10 stator currents (Is(a,b,c)) (no load)

- Fig.II-8 and Fig.II-9 show the results for load test, in transient mode the torque and the speed behaves as the test case at no-load, At the moment when the load is applied (t=1.5s ➔ T_L = 5 N.m) it is noted that the speed decreases, and the electromagnetic torque is stabilized at the value that corresponds to the load considered.
- Fig.II-10 shows the currents which are the same as no load test except when we applied the load they increase (t=1.5s ➔ T_L = 5 N.m)

II-3- Field oriented control (FOC):

II-3-1 Principle of the field oriented control:

The principle of the FOC is that the torque and the flux of the motor are controlled separately in similarity to the separately excited DC motor, where the stator currents are transformed into a rotating reference system aligned with the rotor flux vector, stator or gap, to produce components along the d axis (flux control) and along the q axis (torque control).

II-3-2 Types of FOC:

a. Direct field oriented control (DFOC):

This type of control requires knowledge of the module and the phase of the flux (rotor in our case) at any time. A first method is to directly measure the flow of the machine using sensors positioned in the air gap and to deduce the amplitude and phase.

The sensors are subjected to extreme conditions (temperature, vibration, etc.) Moreover this measurement is tainted by noise depending on the speed caused by the notches, which requires the variable frequency filtering, this measure allows to design a completely decoupled field oriented control (flow and torque) against the installation of flow sensors that increases the cost of manufacturing, for this reason we introduce the estimate (open loop) or observation (closed loop) of the flux from classical measurements (currents, voltages, speed) [34]. As part of this project the indirect field oriented control will be processed.

b. Indirect rotor field oriented control (IRFOC):

Indirect rotor field oriented control does not do flux tuning, flux sensors, estimators and observers are not required. We therefore have no knowledge of the modulus and the phase of the rotor flux, this requires a measurement of the rotor position. This control is simpler but obviously at lower performance compared to the direct control, this is due to the sensitivity of this type of control for changes in the rotor time constant. The advantage is that this control requires little computing time in the microprocessor [35].

In the case of indirect control, the flow is not regulated (neither measured nor estimated). This is given by the setpoint and oriented from the angle θs which is obtained from the stator pulsation ωs. The latter is the sum of the rotor pulsation ωsl estimated and the mechanical pulse P.Ω measured. So this method eliminates the need to use a sensor or an observer of the gap flux.

II-3-3 Equations of IRFOC:

Remembering that the equations of stator and rotor voltage in Reference linked to the rotating field are the following: II-8 and II-9

At the stator:
$$\begin{cases} v_{sd} = R_s i_{sd} + \frac{d\Psi_{sd}}{dt} - w_s \Psi_{sq} \\ v_{sq} = R_s i_{sq} + \frac{d\Psi_{sq}}{dt} + w_s \Psi_{sd} \end{cases}$$

At the rotor:
$$\begin{cases} v_{rd} = 0 = R_r i_{rd} + \frac{d\Psi_{rd}}{dt} - (\omega_s - p\Omega)\, \Psi_{rq} \\ v_{rq} = 0 = R_r i_{rq} + \frac{d\Psi_{rq}}{dt} + \left(\omega_s - p\Omega\right)\Psi_{rd} \end{cases}$$

The implementation of field oriented control is based on the orientation of the rotating axis dq, such that the axis d coincides with the direction of ψr

The orientation of the magnetic flux along the direct axis leads to the cancellation of its quadrature component, we then have:

$$\begin{cases} \Psi_{rq} = 0 \\ \Psi_{rd} = \Psi_r \end{cases} \qquad \text{(II-27)}$$

The equations of the rotor voltages become:

$$\begin{cases} v_{rd} = 0 = R_r i_{rd} + \frac{d\Psi_r}{dt} \\ v_{rq} = 0 = R_r i_{rq} + \left(\omega_s - p\Omega\right)\Psi_r \end{cases} \qquad \text{(II-28)}$$

And the stator fluxes:

$$\begin{cases} \Psi_{sd} = L_s \sigma i_{sd} + \frac{M}{L_r}\Psi_{rd} \\ \Psi_{sq} = L_s \sigma i_{sq} \end{cases} \qquad \text{(II-29)}$$

So the stator voltages become:

$$\begin{cases} v_{sd} = R_s i_{sd} + L_s\sigma \frac{di_{sd}}{dt} + \frac{M}{L_r}\frac{d\Psi_{rd}}{dt} - w_s L_s \sigma i_{sq} \\ v_{sq} = R_s i_{sq} + L_s\sigma \frac{di_{sq}}{dt} + w_s \frac{M}{L_r}\Psi_{rd} + w_s L_s \sigma i_{sd} \end{cases} \quad \text{(II-30)}$$

Estimation of w_s and θ_s :

In the IRFOC, the stator pulsation is determined indirectly from the measurement of the mechanical speed and the following relation [36]:

$$w_{sl} = \frac{M}{L_r} + \frac{i_{sq}}{\Psi_{rd}} \quad \text{(II-31)}$$

From II-7 we obtain

$$w_s = p\Omega + \frac{M}{L_r} + \frac{i_{sq}}{\Psi_{rd}} \quad \text{(II-32)}$$

So θ_s is writing as below:

$$\theta_s = \int w_s . dt \quad \text{(II-33)}$$

Expression of electromagnetic torque:

The new equation of electromagnetic torque is:

$$T_e = k_t \Psi_{rd} i_{sq} \quad \text{(II-34)}$$

Note that the speed given by the expression in (II-13) remains unchanged. The equation of IM becomes:

$$\begin{aligned} v_{sd} &= \sigma L_s \frac{di_{sd}}{dt} + \left(R_s + \frac{R_r L_m^2}{L_r^2}\right) i_{sd} - \sigma L_s \omega_s i_{sq} - \left(\frac{R_r L_m}{L_r^2}\right)\Psi_r \\ v_{sq} &= \sigma L_s \frac{di_{sq}}{dt} + \left(R_s + \frac{R_r L_m^2}{L_r^2}\right) i_{sq} + \sigma L_s \omega_s i_{sd} + \left(\frac{R_r L_m}{L_r^2}\right)\Psi_r \\ \omega_s &= p\Omega + \frac{M}{L_r} + \frac{i_{sq}}{\Psi_{rd}} \\ J\frac{d\Omega}{dt} &= T_e - T_L - f\Omega \\ T_{em} &= k_t \Psi_{rd} i_{sq} \end{aligned} \quad (\text{II}.35)$$

We see in these equations that v_{sd} and v_{sq} depend on currents of the two axes chosen as states variables i_{sd} and i_{sq}, so they influence flux and torque. It is therefore essential decoupling the coupled terms [37] [38].

II-3-4 Decoupling by compensation:

Different decoupling techniques exist, decoupling by state feedback, static decoupling or compensation decoupling, which we will present now [38].

The purpose of the compensation is decoupling the axes d and q. This decoupling makes it possible to write the equations of the motor and the control part in a simple way and thus to easily calculate the coefficients of the regulators. Considering a long dynamic of low speed flux $\left(\frac{d\Psi_{rd}}{dt} = 0\right)$ compared to the currents [39] [40].

s is the operator of Laplace:

So the voltage equation becomes:

$$\begin{cases} v_{sd} = (R_s + sL_s\sigma)i_{sd} - w_sL_s\sigma i_{sq} \\ v_{sq} = (R_s + sL_s\sigma)i_{sq} + w_s\frac{M}{L_r}\Psi_{rd} + w_sL_s\sigma i_{sd} \end{cases} \quad \text{(II-36)}$$

The new control variable is written as below:

$$\begin{cases} v^*{}_{sd} = (R_s + sL_s\sigma)i_{sd} = v_{sd} + w_sL_s\sigma i_{sq} = v_{sd} + e_{sd} \\ v^*{}_{sq} = (R_s + sL_s\sigma)i_{sq} = v_{sq} - \left(w_s\frac{M}{L_r}\Psi_{rd} + w_sL_s\sigma i_{sd}\right) = v_{sq} + e_{sq} \end{cases} \quad \text{(II-37)}$$

With: * refer to the control variable or reference

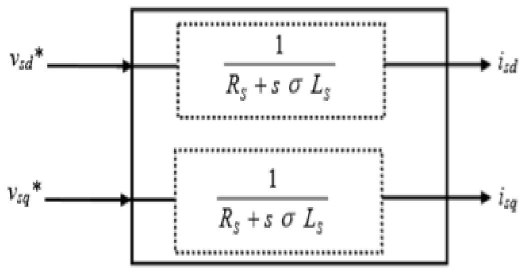

Fig.II-11 the new control variable

The voltages v_{sd}, v_{sq} are then reconstituted from the voltages $v^*{}_{sd}$, $v^*{}_{sq}$

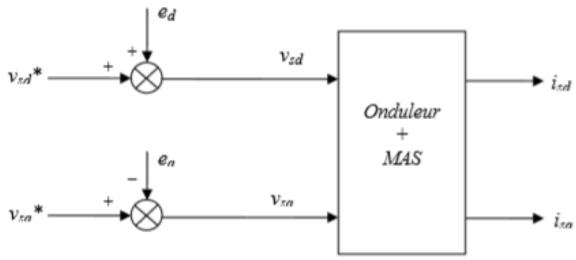

Fig.II-12 reconstitution of the voltages v_{sd}, v_{sq}

II-3-5 Simulation of FOC for IM:

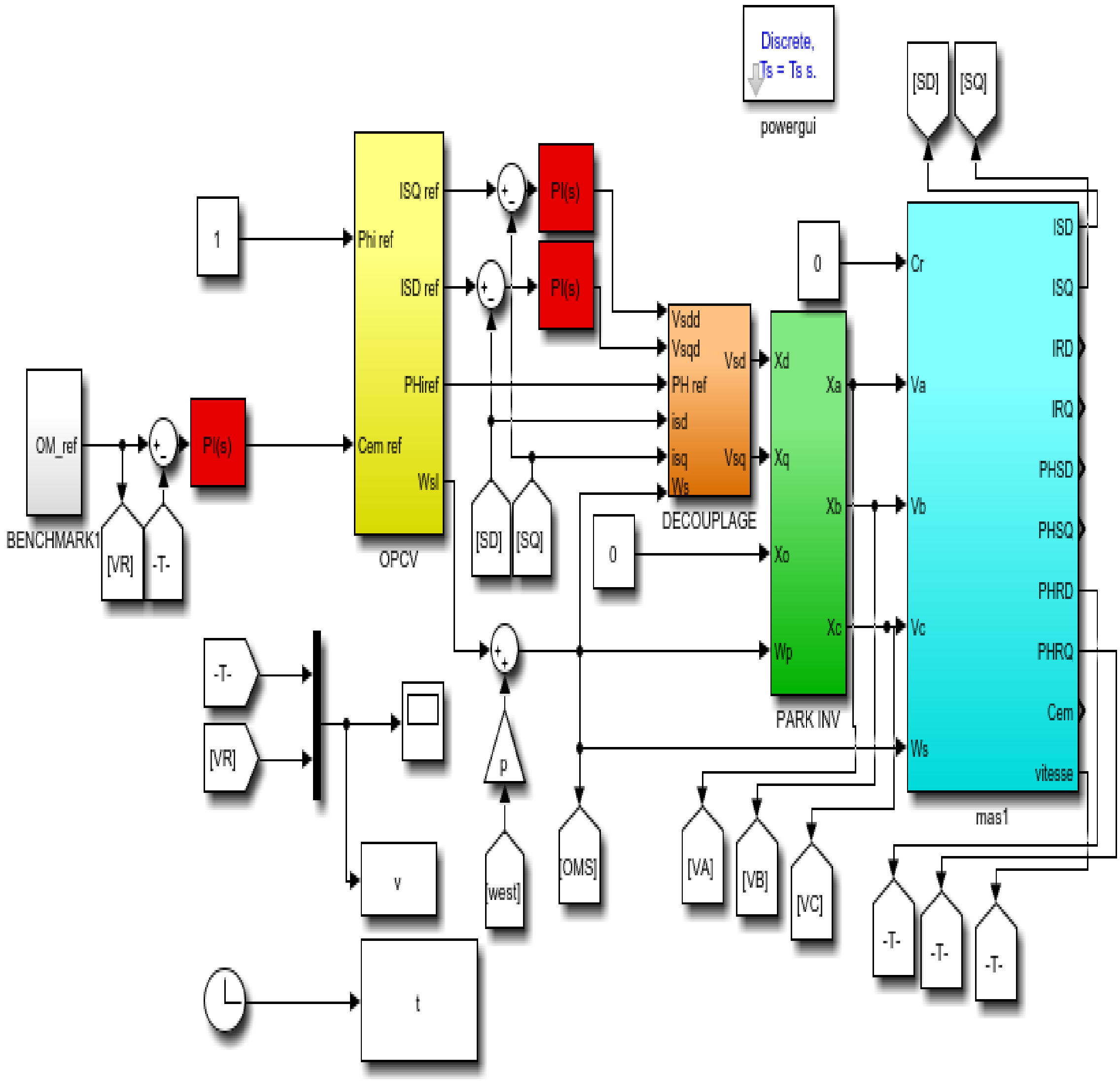

Fig.II-13 block diagram of the FOC in MATLAB Simulink.

a. No load test :

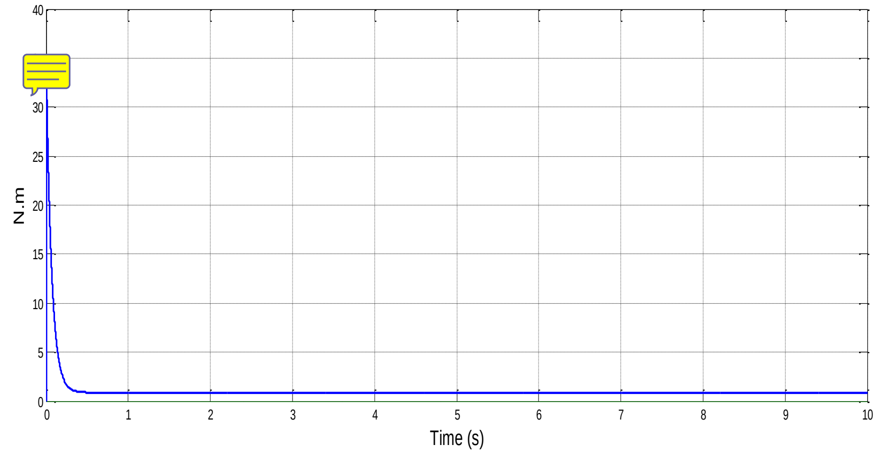

Fig.II-14 Electromagnetic torque

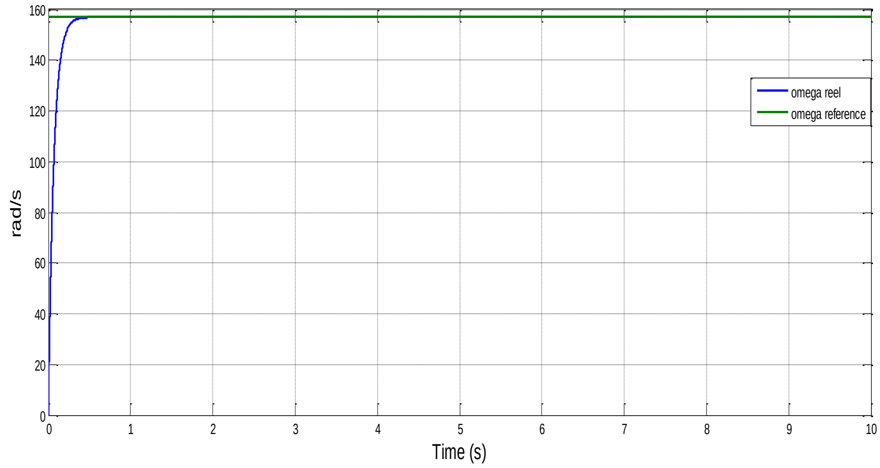

Fig.II-15 Rotor speed (blue) with reference speed (green)

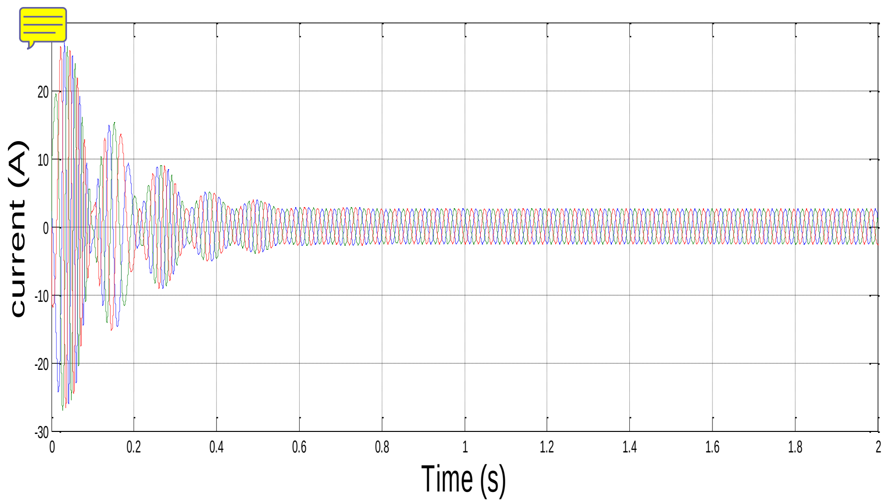

Fig.II.16 Statoric Currents (Is(a,b,c))

b.Test with load: t=3 s, T_L = 8 N.m

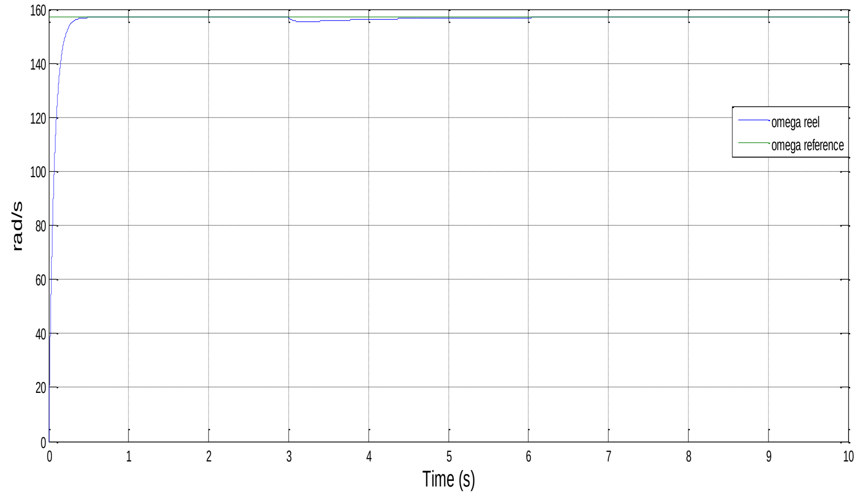

Fig.II-17 rotor speed (bule) and the reference speed (green)

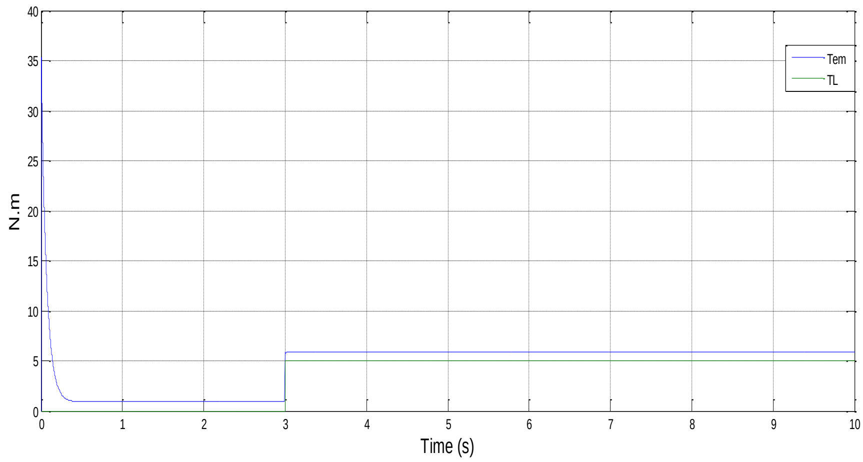

Fig.II-18 Electromagnetic torque (blue) and load torque (green)

Simulation results show that for load application (Fig.II.8), the quantities such as speed, torque, fluxes and currents are influenced by this load but the system is perfectly controlled.

c -Reversing the direction of rotation : (speed from +157 rad/s to -157 rad/s)

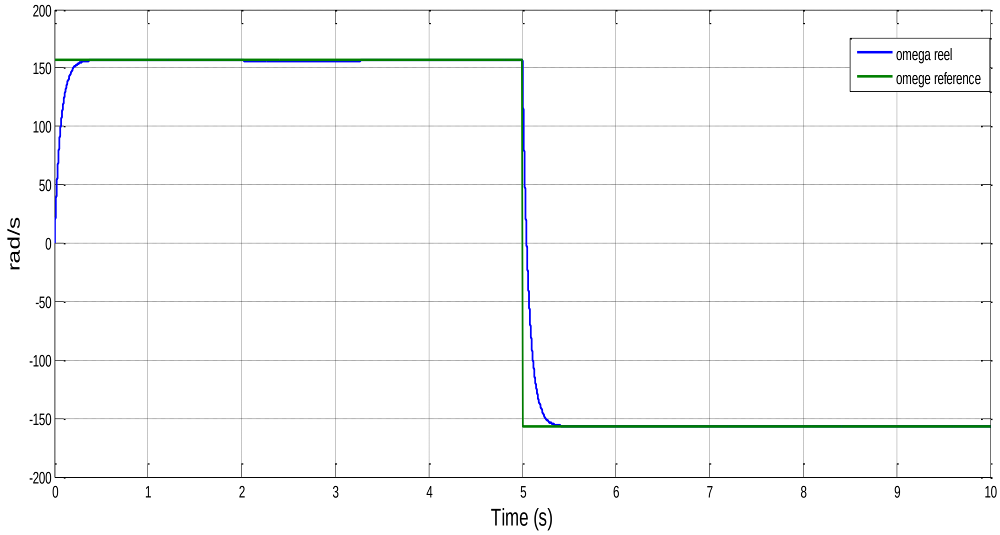

Fig.II-19 Rotor speed (blue) with Reference speed (green)

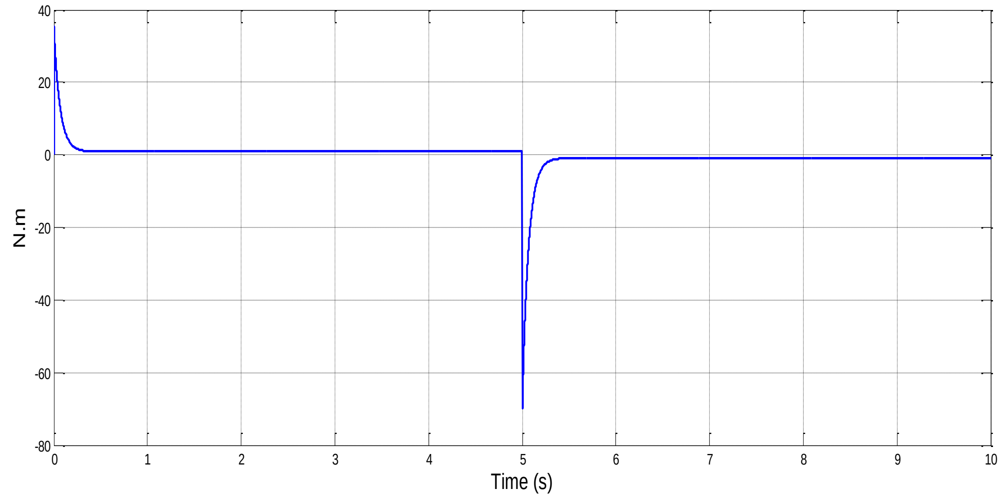

Fig.II-20 Electromagnetic torque (speed inversely)

Following this test of inverting the direction of rotation (+157 rad / s at -157 rad / s) with a load, results show that the response in speed follows perfectly the instruction with the same time of response and a null error. The torque follows the load value imposed with certain peaks when switching from one direction of rotating to another. The coupling between the flux and the torque is intact during this test.

II-4 Conclusion:

In this chapter we have started by describing our motor and its components, and then we have proposed simplification hypotheses which have helped to build several mathematical models for the motor.

In the second part we have chosen the field oriented control as a control technique for our induction motor, we have presented its basic concepts. This control ensures the necessary decoupling, allowing to separate the control of the flux and that of the torque. Simulation results have shown the robustness of the IRFOC control against the variations of motor parameters and rotor speed. In order to overcome the disadvantages of sensors, in the next chapter we will present sensorless speed control system.

Chapter III

Sensorless speed Control of induction motor

III-1 Introduction:

The industry world is interested in reducing the number of sensors because of the complexity and high cost of their installation (additional wiring, maintenance), the cost of the sensor itself being just one element among others. Most of the control techniques of induction motor as the field oriented control requires the measurement not only of stator currents (possibly stator voltages) but also of mechanical speed. Moreover, the load torque is a measurable disturbance but the price of the sensor makes this measurement unrealistic. The control without mechanical sensors (speed, position and load torque) has become a major concern. There are several nonlinear techniques in the literature based on the motor model for determining the mechanical variables of the induction motor. In general, using the state model of the induction model, the mechanical speed can be estimated (using state observer) from stator voltages and currents [41] and [42].

It is well known that there are no systematic methods for designing observers for a given nonlinear system. However, several observation methods are available depending on specific characteristics of the nonlinear system studied. In particular the nonlinear system considered can be seen as an interconnection between several subsystems, where each subsystem satisfying certain conditions, an observer can be synthesized [43][44].

In this chapter we will present and construct an interconnected high gain observer and then connect it to the field oriented control of induction motor using MATLAB SIMULINK.

III-2 Observability study of induction motor without mechanical sensor:

The observability problem of induction motor without mechanical sensors has been studied by several authors [46.47]. In [48] are presented some cases in which the induction motor is observable and unobservable. The problem was to characterize the condition in which the states of induction motor can be observed only from measurable variables (currents and voltages). It follows that when the fluxes $\Psi_{r\alpha}$, $\Psi_{r\beta}$ and the speed Ω are constant even when we use high derivatives of currents, the induction motor observability can not established. Hence, this is sufficient and necessary condition for lost of observability.

This operating case matches the following physical interpretation.

1- When the fluxes are constant ($\dot{\Psi}_{r\alpha} = \dot{\Psi}_{r\beta} = 0$), or equivalently, the excitation stator pulsation is zero ($\boldsymbol{w_s}$ = 0), it implies that $p\Omega + (R_r T_e / p\Psi^2{}_{rd}) = w_s = 0$ or T_e= $-K\Omega$ and K= $p^2\Psi^2{}_{rd}/\ R_r$.

2- If the speed of the motor is constant, then T_e=($f\Omega + T_L$) = $-K\Omega$

The last equation defines the unobservability curve in the map (T_L, Ω) with M $= (p^2\Psi^2{}_{rd}/\ R_r\ ) + f$

Then, taking into account this observability problem, in the next section, the design of the interconnected high gain observer to estimate the mechanical and flux variables

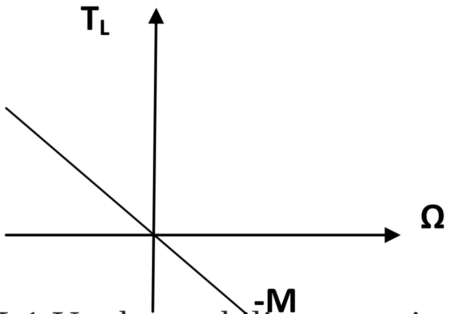

Fig.III-1 Unobservability curve in the map

of the induction motor without mechanical sensors will be presented.

III-3 Methodology:

It is well know that there is no general method to design an observer for general nonlinear control systems. However, the nonlinear system maybe seen as an interconnection between several subsystems. The idea to construct the interconnected observers is to design an observer for the whole system, from the separate synthesis of observers for each subsystem, assuming that, for each of these separate observers, the state of the other subsystem is available.

III-4 Interconnected high gain observer:

If a system does not completely satisfy any of the known properties it may satisfies some of them partly, In other words it may be seen as an interconnection between several subsystems where each of these subsystems satisfies some required properties for an observer to be designable. The idea is then to design an observer for the whole system from separate designs for each subsystem and consequently the problem becomes analogous to some extent to the so called separation principle for observer based control, in that case a controller is designed assuming that all states are available and separately an observer is designed assuming that the control is known for linear systems, the achieved controller based on the estimates of the states given by the achieved observer is proved to be still stabilizing, however, this is no longer guaranteed for general nonlinear systems.

III-5 Interconnected high gain observer for induction motor:

The given nonlinear system represented by:

$$\begin{cases} \dot{X} = A(u,y) + g(u,y,X) \\ \quad\quad y = CX \end{cases} \quad \text{(III-1)}$$

Where $X \in R^n$ constitutes the state vector, $u \in R^m$ is the control input, $y \in R^p$ is the measured output and the matrix A(u,y) is bounded.

Systems of form (III-1) belong to the class of nonlinear state affine systems with a nonlinear term $g(X, u, y)$. This term is in the triangular form in order not to destroy the observability property of the system. A particular case is when the term g depends only on the inputs and outputs, that is, $g(u, y)$, then it is called an input– output injection.

From now on, this class of nonlinear systems will be considered in order to design an observer using the property of inputs' persistency [45].

The model of the induction motor ($IM_{\alpha\beta}$) can be described by:

$$\mathbf{IM\alpha\beta} \begin{cases} \Psi_{r\alpha} = -a\Psi_{r\alpha} - p\Omega\Psi_{r\beta} + aM_{sr}i_{s\alpha} \\ \Psi_{r\beta} = -a\Psi_{r\beta} + p\Omega\Psi_{r\alpha} + aM_{sr}i_{s\beta} \\ i_{s\alpha} = b\left(a\Psi_{r\alpha} + p\Omega\Psi_{r\beta}\right) - \gamma i_{s\alpha} + m_1 u_{s\alpha} \\ i_{s\beta} = b\left(a\Psi_{r\beta} - p\Omega\Psi_{r\alpha}\right) - \gamma i_{s\beta} + m_1 u_{s\beta} \\ \Omega = m\left(\Psi_{r\alpha}i_{s\beta} - \Psi_{r\beta}i_{s\alpha}\right) - c\Omega - \frac{1}{J}T_L \end{cases} \quad \text{(III-2)}$$

III-6 Advantages of the high gain observer:

The advantages of HGO in comparison to other observers can be summarized as follows:

1- High gain observers are relatively simple to design as you do not need to solve complex differential equations nor use complicated formulas. Determining a suitable value for the gain is typically done through experimentation.

2- For large class of nonlinear systems, they can provide global or semi-global stability results for large class of systems. This means that their use can provide stability guaranteed for any arbitrarily chosen initial conditions.

3- They can be relatively fast.

4- They can be robust to modeling uncertainty and external disturbances.

5- For nonlinear systems represented in the normal form, it was shown that when high gain observers are used in output feedback control they can recover the performance of the state feedback controller, in other words, assuming that all the states are measured, you can design a controller that satisfies some performance specifications such as some desired overshoot and settling time. Then, when the states are provided by the high-gain observer, the output feedback controller can provide a very similar performance to the state feedback case.

III-7 Disadvantages of the high gain observer:

The disadvantages of high gain observer can be summarized as follows:

1- They are sensitive to measurement noise. Although there have been some results in the literature that tackled this problem, solving this problem still deserves more research.

2- They suffer from the peaking phenomenon, where, due to the high gain, there is an initial sharp spike in the response of the state estimates. This phenomenon can cause instability for some types of systems. One practical solution to this problem is to use saturation outside the operating range of the system states.

III-8 Application:

Here, a HGO interconnected to an estimator (E) is designed for the class of state affine system (III-3). The model of induction motor (III-2) can be rewritten in an interconnected compact form as

$$\Sigma_{1HGO}:\begin{cases}\dot{\tilde{X}}_1 = A_{1_{HGO}}(u,y,X_2)\tilde{X}_1 + g_1(u,y,X_2,\tilde{X}_1) \\ y_1 = C_1\tilde{X}_1\end{cases} \qquad \text{(III-3)}$$

And

$$\Sigma_2:\begin{cases}\dot{X}_2 = A_2\, X_2 + g_2(u,y,\tilde{X}_1,X_2) \\ y_2 = C_2X_2\end{cases} \qquad \text{(III-4)}$$

$$A_{1_{HGO}}(u,y,X_2) = \begin{pmatrix} 0 & bp\Psi_{r\beta} & 0 \\ 0 & 0 & -\frac{1}{J} \\ 0 & 0 & 0 \end{pmatrix}$$

$$g_1(u,y,X_2,\tilde{X}_1) = \begin{pmatrix} -\gamma i_{s\alpha} + ab\,\Psi_{r\alpha} + m_1u_{s\alpha} \\ m(\Psi_{r\alpha}i_{s\beta} - \Psi_{r\beta}i_{s\alpha}) - c\Omega \\ 0 \end{pmatrix}$$

$$A_2 = \begin{pmatrix} -\gamma & 0 & ab \\ 0 & -a & 0 \\ -ab & 0 & -a \end{pmatrix}$$

$$g_2(u,y,\tilde{X}_1,X_2) = \begin{pmatrix} -bp\Omega\Psi_{r\alpha} + m_1u_{s\beta} \\ -p\Omega\Psi_{r\beta} + aM_{sr}i_{s\alpha} \\ p\Omega\Psi_{r\alpha} + a(M_{sr} + b)\, i_{s\beta} \end{pmatrix}$$

Where $\tilde{X}_1 = (\tilde{x}_{11},\tilde{x}_{12},\tilde{x}_{13})^T$ with $\tilde{x}_{11} = i_{s\alpha}$, $\tilde{x}_{12} = \Omega$,$\tilde{x}_{13} = T_L$;

$C_1 = (1 \quad 0 \quad 0)$;

$X_2 = (x_{21},x_{22},x_{23})^T$ with $x_{21} = i_{s\beta}$,$x_{22} = \Psi_{r\alpha}$, $x_{23} = \Psi_{r\beta}$;

$C_2 = (1 \quad 0 \quad 0)$;

$u = [u_{s\alpha}, u_{s\beta}]^T$ and $y = [i_{s\alpha},\ i_{s\beta}]^T$. We assume that T_L is a constant and unknown torque.

_Note that the dimension of the first subsystem is equal to 3 it is due to that the load torque is now considered as a variable to be estimated by the observer.

Objective:

By using only the stator currents measurements of induction motor, the goal is then to design a HGO for subsystem (III-3) in order to reconstruct the mechanical variables (Ω, T_L) and an estimator (E) for (III-4) to estimate the magnetic variables ($\Psi_{r\alpha}$, $\Psi_{r\beta}$).

_Defining the auxiliary signal $v_1 = [u, y, X_2]^T$, and in order to design an observer for (III-3), the following assumptions were introduced:

1. The signals v_1 are bounded and assumed to be regularly persistent [43] in order to guarantee the observability property of the subsystem.

2. $A_1(u, y, X_2, \widetilde{X}_1)$ is globally Lipschitz w.r.t X_2 and uniformly Lipschitz w.r.t (u.y).

3. g_1 $(u, y, X_2, \widetilde{X}_1)$ is globally Lipschitz with respect to X_2 and uniformly Lipschitz w.r.t (u.y).

4. g_2 $(u, y, \widetilde{X}_1, X_2)$ is globally Lipschitz with respect to $\widetilde{X}_1$and uniformly Lipschitz w.r.t (u.y).

Notice that the regularly persistent property of the input is necessary in order to guarantee the observer convergence. Furthermore, the regular persistency of the input is such that the input excites the system sufficiently in order to guarantee the observability of the system, and hence the convergence of the observer.

Assuming that the above assumptions are verified, an observer for subsystem (III-3) is given by:

$$O_{HG}: \begin{cases} \dot{\tilde{Z}}_1 = A_{1_{HGO}}(u, y, Z_2)\tilde{Z}_1 + g_1(u, y, \tilde{Z}_1, Z_2) + S^{-1}C_1^T(y_1 - \hat{y}_1) \\ \dot{S} = -\theta S - A_{1_{HGO}}(u, y, Z_2)S - SA_{1_{HGO}}^T(u, y, Z_2) + C_1^T C_1 \\ \hat{y}_1 = C_1 Z_1 \end{cases} \quad \text{(III-5)}$$

Where $\tilde{Z}_1 = (\tilde{Z}_{11}, \tilde{Z}_{12}, \tilde{Z}_{13})^T$ with $\tilde{Z}_{11} = \hat{i}_{s\alpha}$, $\tilde{Z}_{12} = \hat{\Omega}$ and $\tilde{Z}_{13} = \hat{T}_L$ are the estimated states. Noting that $S^{-1}C_1^T$ is the gain of the observer (III-5) and $S > 0$ and the parameter θ is a positive constant.

The estimator (E) for subsystem (III-4) is given by the following equation

$$E: \begin{cases} \dot{Z}_2 = A_2 Z_2 + g_2(u, y, \tilde{Z}_1, Z_2) \\ \quad\quad \hat{y}_2 = C_2 Z_2 \end{cases} \tag{III-6}$$

Assumption: The initial conditions are known such that estimator (III-6) of (III-4) is well initialized. The induction motor parameters are known with a certain precisions such that the errors between $\hat{\Psi}_{r\alpha}$ and $\hat{\Psi}_{r\beta}$ given by (III-6) and their real values given by (III-4) are supposed to be sufficiently bounded.

How to estimate the gain ?

There is no mathematical method to estimate the gain of the observer, how large or how small is dependent on each project. The gain may be considered as large in one case but actually small in other cases. We may have to do a lot of experiments to identify the proper values of the gain for our project. The rule of thumb is that the gain tends to be larger when there is less noise in measurements, and smaller when more noise in measurements compared to input disturbances.

III-9 Simulation results:

It is necessary to start by noting that:

1- In order to overcome the singularity problem in the observer simulation we added a very small value to the denominators of the elements of the inverse matrix S^{-1}.

2- θ is chosen to be equal to 1000.

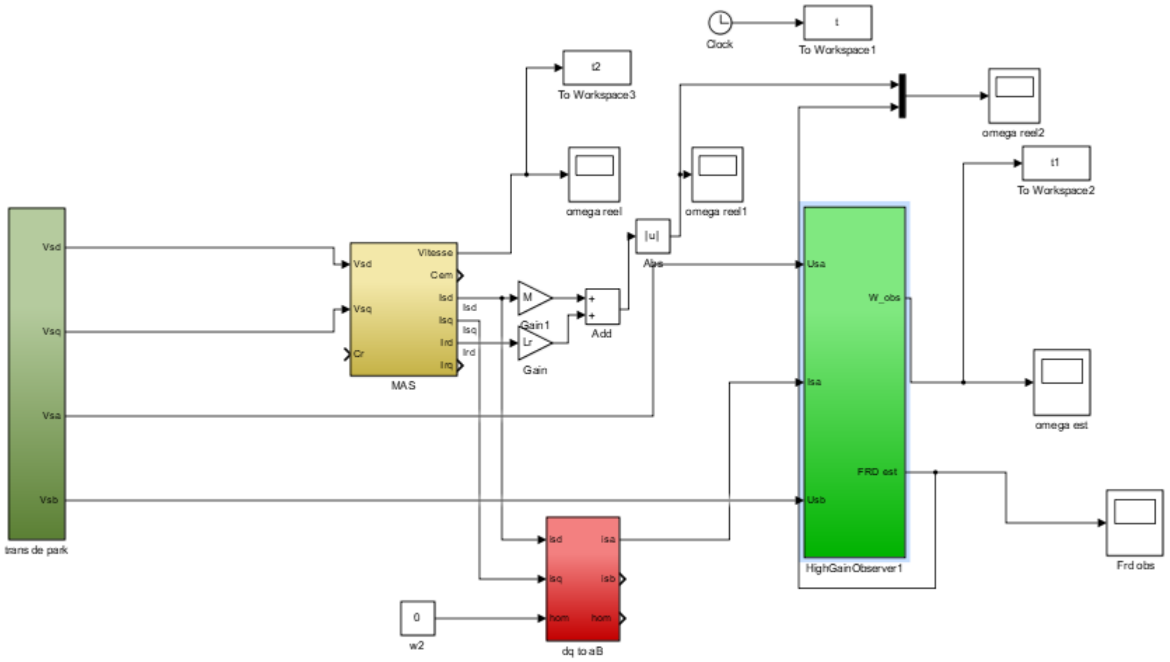

Fig.III-2 block diagram of induction motor model linked to the interconnected high gain observer using MATLAB SIMULINK

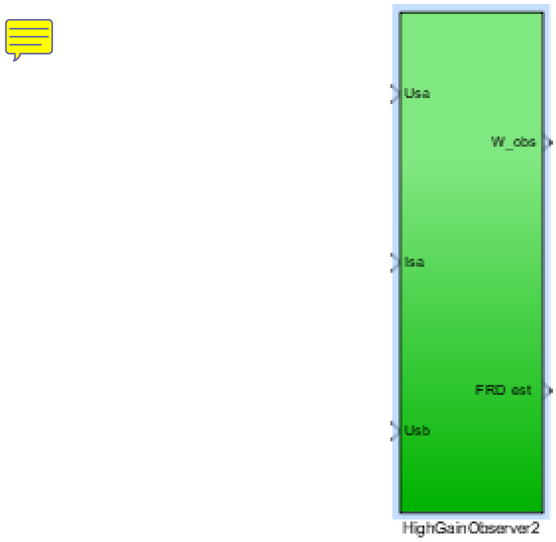

Fig.III-3 the block of the interconnected high gain observer

- As it is shown in the Fig.III-3 we only need the stator voltages $u_{s\alpha}$, $u_{s\beta}$ and the stator current $i_{s\alpha}$ as inputs to the observer to estimate the speed (w_obs).
- 'FRD est' refers to Ψ_{rd} estimated by the observer.

The results for nominal parameters of induction motor

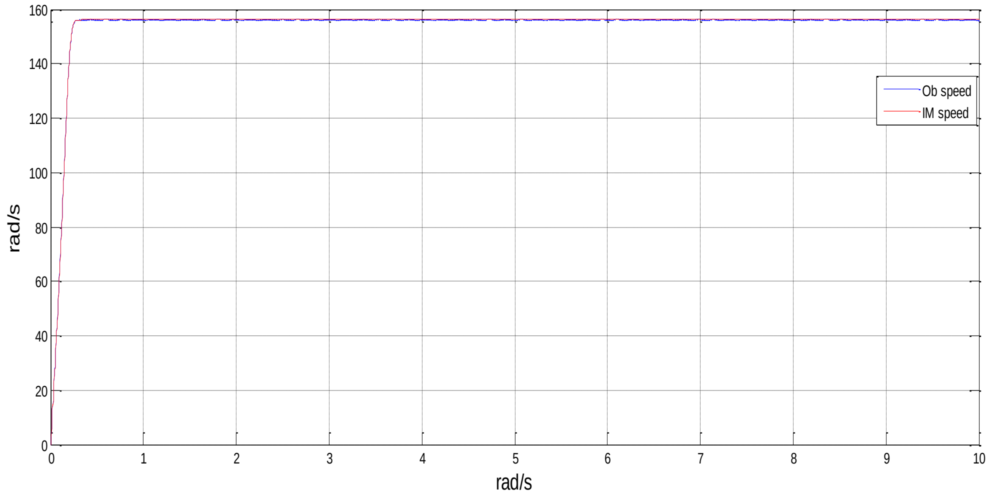

Fig.III-4 IM and observer speed

Fig III-4 shows that speed given by the observer converges very well to the nominal speed and it is almost identical to the IM speed.

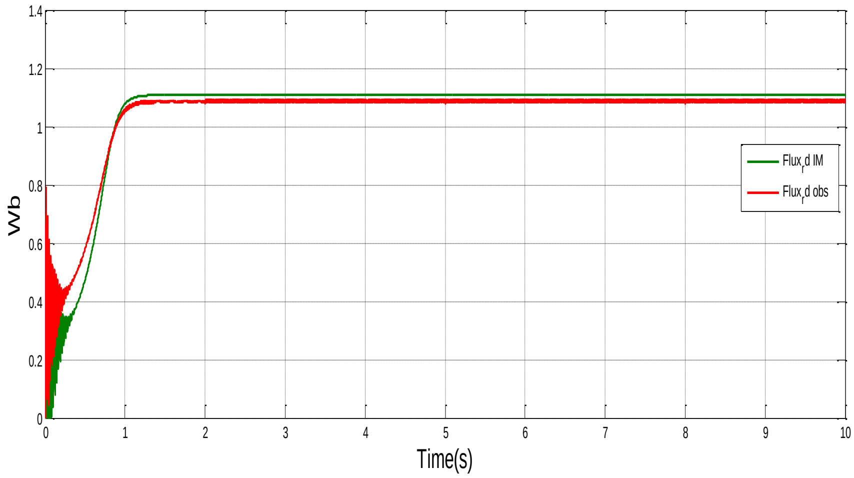

Fig.III-5 IM and observer flux

Fig III-5 shows that flux estimated by the observer is identical to that of induction motor with a very small error.

III-9-1 Robustness analysis to variation of parameters of IM

a. Stator resistance variation (+30% of Rs) (with no load):

Rs=7.2750 ohm

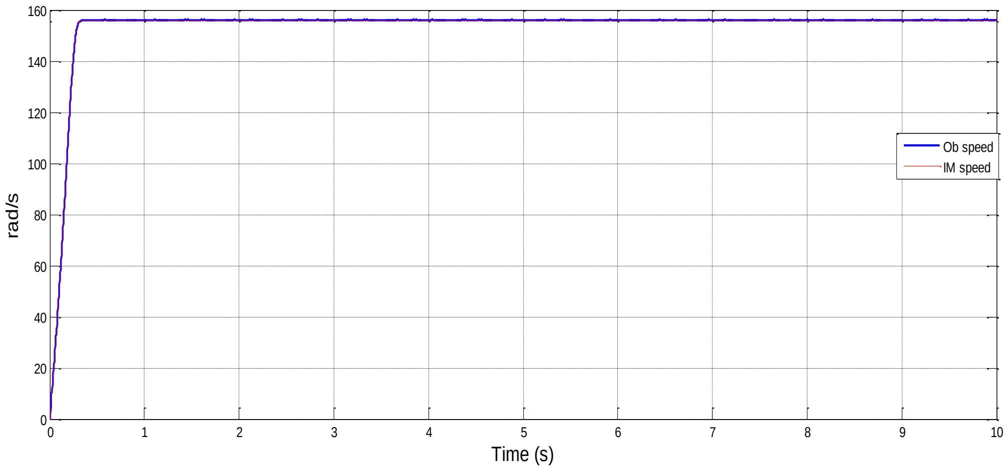

Fig.III-6 IM and observer rotor speed (with +30% of Rs)

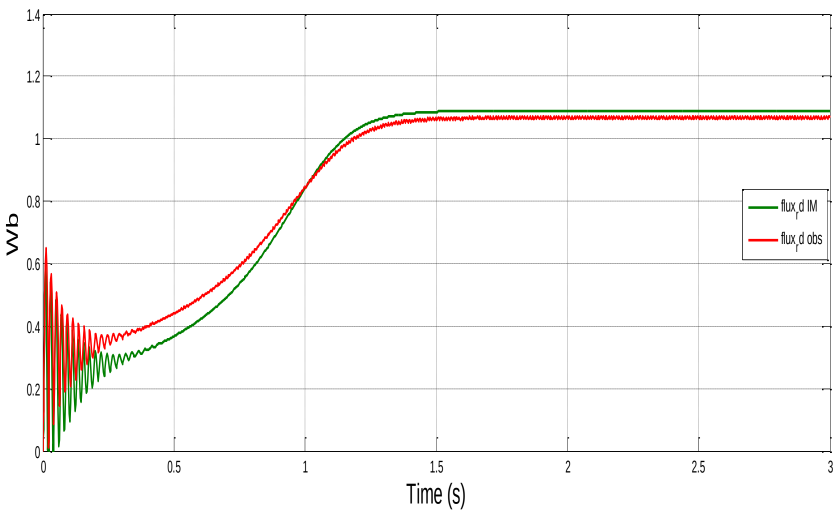

Fig.III-7 IM and observer Flux (with +30% of Rs)

The figures (III-6) and (III-7) above show the results (speed and flux) after +30% of stator resistance, its clear that these results are quiet similar to those ones on the nominal parameters. We conclude that changing the stator resistance has no effect on the speed and the flux observation.

b. Stator inductance variation (+15% of Ls) (with no load):

Now we will show the results for +15% of Ls with no load.

Ls=0.3151H

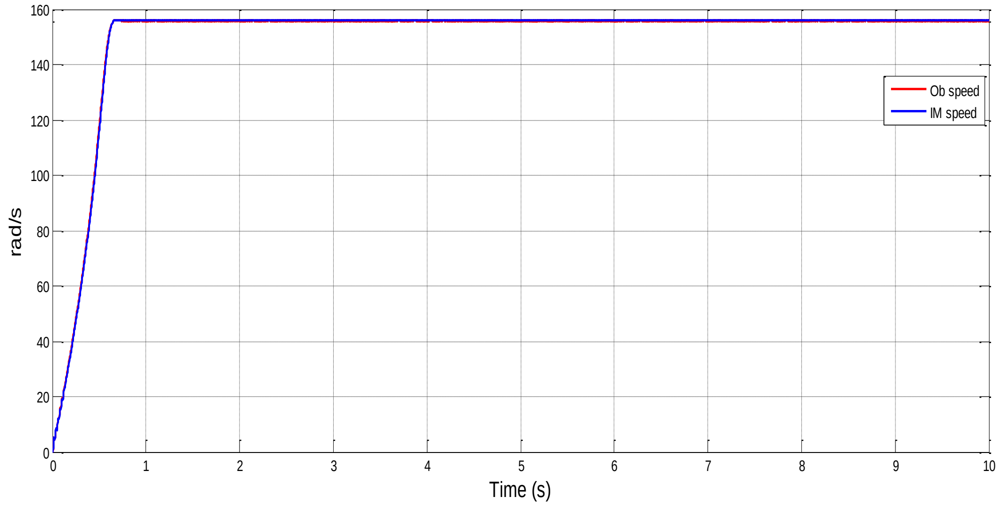

Fig.III-8 IM and observer speed with +15% of Ls

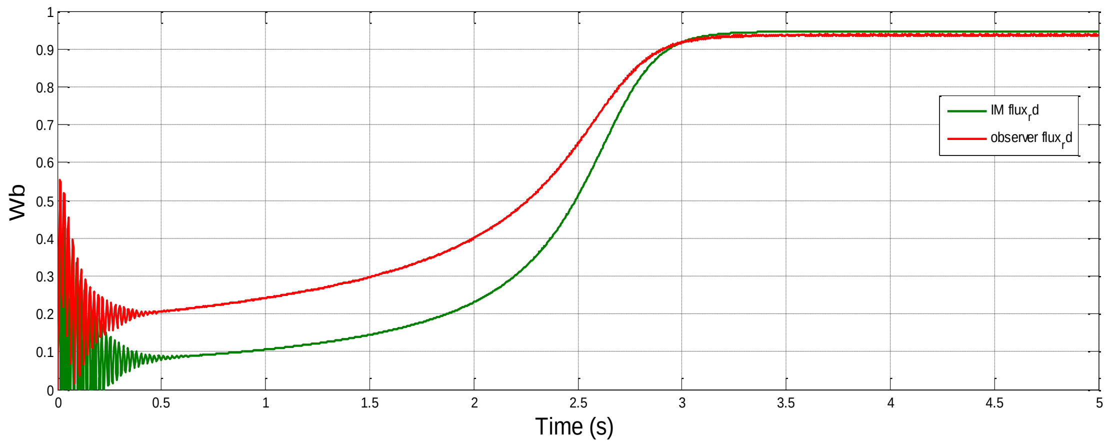

Fig.III-9 IM and observer flux with +15% of Ls

The figures (III-8) and (III-9) show the speed and the flux results respectively for +15% of Ls, the main clear effect is that the speed and flux become slower to achieve their nominal values, in the other side the observer tracked very well the induction motor quantities except a small error for the flux.

We conclude that changing Ls affect the estimation of the flux in the transient mode.

c. Test reversing the direction of rotation (with no load):

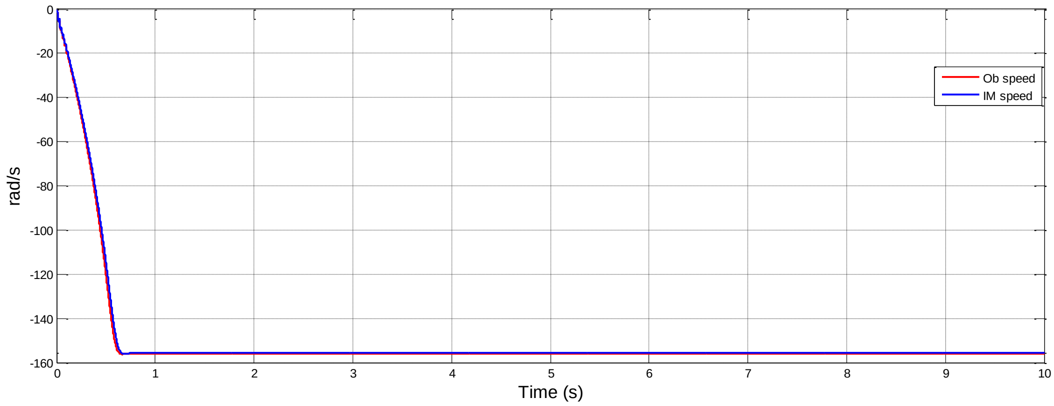

Fig.III-10 IM and observer speed (test reversing the direction of rotation)

Fig.III-10 shows that the observerd speed converges to the reference speed of the IM.

d. Test with load for nominal parameters:

We have applied a load torque.

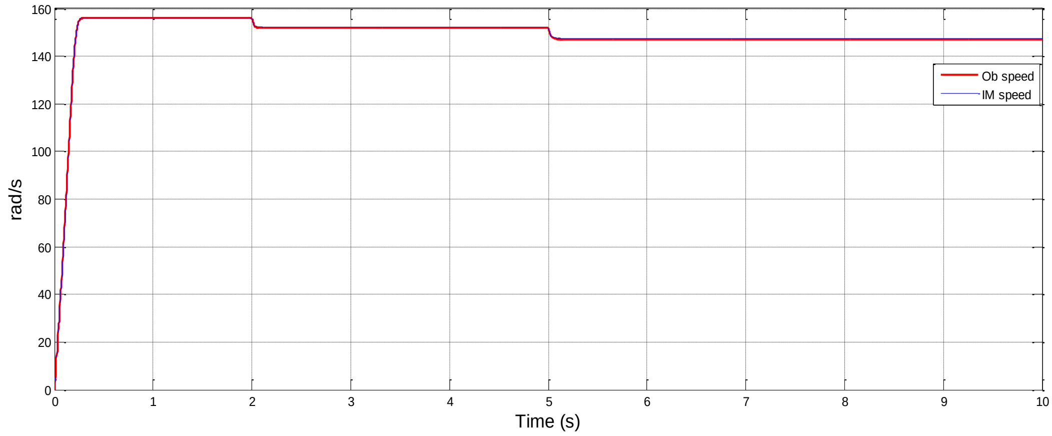

Fig.III-11 IM and observer speed with load

Fig.III-11 shows that the speed estimated by the observer tracks very well the IM speed even when we applied a load torque twice at t=2s, and t=5s, they both decrease as a known consequence of applying a load to the IM.

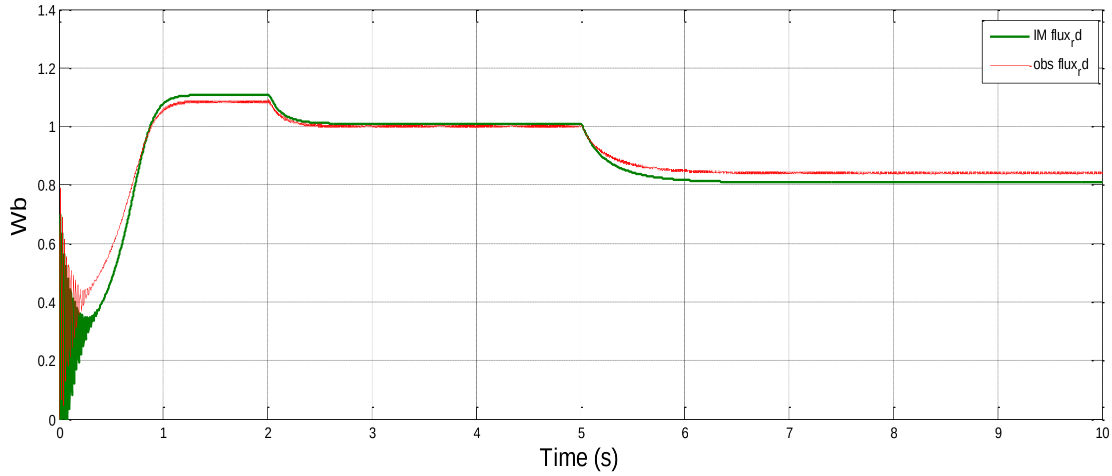

Fig.III-12 IM and observed flux with load

Fig.III-12 shows that our observer works properly, the observed flux is identical to the IM flux with a very small error even when we applied a load torque at t=2s, and t=5s.

III-9-2 Sensorless speed field oriented control of IM:

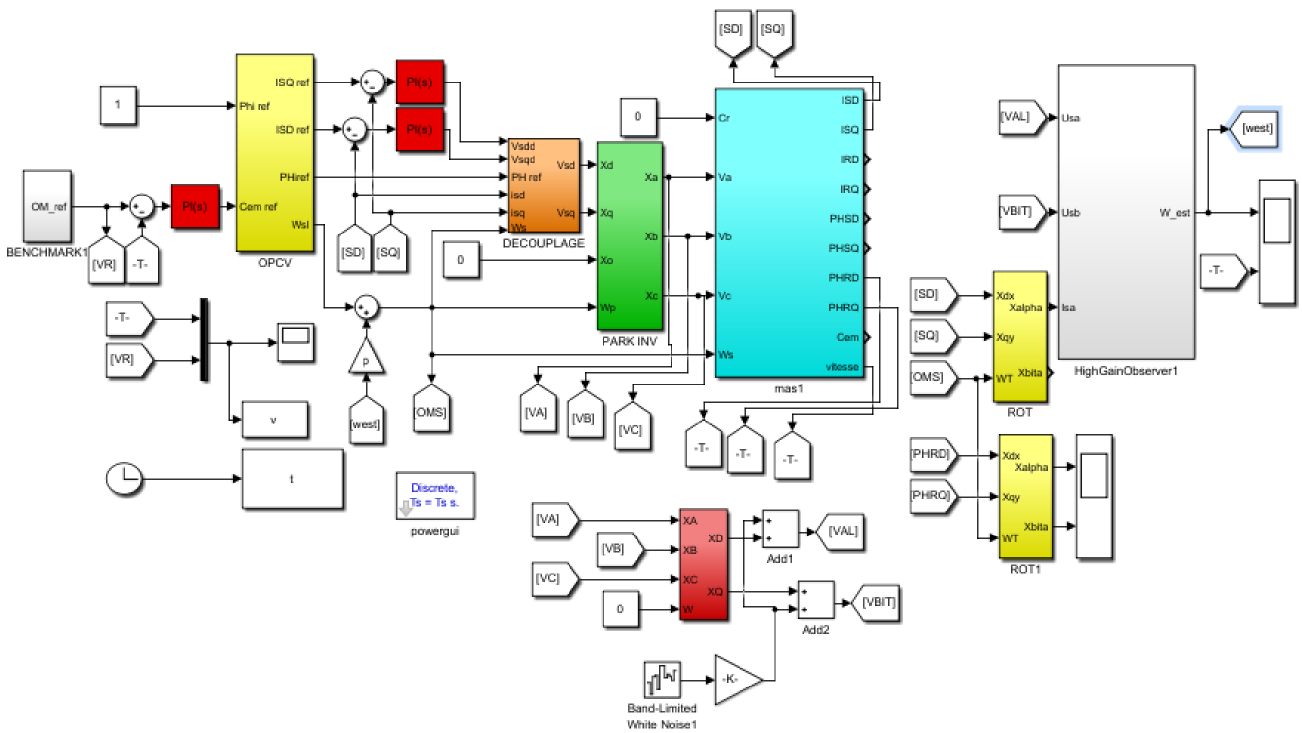

Fig.III-13 sensorless speed field oriented control of IM using MATLAB SIMULINK

a. Test with no load for nominal parameters:

: IM speed

: Estimated speed

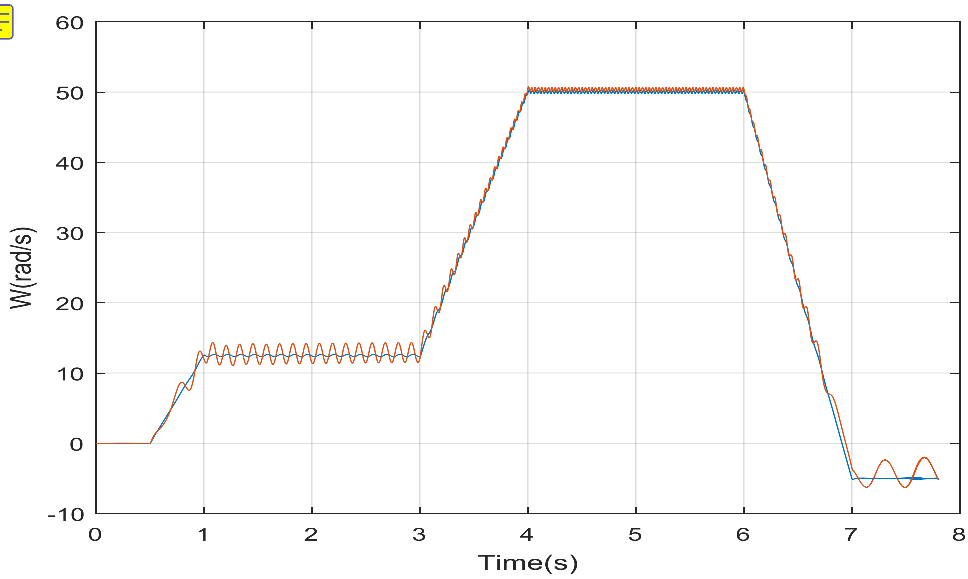

Fig.III-14 IM speed and estimated speed

Motor speed and estimated speed are shown in Fig.III-14 as can be seen from these experimental results, the motor speed tracks satisfactorily the actual speed. In addition we see the fast response of the observer which is due to the high gain that takes the poles of the observing system far to the left of the complex plane, this happens under observable conditions (speed greater enough than 0). Nevertheless, it exists a static error appears when applying a low speed that belongs to the unobservability conditions which is exactly at zero speed and zero load torque applied.

b. Test of robustness to parameters variation:

b-1.+30% of Rs

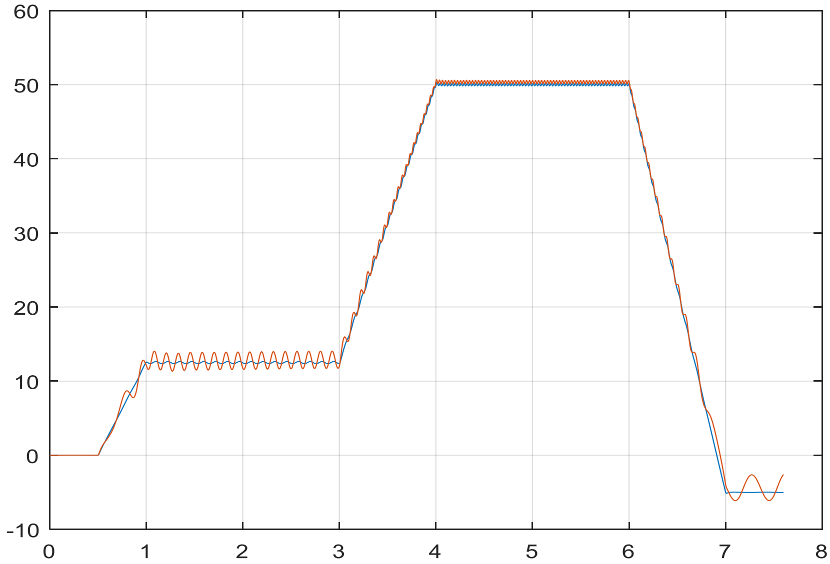

Fig.III-15 IM speed and estimated speed (+%30 of Rs)

Fig.III-15 shows results of robustness test of sensorless speed field oriented control to Rs variations, it is clear that the observed speed tracks very well the speed of IM.

b-2.-5% of Ls

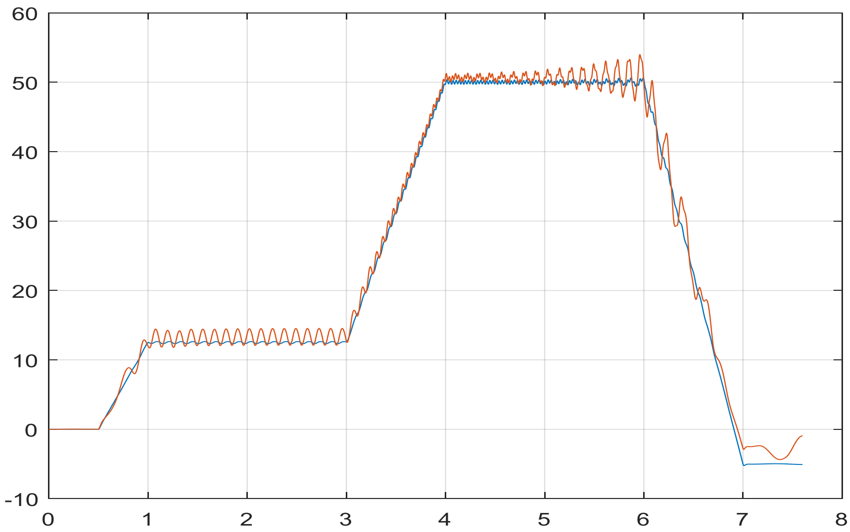

Fig.III-16 IM speed and estimated speed (-%5 of Ls)

Fig.III-16 shows the results of robustness test of sensorless speed field oriented control to Ls variations, it is clear that speed error between induction motor and observer is obvious in this test. From Fig.III-15 and Fig.III-16 we can conclude that our sensorless speed FOC of IM is robust to Rs variations but not very robust to Ls varations.

III-10 Conclusion:

In this chapter we have presented a sensorless field oriented control based on an interconnected high gain observer, we have started by the observability study of induction motor and then we have presented and constructed the interconnected HGO using MATLB SIMULINK and finally we have finished by simulation results that have validated the reliability of the whole sensorless speed control system.

General conclusion:

The work, within the framework of this project, has made it possible to elaborate the study of the indirect rotor field oriented control of the induction motor without speed sensor.

For this, we have used the interconnected high gain observer in sensorless speed control of induction motor. This observer is capable of suitably replacing the speed sensor in the indirect rotor field oriented control strategy. This is to reduce the cost and increase the performance of the control of induction motor by eliminating the speed sensor.

To achieve these goals, in the first chapter we have started by defining the state observer and presenting some well know observers (Luenberger observer, Kalman filter and high gain observer), and we have ended by presenting different types of Sensorless speed control techniques.

In the second chapter, in the first part we have modeled the induction motor in two frames dq and αβ using MATLAB SIMULINK, in the second part we have presented the two types of field oriented control and the principles of indirect rotor field oriented control which is based on separating the flux and the electromagnetic torque and we have ended by associating the IM model with IRFOC, the results obtained from MATLAB SIMULINK have shown the good performance of IRFOC for IM.

In the third chapter we have presented the chosen interconnected high gain observer and its equations. After constructing and testing its reliability we have used the interconnected high gain observer in IRFOC of induction motor, the simulation results obtained from MATLAB SIMULINK have validated the good performance of sensorless speed FOC for IM. We have ended this chapter by testing the robustness of the sensorless control system to motor parameters' variations which has been robust to stator resistance variation but not very robust to stator inductance variation.

This project has allowed us to experiment interconnected high gain observer and to validate its advantages and disadvantages.

Scope for Future work:

1- The first future work that can be proposed is to implement a sensorless speed control for real induction motor.

2- The estimation of the motor parameters maybe an interesting future work for improving the overal performance of the sensorless speed control system.

3- A comparative study can be made with other types of observers to check their suitability for the induction motor.

References

[1] : Jana, A. K. (2010). A nonlinear exponential observer for a batch distillation. In: 11th International Conference on Control Automation Robotics & Vision (ICARCV), 2010, p. 1393-1396.

[2]: Luenberger, D. G. (1964). Observing the state of a linear system. IEEE Transactions on

Military Electronics, 8, 74-80

[3]:Luenberger, D. (1966). Observers for multivariable systems. Automatic Control, IEEE Transactionson,11,190-197.

|4]: Luenberger, D. (1971). An introduction to observers. IEEE Transactions on Automatic Control, 16,596-602.

[5]: Welch, G., & Bishop, G. (1995). An introduction to the Kalman filter.

[6]: H. K. Khalil and L. Praly. “High-gain observers in nonlinear feedback control”. *International Journal of Robust and Nonlinear Control*, 24(6):993–1015, 2014.

[7]: Doyle JC, Stein G. Robustness with observers. IEEE Transactions on Automatic Control 1979; AC-24(4):607–611.

[8]: Petersen IR, Holot CV. High-gain observers applied to problems in disturbance attenuation, H-infinity optimization and the stabilization of uncertain linear systems. Proceedings of American Control Conference, Atlanta, GA, June 1988; 2490–2496.

[9]: Khalil HK, Saberi A. Adaptive stabilization of a class of nonlinear systems using high-gain feedback. IEEE Transactions on Automatic Control 1987; AC-32(11):1031–1035.

[10]: Gauthier, J.P.; Hammouri, H.; Othman, S. (1992), "A simple observer for nonlinear systems applications to bioreactors", IEEE Transactions on Automatic Control, **37** (6): 875–880,

[11]: F. Esfandiari and H.K. Khalil. (1992) Output feedback stabilization of fully linearizable systems. Int. J. Contr., 56:1007-1037.

[12]: A. A. Apte and A. Deshpande, (2015) “Study of Sensorless Control Algorithms for a Permanent Magnet Synchronous Motor Vector Control Drive,”

[13]: TI. (1999) ,“Implementation of a speed field oriented control of 3-phase PMSM motor using TMS320F240,” Texas Instrum. Appl. Rep. SPRA588.

[14]: P. P. Acarnley, J. F. Watson, and S. Member. (2006) ,“Review of PositionSensorless Operation of Brushless Permanent-Magnet Machines,” vol. 53, no. 2, pp. 352–362.

[15]: J.W. Finch and D. Giaouris. (2008) ,“Controlled ac electrical drives,” IEEE Trans. Ind. Electron., vol. 55, no. 2, pp. 481–491.

[16]: A B Kulkarni, M Ehsani. “A novel position sensor elimination technique for the interior permanent-magnet synchronous motor drive,” IEEE Transactions on Industry Applications, 1992, 28(1), pp.144-150.

[17]: Li Yongdong and Zhu Hao. (2008), “Sensorless Control of Permanent Magnet Synchronous Motor – A Survey,” IEEE Vehicle Power and Propulsion Conference (VPPC).

[18]: M. Naidu and B. K. Bose. (1992), “Rotor position estimation scheme of a permanent magnet synchronous machine for high performance variable speed drive,” pp. 48–53 vol.1.

[19]: O. Benjak and D. Gerling. (2010), “Review of position estimation methods for IPMSM drives without a position sensor part II: Adaptive methods,” 19th Int. Conf. Electr. Mach. ICEM 2010, pp. 1–6.

[20]: J. Kang, X. Zeng, Y. Wu, and D. Hu.(2009), “Study of position sensorless control of PMSM based on MRAS,” in Proc. IEEE Ind.Technol.Conf ,pp.1–4.

[21]: H. M. Kojabadi and M. Ghribi, “MRAS-based adaptive speed estimator in PMSM drives,” Int. Work. Adv. Motion Control. AMC, vol. 2006, pp. 569–572, 2006.

[22]: M. Elbuluk, “A sliding mode observer for sensorless control of permanent magnet synchronous motors,” Conf. Rec. 2001 IEEE Ind. Appl. Conf. 36th IAS Annu. Meet. (Cat. No.01CH37248), vol. 2, no. C, pp. 1273–1278, 2001.

[23]: Salvatore L, Cupertino F, Cascella G L. “A new approach to sensorless vector control of SMPM with adaptive sliding-mode observer,” Proceedings of the 2002 IEEE International Symposium on Industrial Electronics, SIE 2002,Vol.2, 8-11 July 2002, pp. 489-494.

[24]: Ludovic Chretien and Iqbal Husain, “Position Sensorless Control of Non- Salient PMSM from Very Low Speed to High Speed for Low Cost Applications,”IEEE, 2007, pp. 289.

[25]: G. Xie, K. Lu, S. K. Dwivedi, J. R. Rosholm, and F. Blaabjerg, “Minimum-Voltage Vector Injection Method for Sensorless Control of PMSM for Low-Speed Operations,” IEEE Trans. Power Electron., vol. 31, no. 2, pp. 1785–1794, 2016.

[26]: Ji-Hoon Jang, Seung-Ki Sul, Jung-Ik Ha, Ide K, Sawamura, M. “Sensorless Drive of Surface-Mounted Permanent-Magnet Motor by High-Frequency Signal Injection Based on Magnetic

Saliency," IEEE Transactions on Industry Applications, Vol.39,Issue 4, July/Aug. 2003, pp. 1031-1039.

[27]: Eskola M, Tuusa H. "Sensorless control of salient pole PMSM using a low-frequency signal injection," European Conference on Power Electronics and Applications, 11-14 Sept. 2005, pp.10.

[28]: Alfio Consoli, Giuseppe Scarcella, and Antonio Testa, "Speed- and current-sensorless fieldoriented induction motor drive operating at low stator frequencies," IEEE Trans. on Ind. Applicat.,Vol.40,NO.1,January/February 2004, pp. 186-193.

[29]: Joachim Holtz, and Juntao Quan, "Drift- and Parameter-Compensated Flux Estimator for Persistent Zero-Stator-Frequency Operation of Sensorless-Controlled Induction Motors," IEEETrans.onInd.Applicat.,Vol.39,No.4,July/August 2003, pp. 1052-1060.

[30]: Joachim Holtz, "Sensorless Control of Induction Motor Drives," IEEE Proc., Vol. 90, No. 8, pp.1359-1394, August 2002.

[31]: Joachim Holtz, and Juntao Quan, "Sensorless vector control of induction motors at very low speed using a nonlinear inverter model and parameter identification," IEEE Trans. on Ind. Applicat.,Vol.38,NO.4,July/August2002,pp.1087-1095.

[32] : B. Houssini, «Prototypage rapide a base de FPGA d'un algorithme de contrôle avancé Le Moteur a induction», Université du Québec, 2010.

[33] **:** J. P. Caron, J. P. Hautier**,** "Modélisation et commande de la Machine Asynchrone", Edition Technip1995 ISBN 2-7108-0683-5.

[34] : B. Walid H. Lamnaouar et H. Houari, «Commande de la machine a induction par technique MRAS », Mémoire d'ingénieur, Université Msila, 2007.

[35] : C. Djamel, «Commande d'une machine asynchrone par logique floue», Thèse de master académique, Université d'EL-Oued, Juin 2014.

[36] : A. M. Trzynadlowski " Control of Induction Motors", Edition Academic Press 2001, 0-12-701510-8.

[37] : G. Grellet, G. Clerc "Actionneurs Electrique Principes-Modèles-Commande", Edit. Eyrolles 97. ISBN 2-212-09352-7.

[38] : S. Khojet El Khil "Commande Vectorielle d'une Machine Asynchrone Doublement Alimentée (MADA) ", Thèse de Doctorat, INP de Toulouse 2006.

[39] : Ramón Blasco Giménez "High Performance Sensorless Vector Control of Induction Motor Drives", Thèse de Doctorat Nottingham 1995.

[40] : Mihai Comanescu "Flux and Speed Estimation Techniques for Sensorless Control of Induction Motors ", Thèse de Doctorat Université de l'état d'Ohio USA 2005.

[41] :] H.Kubota,K.Matsus eandT.Nakano,"DSP-Based Speed Adaptative Flux Observer of Induction Motor" IEEE Transactions on Industry Applications, Vol.29, No.2, pp.344-348, Mar./Apr. 1993.

[42] : X. Xu and D. Novotny, "Implementation of Direct Stator Flux Orientation Control on a Matsuse DSP-Based Speed Adaptative Flux Observer of Induction Motor", IEEE Transactions on Industry Applications, Vol.29, No. 2, pp.344-348, Mar./Apr. 1993.

[43]: G. Besançon and H. Hammouri, "On Observer Design for Interconnected Systems", Journal of Mathematical Systems, Estimation and Control, vol.8, No.4, 1998.

[44]: M.Ghanes,J.DeLeonand A.Glumineau"Validation of an Interconnected HighGain Observer for Sensorless Induction Motor On Low Frequencies Benchmark: Application to an Experimental Set-up",IEEProc.Control Theory and Applications. Vol. 152, No. 4, pp. 371-378, July 2005.

[45]: H. Hammouri and J. DeLeon, "Observer synthesis for state-affinne systems", Proc 29th IEEE Conference on Decision and Control, Honolulu, Hawaii, pp. 784-785, 1990.

[46] : C. Canudas de Wit, A. Youssef, J.P. Barbot, Ph. Martin and F. Malrait, "Observability Conditions of Induction Motors at low frequencies", IEEE Conference on Decision and Control, Sydney, Australia, December 2000

[47]: S. Ibarra-Rojas, J. Moreno and G. Espinosa, "Global observability analysis of sensorless induction motor", Automatica, Vol.40, Issue : 6, pp. 1079-1085, June 2004.

[48]: GHANES M, DE LEON J, GLUMINEAU A : 'Observability study and observer-based interconnected from for sensorless induction motor' IEEE conf.Decision and control,SanDiego, CA, USA, 13-15 December 2006