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Fuzzy-Logic Control of Induction Motor.

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Dedication

« I dedicate this modest work in particular to my dear mother and dear father for all the sacrifices and support they gave me throughout my academic and life career, i also dedicate it to all my relatives and the family of the scientific community and all those who had a role in the completion of this thesis. »

Acknowledgement

" First of all, i would like to express my sincere thanks and gratitude in particular to my dear mother and dear father as well to everyone who had a role in the completion of this modest work, especially, Dr. Djokhrab Ala Eddine, for adopting and believing in this work as well. I would like to express my gratitude to the members of the jury who took care of reviewing and analyze my humble work in the midst of their intense preoccupations in order to provide an added value to the thesis. I also thank all the profs of the electrical engineering department, as well as the administration and university staff. "

ABSTRACT

تطبيق تقنية المنطق الضبابي لـلتحكم في محرك حثي غير متزامن

ملخـــص: يوم بعد يوم يزداد الاعتماد على المحركات الكهربائية الحثية غير المتزامنة على حساب الأنواع الأخرى وذلك في مختلف التطبيقات اليومية الحياتية والصناعية وهذا لتميزها ببنائها البسيط والمتين مما جعلها خيارا جذابا ,ولكن لا يخلو هذا النوع من المحركات من التعقيدات والمشاكل في التحكم فيه, من بين هذه المشاكل فقدان في استجابة العزم والسرعة المحرك عند تطبيق حمولة عليه وأيضا تغير ديناميكيته وخصائصه الداخلية. قدمت هاته المذكرة مقاربة جديدة للتحكم الشعاعي الموجه (FOC) للمحركات الحثية غير المتزامنة عن طريق تطوير تحكم فعال بإدخال تقنيات اكثر فعالية على التحكم الشعاعي الموجه المباشر المطبق على الجزء الدوار (DFOC) لضمان الفصل بأفضل شكل ممكن للعزم عن التدفق الكهرومغناطيسي وأيضا صلابة ضد تغير الديناميكيات الداخلية للمحرك, تتمثل هاته التقنيات في استبدال المتحكمات الكلاسيكية (PI) بمتحكمات أخرى احدث من سابقتها تدعى بمتحكمات المنطق الضبابي (FLC) التي وفرت مرونة و دقة أكثر من نظيرتها على مستوى استجابة السرعة و العزم للمحرك الحثي الغير متزامن و فصل مثالي للعزم و التدفق الكهرومغناطيسي وأيضا صلابة أكبر للصمود في وجه تغير ديناميكيات المحرك بمرور الوقت او بتدخل عوامل خارجية. نتائج هاته الدراسة وضحت وناقشت النتائج المتحصل عليها لكل تقنية على حدى وذلك على طول هاته المذكرة بهدف التأكد والنظر في صلابة وجودة أداء كل تقنية. تم تنفيذ كل التقنيات المذكورة بالاستعانة ببرنامج الماتلاب للمحاكات والتصميم.

الكلمات المفتاحية :المحرك الحثي اللاّتزامني (IM), عاكس الجهد المتباطئ (PWM) , التحكم الشعاعي المباشر للجزء الدوار (DFOC),المتحكم التناسبي التكاملي (PI), المتحكم المنطقي الضبابي (FLC)..

Application of Fuzzy-Logic for Controlling an Asynchronous Induction Motor.

Abstract: Day after day, the reliance on asynchronous induction motors is increasing at the expense of other types in various daily, life and industrial applications. Loss of Motor torque and speed response when a load is applied to it, as well as a change in its dynamics and internal characteristics. This note introduces a new approach to Field Directed Control (FOC) for asynchronous induction motors by developing an efficient control by introducing FL techniques to Direct Field Oriented Control (DFOC) to ensure optimum decoupling of torque and electromagnetic flux as well as rigidity Against changing the internal dynamics of the Motor, these techniques are to replace the classic controllers Proportional Integral (PI) with other modern controllers than their predecessor called Fuzzy Logic Controllers (FLC), which provided more flexibility and accuracy than their counterparts. I was studied and developed to control and Perfect response in terms of speed and torque of the motor, perfect decoupling of torque and electromagnetic flux, as well as greater robustness to withstand the changing dynamics of the motor over time or the intervention of external factors. The results of this study clarified and discussed the results obtained for each technique separately, along the length of this note, in order to ascertain and consider the robustness and quality of performance of each technique. All the mentioned techniques were implemented using MATLAB software for simulation and design.

KEYWORDS: Induction Motor (IM), Voltage Inverter with Hysteresis (H-PWM), Direct Field Oriented Control of the Rotor (DFOC), Proportional Integral Controller (PI), Fuzzy Logic Controller (FLC).

TABLE OF CONTENTS

ABSTRACT I
TABLE OF CONTENTS III
LIST OF FIGURES VI
LIST OF TABLES VIII
LIST OF ABBREVIATIONS IX
LIST OF SYMBOLS X
GENERAL INTRODUCTION 1
CHAP_I MODELING AND SIMULATION OF THE INDUCTION MACHINE 4
I. 1 - INTRODUCTION: 5
I. 2- HISTORICAL REVIEW OF INDUCTION MACHINE 5
I.2.1- Historical review of induction machine: 6
I.2.2- Induction machine definition: 7
I.2.3- Induction machine types: 7
I.2.4- Constitution of the induction machine: 7
I.2.4.a - Stator construction: 8
I.2.4.b - Rotor construction: 8
I.2.5- Principle of operation of an induction motor: 9
I.2.6- Advantages and disadvantages of the induction machine: 10
I.2.6.a - The advantages: 10
I.2.6.b - The disadvantages: 11
I.2.7- Fields of application of induction motors: 11
I.3.1- Simplifying assumptions: 12
I.3.2- Modeling of the IM in the "a, b, c" three-phase plane: 12
I.3.3- General equations of the three-phase induction machine in "a, b, c" plane: 13
I.3.3.1- The electrical equations: 13
I.3.3.2- The magnetic equations: 14
I.3.3.3- Mechanical equations: 15
I. 3 - PARK MODEL OF INDUCTION MACHINE: 16
I.4.1- Transformation of "PARK": 16
I.4.2- Application of the park transformation to the IM: 17
I.4.2.1 - Voltage equations: 18
I.4.2.2 - Flux equations: 18
I.4.2.3 - Mechanical equations: 18
I. 4 - CHOICE OF REFERENTIAL: 19
I. 5. 1 - Referential related to the stator: 19
I. 5. 2 - Referential related to the rotor: 19
I. 5. 3 - Referential related to the rotating field: 19
I. 5 - STATE SPACE REPRESENTATION OF THE INDUCTION MACHINE: 20
I. 6 - SIMULATION OF THE INDUCTION MACHINE MODEL: 22
I.7.1 - Asynchronous machine simulation block diagram: 22
I.7.2 - Simulation and interpretation of results: 24
I. 8 - THREE-PHASE VOLTAGE INVERTER: 25
I.8.1 - Different types of inverter for supplying IM: 25
I.8.1.1 - Current inverter: 25

I.8.1.2 - Voltage inverter: ... *26*
I.8.2 - PWM VOLTAGE INVERTER WITH HYSTERESIS: ... 26
I.8.2.1 - Principle: ... *27*

I. 9 - MODELING OF THE PWM INVERTER: ... 28

I. 10 - SIMULATION AND INTERPRETATION OF THE RESULTS: ... 30

I. 11 - CONCLUSION: ... 32

CHAP_II ROTOR FLUX ORIENTATION VECTOR CONTROL ... 33

II. 1 - INTRODUCTION: ... 34

II. 2 - FOC HISTORICAL REVIEW: ... 34

II. 3 - ADVANTAGES AND DISADVANTAGES OF VECTOR CONTROL: ... 35

II. 4 - FIELD ORIENTATION CONTROL (FOC): ... 35

II. 5 - CHOICE OF FIELD ORIENTATION: ... 36

II. 6 - PRINCIPLE OF FOC BY ORIENTATION OF THE ROTOR FLUX: ... 37

II. 7 - TYPES OF FIELD ORIENTATION CONTROL: ... 39

II.7. 1 - DIRECT VECTOR CONTROL: ... 39
II.7. 2 - INDIRECT VECTOR CONTROL: ... 40

II. 8 - STRUCTURE OF FOC OF THE ROTOR FLUX OF IM SUPPLIED WITH VOLTAGE: ... 40

II.8. 1 - SYSTEM OF EQUATIONS RELATED TO ROTOR FLUX: ... 41
II.8. 2 - INPUT-OUTPUT DECOUPLING: ... 42
II.8. 3 - DECOUPLING BY COMPENSATION METHOD: ... 42
II.8. 4 - FIELD WEAKENING BLOCK: ... 43
II.8. 5 - PRINCIPLE OF OPERATION: ... 44
II.8. 6 - ROTOR FLUX ESTIMATION: ... 44
II.8. 7 - REGULATION OF THE SYSTEM: ... 45
II.8.7. 1 - Design of regulators: ... *45*
II.8.8 - REGULATORS CHARACTERISTICS: ... 46
II.8.8.1 - Stability: ... *46*
II.8.8.2 - Accuracy: ... *46*
II.8.8.3 - Rapidity: ... *46*

II.9- CALCULATION OF REGULATORS: ... 46

II.9.1 - DIRECT STATOR CURRENT isd REGULATION: ... 46
II.9.2 - QUADRATURE STATOR CURRENT isq REGULATION: ... 48
II.9.3 - VELOCITY REGULATION Ωr : ... 48
II.9.4 - FLUX REGULATION φr: ... 49

II. 10 - SIMULATION AND INTERPRETATION: ... 50

II.10 .1 - SIMULATION OF THE SPEED VARIATION: ... 50
II.10 .2 - SIMULATION OF THE INVERSION OF THE ROTATIONNEL DIRECTION : ... 51
II.10 .3 - SIMULATION WITH LOAD TORQUE VARIATION: ... 52
II.10 .4 - ROBUSTNESS TEST FOR THE VARIATION OF THE ROTOR RESISTANCE: ... 54

II. 11 - CONCLUSION: ... 55

CHAP_III APPLICATION OF FUZZY LOGIC CONTROL ON DFOC ... 56

III.1 - INTRODUCTION: ... 57

III.2 - OVERVIEW OF THE HISTORY AND APPLICATIONS OF FUZZY LOGIC: ... 57

III.3 - ADVANTAGES AND DISADVANTAGES OF FUZZY LOGIC CONTROL: 58

III.3.1 - THE ADVANTAGES OF FUZZY LOGIC CONTROL ARE: 58
III.3.2 - THE DISADVANTAGES OF FUZZY LOGIC CONTROL ARE: 59

III.4 - FUZZY LOGIC IN INDUSTRY: 59

III.5 - FUZZY SET THEORY: 59

III.5.1 - DEFINITION: 59
III.5.2 - FUZZY SETS: 61
III.5.3 - DIFFERENT FORMS OF MEMBERSHIP FUNCTIONS: 61
III.5.3.1 - Triangular and trapezoidal membership function: 62
III.5.3.2 - Piecewise exponential membership function: 62
III.5.3.3 - The Gaussian membership function: 62
III.5.3.4 - Singleton membership function: 62
III.5.4 - FUZZY LOGIC OPERATORS: 63
III.5.4.1 - "NON" Operator: 63
III.5.4.2 - "AND" Operator: 63
III.5.4.3 - "OR" Operator: 64
III.5.5 - FUZZY REASONING: 64
III.5.5.1 - Fuzzy implication: 64
III.5.6 - THE GENERALIZED MODUS PONENS: 65

III.6 - CONTROL USING FUZZY LOGIC: 65

III.6.1 - STRUCTURE OF A FUZZY LOGIC CONTROL: 66
III.6.1.1 - The fuzzification interface: 66
III.6.1.2 - Rules base: 67
III.6.1.3 - Standardization: 67
III.6.1.4 - Inference engine: 67
III.6.1.5 - Defuzzification interface: 70

III.7 - APPLICATION OF FUZZY LOGIC TO IM SPEED CONTROL: 71

III.7.1 - DESIGNING AN FLC: 71
III.7.2 - DEVELOPMENT OF A FUZZY CONTROLLER: 71
III.7.2.1 - Fuzzification stage: 73
III.7.2.2 - Stage of establishing inference rules: 74
III.7.2.3 - Defuzzification: 75

III.8 - SIMULATION AND INTERPRETATION: 75

III.8 .1 - SIMULATION OF THE SPEED VARIATION: 76
III.8 .2 - SIMULATION OF THE INVERSION OF THE ROTATIONAL DIRECTION: 77
III.8 .3 - SIMULATION WITH LOAD TORQUE VARIATION: 78
III.8 .4 - ROBUSTNESS TEST FOR THE VARIATION OF THE ROTOR RESISTANCE: 80

III.9 - CONCLUSION: 81

GENERAL CONCLUSION 82

ANNEX i

REFERENCES ii

LIST OF FIGURES

Fig I- 1 : Ferrari's induction motor (1885) and Tesla's induction motor (1886). 5

Fig I- 2 : Classification of electric motors 6

Fig I- 3 : Parts of squirrel cage induction motor 8

Fig I- 4 : Stator of three-phase induction motor 8

Fig I- 5 : "Squirrel cage rotor" and "Phase wound rotor" 9

Fig I- 6 : Principle operation of an induction motor 10

Fig I- 7 : Schematic representation of a three-phase asynchronous machine. 12

Fig I- 8 : Transition from a three-phase to a two-phase system 16

Fig I- 9 : Choice of repository 20

Fig I- 10 : Simulation diagram of an IM powered by an electrical network. 23

Fig I- 11 : Block diagram of the Induction machine model. 23

Fig I- 12 Simulation results of the IM powered by the sine wave voltage source with application of load $Cr = 25N.m$ at $t = 1s$ 24

Fig I- 13 : Current inverter 25

Fig I- 14 : Voltage inverter 26

Fig I- 15 : Principle of PWM control 27

Fig I- 16 : PWM simulation diagram with hysteresis. 28

Fig I- 17 : "Diagram of the voltage inverter" and "Simplified diagram" 28

Fig I- 18 : IM simulation diagram powered by PWM hysteresis voltage inverter 30

Fig I- 19 : Simulation results of the IM powered by Inverter with Hysteresis and with application of load $Cr = 25N.m$ at $t = 1s$ 31

Fig II- 1 : Orientation of the rotor flux on the axis d. 36

Fig II- 2 : Equivalence between the control of a separate-field motor and the vector control of an IM 39

Fig II- 3 : The concept of decoupling. 41

Fig II- 4 : Structure of the rotor flux orientation control 42

Fig II- 5 : Reconstruction of voltages $Vsd\ and\ Vsq$ 43

Fig II- 6 : Field weakening block. 44

Fig II- 7 : Schematic diagram of direct vector control with oriented rotor flux of the IM 44

Fig II- 8 : Diagram of the open-loop flux estimator 45

Fig II- 9 : Representation of PI control 45

Fig II- 10 : Block diagram of " isd" stator current regulation 47

Fig II- 11 : Block diagram of $\boldsymbol{isq}$ stator current regulation 48

Fig II- 12 : Diagram of speed regulation block 48
Fig II- 13 : Scheme of flux regulation block φr 49
Fig II- 14 : Block diagram of direct vector control 50
Fig II- 15 : DFOC simulation results during IM shaft speed variation 51
Fig II- 16 : DFOC simulation results during IM shaft speed inversion 52
Fig II- 17 : DFOC simulation results during no-load starting followed by a load torque variation . 53
Fig II- 18 : DFOC simulation results of robustness test for rotor resistance variation. 54

Fig III- 1: Classification of people into three sets according to classical logic. 60
Fig III- 2 : Classification of people into three sets according to fuzzy logic. 60
Fig III- 3 : Different forms for membership functions. 61
Fig III- 4 : NON-Operator. 63
Fig III- 5 : AND, OR Operator 64
Fig III- 6 : Structure of a fuzzy controller 66
Fig III- 7 : Main steps when designing an FLC. 71
Fig III- 8 : Synoptic diagram of a speed regulator 72
Fig III- 9 : The membership functions 74
Fig III- 10 : Diagram of a DFOC Based on FLC to Controlling IM. 76
Fig III- 11 : DFOC with an FLC simulation results during IM shaft speed variation 77
Fig III- 12 : DFOC with an FLC simulation results during IM shaft speed inversion. 78
Fig III- 13 : DFOC with an FLC simulation results during no-load starting followed by a load torque variation. 79
Fig III- 14 : DFOC with an FLC simulation results of Robustness test for rotor resistance variation. 80

LIST OF TABLES

Tab III- 1 : Abbreviations of fuzzy sets 68
Tab III- 2 : Matrix of complete inferences.. 69
Tab III- 3 : Inferences rules table.. 74
Tab III- 4 : Rules table for speed FLC 75

LIST OF ABBREVIATIONS

AC : Alternative Current.

DC : Direct Current.

IM : Induction Motor

ACIM : AC Induction Motor..

P.I : Proportional Integral.

PWM : Pulse Width Modulation.

FOC : Field Oriented Control.

DFOC : Direct Field Oriented Control.

ORF : Orientation of The Rotor Flux.

DVC : Direct Vector Control.

FL : Fuzzy Logic.

FLC : Fuzzy Logic Controller.

$\boldsymbol{\mu}$: Membership Function.

LIST OF SYMBOLS

$\boldsymbol{t}$: Time.

$\boldsymbol{s}, \boldsymbol{r}$: Respective indices of the stator and the rotor.

$\boldsymbol{\Omega_s}$: Synchronous speed of rotation of the rotating field in $[rad. sec^{-1}]$.

$\boldsymbol{p}$: Number of pole pairs.

$\boldsymbol{f}$: The frequency in Hz.

$\boldsymbol{n_s}$: Synchronous rotational speed of the rotating field $[tr. min^{-1}]$.

$\boldsymbol{n_r}$: Rotor speed $[tr. min^{-1}]$.

$\boldsymbol{\omega}$: Pulsation of alternating currents in $[rad. sec^{-1}]$.

$\boldsymbol{\omega_s}$: Pulsation of stator currents $[rad/sec]$.

$\boldsymbol{\omega_r}$: Pulsation of rotor currents $[rad/sec]$.

$\boldsymbol{\omega_{slip}}$: Slip Pulse $[rad/sec]$.

$\boldsymbol{slip}$: Slip coefficient.

$\boldsymbol{\omega_{coor}}$: Axis frame pulsation (d, q) $[rad/sec]$.

$\boldsymbol{\omega_{rated}}$: Rated angular speed of the machine$[rad/sec]$.

$\boldsymbol{V_{s(abc)}}$, $\boldsymbol{V_{r(abc)}}$: Stator and rotor voltage vectors $[V]$.

$\boldsymbol{i_{s(abc)}}$, $\boldsymbol{i_{r(abc)}}$: Stator and rotor current vectors $[A]$.

$\boldsymbol{\varphi_{s(abc)}}$, $\varphi_{r(abc)}$: Stator and rotor flux vectors $[Wb]$.

$\boldsymbol{R_s}$, $\boldsymbol{R_r}$: Stator and rotor self-resistance $[\Omega]$.

$\boldsymbol{L_s}$, $\boldsymbol{L_r}$: Stator and rotor self-inductors $[H]$.

$\boldsymbol{M_{sr}}$: Mutual inductance between stator phases $[H]$.

$\boldsymbol{M_{rs}}$: Mutual inductance between rotor phases $[H]$.

$\boldsymbol{M}$: Mutual inductance between stator and rotor phases $[H]$.

$\boldsymbol{C_e}$: The electromagnetic couple $[N. m]$.

$\boldsymbol{C_r}$: The resisting couple $[N. m]$.

$\boldsymbol{J}$: Moment of inertia of rotating parts $[Kg. m^2]$.

$\boldsymbol{f}$: ACIM viscous coefficient of friction $[N. m. sec. rad^{-1}]$.

$\boldsymbol{\sigma}$: Blondel's total dispersion coefficient.

$\boldsymbol{T_s}$, $\boldsymbol{T_r}$: Stator and rotor time constant.

$\boldsymbol{d}$: Direct axis index.

$\boldsymbol{q}$: Quadrature axis index.

$\boldsymbol{o}$: Homopolar axis index.

$[\boldsymbol{P}(\boldsymbol{\theta_d})]$: PARK's transformation matrix.

$[\boldsymbol{P}(\boldsymbol{\theta_d})]^{-1} = [\boldsymbol{P}(\boldsymbol{\theta_d})]^{T}$: The inverse PARK transformation matrix.

$\boldsymbol{\theta_s}$: Stator position $[rad]$.

$\boldsymbol{\theta}$: Position between the stator axis as and the rotor axis ar $[rad]$.

$\boldsymbol{\theta_r}$: Electric rotor position $[rad]$.

$\boldsymbol{V_{ds}}$, $\boldsymbol{V_{qs}}$: Components of the stator voltage along the d axis and the q axis $[V]$.

$\boldsymbol{V_{dr}}$, $\boldsymbol{V_{qr}}$: Components of the rotor voltage along the d axis and the q axis $[V]$.

$\boldsymbol{i_{ds}}$, $\boldsymbol{i_{qs}}$: Components of the stator current along the d axis and the q axis $[A]$.

$\boldsymbol{i_{dr}}$, $\boldsymbol{i_{qr}}$: Components of the rotor current along the d axis and the q axis $[A]$.

$\boldsymbol{\varphi_{ds}}$, $\boldsymbol{\varphi_{qs}}$: Components of the stator flux along the d axis and the q axis $[Wb]$.

$\boldsymbol{\varphi_{qr}}$, $\boldsymbol{\varphi_{qr}}$: Components of the rotor flux along the d axis and the q axis $[Wb]$.

$\boldsymbol{V_{dc}}$: This is the DC supply voltage of the inverter.

$\boldsymbol{K_i}$: Integral gain.

$\boldsymbol{K_p}$: Proportional gain.

$\boldsymbol{K_{i1}}$, $\boldsymbol{K_{p1}}$: Proportional and integral gains of the direct stator current PI regulator (d).

$\boldsymbol{K_{i2}}$, $\boldsymbol{K_{p2}}$: Proportional and integral gains of the quadrature stator current PI regulator (q).

$\boldsymbol{K_{i3}}$, $\boldsymbol{K_{p3}}$: Proportional and integral gains of speed PI regulator.

$\boldsymbol{K_{if}}$, $\boldsymbol{K_{pf}}$: Proportional and integral gains of the rotor flux PI regulator.

$\boldsymbol{\tau}$: Time constant.

$\boldsymbol{\xi}$: Damping constant.

$\boldsymbol{\omega_{ref}}$: Reference speed $[rad/sec]$.

$\boldsymbol{x_{in1}\, and\; x_{in2}}$: Fuzzy sets of the input variables.

$\boldsymbol{x_{out}}$: Fuzzy output variable.

$\boldsymbol{En, dEn}$: Fuzzy sets of input variables.

$\boldsymbol{dUn}$: Fuzzy output variables.

General Introduction

Because asynchronous motors have a simple construction and are straightforward to maintain, they have a sizable market. Induction motors are widely used in a variety of applications, ranging from industrial applications such as pumps, fans, and blowers to residential appliances, due to their low cost and durability. Induction motors have traditionally been operated at a single speed dictated by the frequency of the main voltage and the number of poles in the motor. Because there is no linear relation between the motor current and the generated torque as there is for a DC motor, controlling the speed of an induction motor is significantly more difficult than controlling the speed of a DC motor. Furthermore, unlike dc motors, induction motors may be operated without maintenance for a long period due to their brushless architecture. The squirrel cage motor is the least priced and most extensively used induction motor. To establish a magnetic field in the rotor, no current supply from outside the rotor is required. This is why this motor is just so reliable and affordable [1].

Each stage of the modeling, analysis, and control of all electrical machinery necessitates the creation of correct system models. The degree of precision required of these models is entirely dependent on the modeling stage in question. According to Nabae et al. (1980) and Murata et al. (1986), the mathematical description utilized in machine modeling demands very fine tolerance levels (1990). However, some assumptions may be made in the building of acceptable models for control purposes, simplifying the final machine model significantly. Nonetheless, for both steady state and transient operation situations, these models must include the basic features of both the electromagnetic and mechanical systems (Nowotny and Lipo - 1996). Furthermore, because contemporary electric machines are generally powered by switching power conversion stages, the created motor models must be applicable to a wide range of applied voltage and current waveforms [4] At the same time, with the advancement of power electronics, motor control microprocessors, and novel motor control theories, asynchronous motor control technology is gaining popularity. When adopting vector control technology, asynchronous motors have a simple construction, cheap cost, and performance equivalent to DC motor control. Asynchronous motor vector control also provides greater accuracy, a larger speed-regulating range, and faster response [2].

In the 1970s, vector control was carried out by Blaschke, known as Field Oriented Control (FOC) [38]. Its principle consists in eliminating the coupling between the inductor and the armature of the asynchronous induction machine, therefore it makes it possible to obtain an operation comparable to that of a DC machine. However, experience has shown the weaknesses of this method in the face of parameter uncertainties, whether they are measured, such as motor speed, or vary during operation, such as rotor and stator resistances.

The application of FOC requires the use of a flux regulator requires a flux sensor, which is often very expensive. Its assembly in drive systems is very delicate and requires a great deal of precision to achieve effective results. In order to eliminate this obstacle, it is necessary to apply the techniques of the automatic, allowing the reconstruction of flux. This process is called the estimator.

The main objective of this work is the evaluation by numerical simulation of the performance of the DFOC by orientation of the rotor flux of an IM associated with an introduction to fuzzy logic

The stations that I went through to accomplish this theses began with the first chapter, which was devoted to the modeling of the Voltage-controlled Induction Machine, the model adopted is based on the transformation of PARK, the application of the latter to the asynchronous induction machine will allow to have a two-axis model (d, q) represents the image of the three-phase model (a, b, c), the model will be tested by simulation, and evaluated through the various results, in parallel a frequency converter has been modeled, in terms of the voltage inverter, controlled by PWM with hysteresis, the machine-converter association will be simulated to see the impacts of this converter on the machine.

Then I crossed the next station, which is the actress of the second chapter, which was devoted to the application of DFOC by orientation of the rotor flux, we will also present the regulation of the speed of the IM by the classic PI controller.

Afterwards the road brought me to the last station, which was represented in the third chapter, where I dealt with an introduction to Fuzzy Logic. The basic principles of control and regulation by fuzzy logic are then studied. FLC is finally applied to design a speed governor and achieve robust control.

:Chapter_I :

" Modeling and Simulation of The Induction Machine"

I. 1 - Introduction:

Because the energy supply has been increasingly strained since the turn of the century, it is critical to preserve power [2].

In this chapter, we will present the constitution and the operating principle of the asynchronous motor also its three-phase mathematical model, its transformation into the two-phase system and the modeling of its power supply. Then, we will give the simulation results of the asynchronous motor supplied directly by the three-phase power system via a converter. [1]

I. 2 - Historical review of induction machine: [20]

Faraday discovered the electromagnetic induction law around 1831 and Maxwell formulated the laws of electricity (or Maxwell's equations) around 1860. The knowledge was ripe for the invention of the induction machine which has two fathers: Galileo Ferraris (1885) and Nicola Tesla (1886). Their

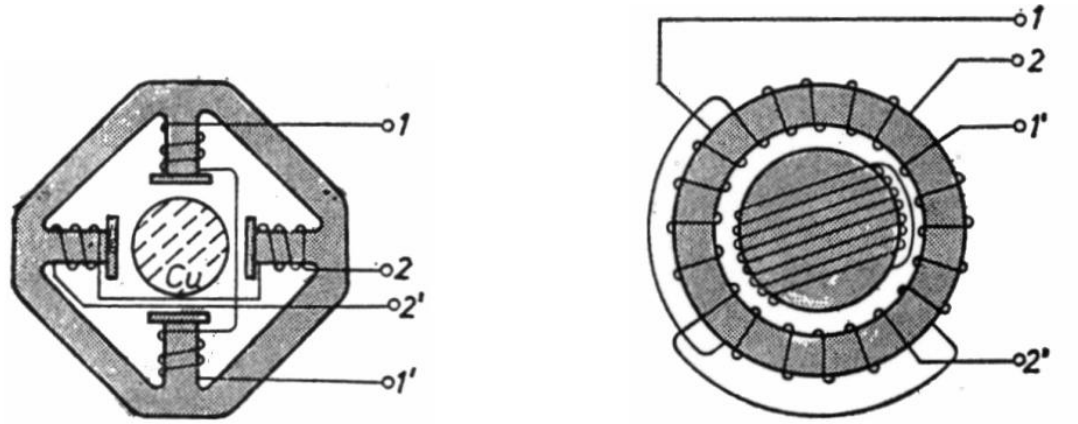


Fig I- 1 : Ferrari's induction motor (1885) and Tesla's induction motor (1886).

Both motors have been supplied from a two-phase **AC**. power source and thus contained two phase concentrated coil windings 1-1' and 2-2' on the ferromagnetic stator core.

In Ferrari's patent the rotor was made of a copper cylinder, while in the Tesla's patent the rotor was made of a ferromagnetic cylinder provided with a short-circuited winding.

Though the contemporary induction motors have more elaborated topologies Fig (I-1) and their performance is much better, the principle has remained basically the same.

That is, a multiphase AC. stator winding produces a traveling field which induces voltages that produce currents in the short-circuited (or closed) windings of the rotor. The interaction between the stator produced field and the rotor induced currents produces torque and thus operates the induction motor. As the torque at zero rotor speed is nonzero, the induction motor is self-starting. The three-phase AC. power grid capable of delivering energy at a distance to induction motors and other consumers has been put forward by DolivoDobrovolsky around 1880.

However, at least for transportation, the DC motor took over all markets until around 1985 when the IGBT PWM inverter was provided for efficient frequency changers. This promoted the induction motor spectacular comeback in variable speed drives with applications in all industries.

I.2.1- Historical review of induction machine:

Following the oil crises of the 1970s and increased global awareness of climate change, several nations began to consolidate the idea of sustainable development, which is based on, among other things, energy efficiency and energy management. Electric motor drive systems account for about 68 % of global power consumption in the industrial sector and 46 % of global electricity consumption. As a result, improving the industry's energy efficiency and lowering both energy consumption and production costs requires a focus on electric motor efficiency and reliability [3].

Electrical machines are devices which convert electrical energy into mechanical and vice versa. Based on the type of current that an electrical machine operates with (direct – DC or alternating – AC), it can be classified as either a DC machine or an AC machine. DC machines require regular and thorough maintenance. In addition, they typically have lower efficiency than AC machines. Therefore, AC machines are of more significant research interest. The AC machines are further divided into synchronous and induction (or asynchronous) machines. Synchronous machines operate at a constant speed regardless of load, while the rotational speed of induction machines decreases as the load increases. Both synchronous and induction machines can be used as generators or as motors. Generators convert mechanical energy into electrical energy, while motors convert electrical energy into mechanical energy [5].

This led to a great diversity of electric motors and an increase in their wide spread according to their uses and the most appropriate for each of them. Therefore, it can be divided into two groups, as shown in the Fig (I-2): [6]

- Commutator motors (also known as DC motors)
- Commutatorless motors (known as AC motors)

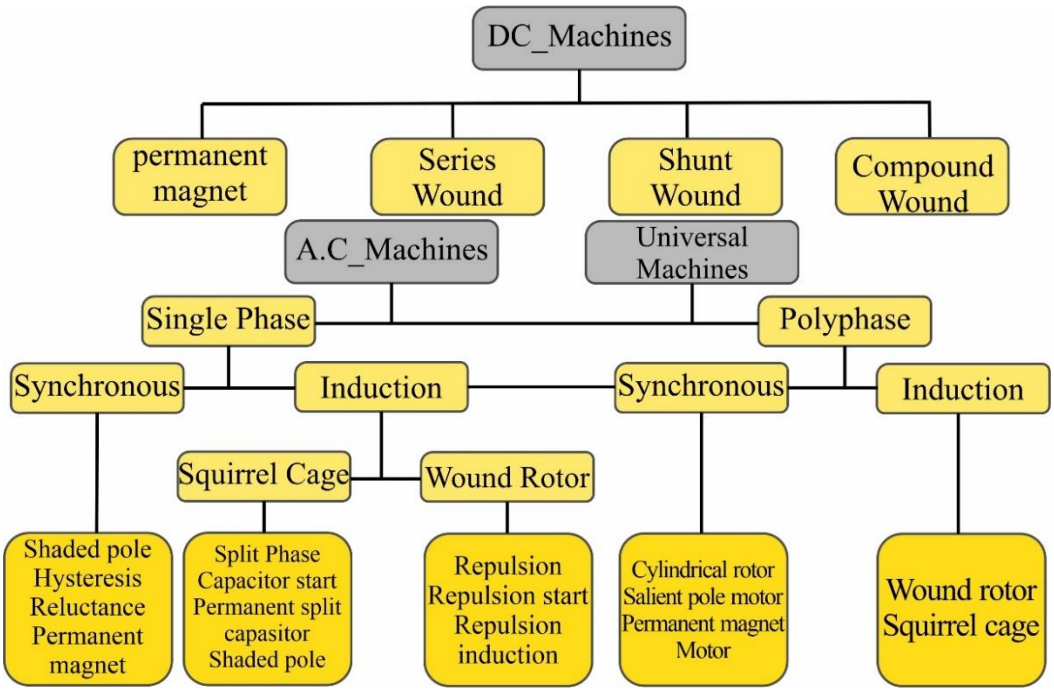


Fig I- 2 : Classification of electric motors

I.2.2- Induction machine definition:

The asynchronous machine is an alternating current machine, the rotor always lags behind the stator field because of slip, the asynchronous machine is said to be (induction machine) because the energy is transferred from the stator to the rotor or vice versa by induction electromagnetic. [7]

Among the asynchronous machines, we can distinguish two types:

- Induction machines.
- Collector's machines.

I.2.3- Induction machine types:

Depending on the construction of the rotor circuit there are two types of induction motors: [8]

I.2.3.a - Squirrel cage induction motor:

Rotors is very simple and consist of bars of aluminum (or copper) with shorting rings at the ends.

I.2.3.b - Wound rotor induction motor:

Rotor consists of three phase windings (star connected) with terminals brought out to slip rings for external connections. Squirrel cage type is more common compared to the wound rotor type due to:

a. Robust, as no brushes, no contacts on the rotor shaft.
b. Simple in construction and easy to manufacture.
c. Almost maintenance-free, except for bearing and other mechanical parts.
d. High efficiency as rotor has very low resistance and thus low copper loss.

I.2.4- Constitution of the induction machine:

There are two main types of components which are used in induction motor manufacturing as follows: [8]

a) Active components: which are classified into two categories:

- Magnetic materials (0.5 mm electrical steel).
- Electrical materials (copper wires, insulations, bars, end rings, slip rings, brushes, and lead wires).

b) Constructional components: like frame, end shields, shaft, bearings, and fan. These components are shown in Fig (I-3).

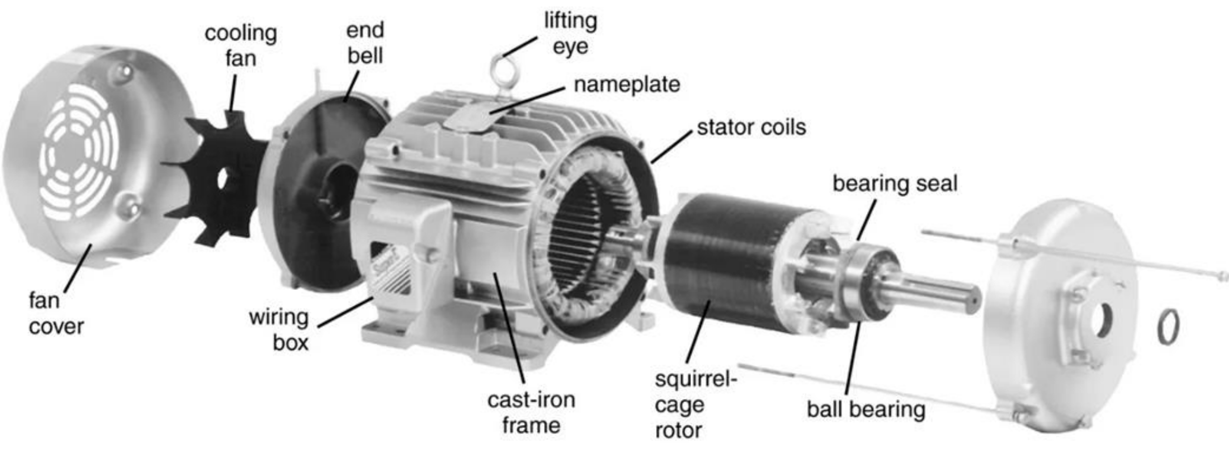


Fig I- 3 : Parts of squirrel cage induction motor [9]

I.2.4.a - Stator construction:

The stator is made up of several thin laminations (0.5 mm) of electrical steel (silicon steel), they are punched and clamped together to form a hollow cylinder (stator core) with slots, as shown in Fig (I-4). Coils of insulated wires are inserted into these slots. Each group of coils, together with the core that it surrounds, forms an electromagnet, forms an electromagnet (a pair of poles). The number of poles of an induction motor depends on the internal connection of the stator windings. [10]

The stator of an induction motor is identical to that of a synchronous motor (SM), three windings coupled in star or in delta and staggered between them by $2\pi/3$ which are supplied by a system of balanced voltages.

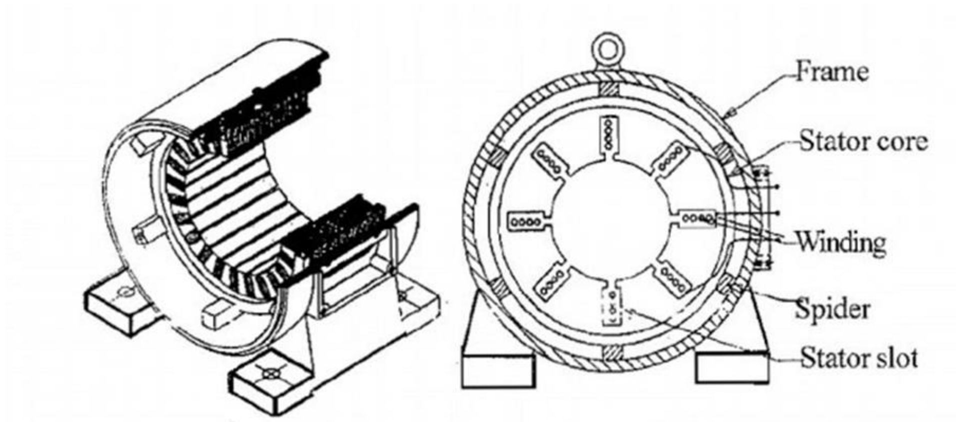


Fig I- 4 : Stator of three-phase induction motor [11]

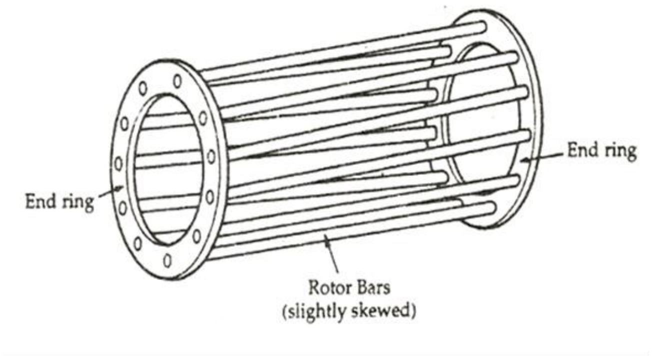
I.2.4.b - Rotor construction:

The rotors of asynchronous machines can be of two types: wound or squirrel cage and the rotor is not electrically linked to any source of energy. (neither continuous nor alternative) which greatly simplifies its construction, there are two types of rotor [13].

The squirrel cage rotor is made up of several thin electrical steel lamination (0.5mm) with evenly spaced bars, which are made up of aluminum or copper, along the periphery. In the most popular type of rotor (squirrel cage rotor), these bars are connected at ends mechanically and electrically by the use of end rings as in Fig (I-5). Almost 90 % of induction motors have squirrel cage rotors. The rotor slots are not exactly parallel to the shaft. Instead, they are given a skew for two main reasons, firstly to make the motor run quietly by reducing magnetic hum and to decrease slot harmonics, secondly to help reducing the locking tendency of the rotor (the rotor teeth tend to remain locked under the stator teeth due to direct magnetic attraction between the two). The rotor is mounted on the shaft using bearings on both ends. [10] [13]

The wound rotor has a set of windings on the rotor slots which are not short circuited, but they are terminated to a set of slip rings. These are helpful in adding external resistors and contactors, as in Fig (I-5). [13]

The motor rotor supports a winding similar to that of the MS stator, three-phase winding offset by 2π/3 with the same number of poles as that of the MS stator. These 3 windings are star-coupled and short-circuited on themselves. [13]



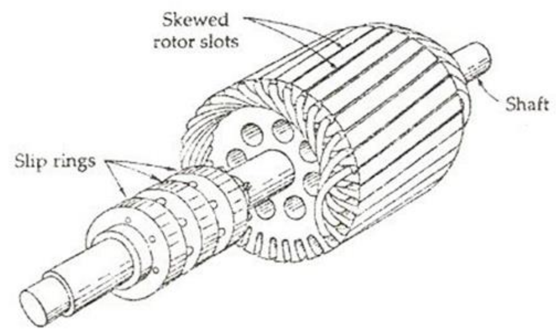


Fig I- 5 : “Squirrel cage rotor” and “Phase wound rotor” [12]

I.2.5- Principle of operation of an induction motor:

When 3-phase supply is fed to the stator winding of a 3-phase wound induction motor, a resultant rotating magnetic field at constant angular velocity is produced in the stator core.

Let this field is revolting in an anti-clockwise direction at synchronous speed n_s.

Where,

$$\boldsymbol{n_s = \frac{120 \cdot f}{P}} \quad [\boldsymbol{rpm}] \text{--(I- 1)}$$

Where f is the frequency of the input electrical power, and P is the number of poles of the induction machine. [14]

The rotating magnetic field is cut by the stationary rotor conductors and an emf is induced in the rotor conductors. As the rotor conductors are short circuited, current flows through them, Furthermore, a resultant field is produced by the rotor current carrying conductors. This field tries to come in line with the stator rotating field, as a result, an electromagnetic torque is developed and rotor starts rotating in same direction as that of stator rotating field. The rotor then run at a mechanical speed close to and less than the synchronous speed as it tries to attain synchronous speed but never reach. It is because if the rotor revolves at the synchronous speed then the relative speed between rotating stator field and rotor will be zero, therefore, neither emf will be induced in rotor conductors or current nor rotor field and hence no torque will be produced. Thus, an induction motor can never run at synchronous speed. [15]

This difference in rotational speed of the rotor compared to that of the rotating field brings us back to defining a parameter which characterizes asynchronous machines, which is called slip. [20]

The slip " s " is a quantity that considers the difference between the rotational speed of the asynchronous machine and the synchronous speed, since it is a number confined to the following numerical range $\mathbf{0 < s < 1}$ It is expressed as a percentage and defined by: [15]

$$s = \frac{n_s - n}{n_s} \qquad \text{(I- 2)}$$

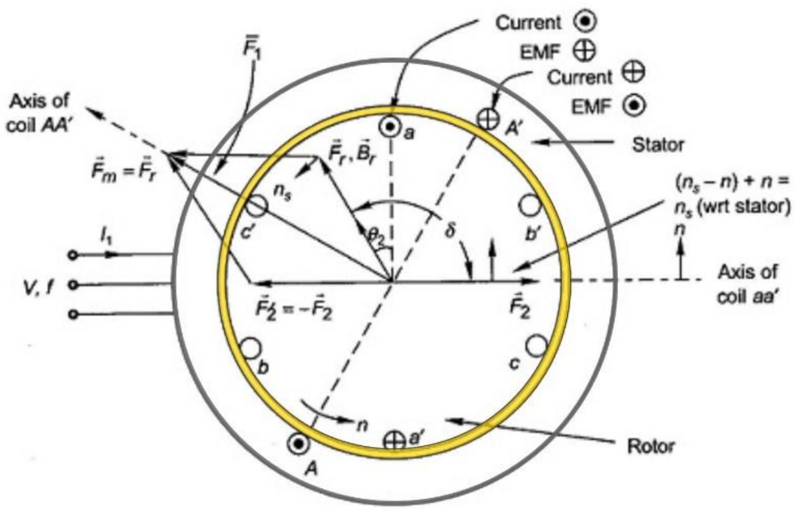


Fig I- 6 : Principle operation of an induction motor [17]

I.2.6- Advantages and disadvantages of the induction machine:

Like other electric machines, the IM has some advantages and disadvantages which are related to several factors: its structure, control strategy and applications. [16]

I.2.6.a - The advantages:

- The most important advantage of an induction motor is that its construction is quite simple in nature. The construction of the Stator is similar in both Synchronous motors as well as induction motors. However, a slip ring is required to feed DC Supply to the Rotor in the case of a Synchronous Generator. These Slip rings are not required in a Squirrel cage induction motor because the windings are permanently short circuited. When compared with a DC Motor, the induction motor does not have Brushes and hence, maintenance required is quite low. This leads to a simple construction.
- The working of the motor is independent of the environmental condition. This is because the induction motor is Robust and mechanically strong.

- A Squirrel cage induction motor does not contain Brushes, Slip rings and Commutators. Due to this reason, the cost of the motor is quite low. However, Slip Rings are used in Wound type induction motor to add external resistance to the rotor winding.
- Due to the absence of Brushes, there are no sparks in the motor. It can also be operated in hazardous conditions.
- An induction motor is a highly efficient machine with full load efficiency varying from 85 to 97 percent.

I.2.6.b - The disadvantages:

- A single-phase induction motor, unlike a 3-phase induction motor, does not have a self-starting torque. Auxiliaries are required to start a single-phase motor.
- During light load conditions, the power factor of the motor drops to a very low value. This is because during the start, the motor draws a large magnetizing current to overcome the reluctance offered by the air gap between the Stator and the Rotor. Also, the induction motor will take very less current from the supply main. The vector sum of Load current and Magnetizing current lags the voltage by around 75-80 degrees and hence, the power factor is low. Due to high magnetizing current, the copper losses of the motor increase. This in turn leads to decrease in the efficiency of the motor.
- Speed control of an induction motor is very difficult to attain. This is because a 3-phase induction motor is a constant speed motor and for the entire loading range, the change in speed of the motor is very low.
- Induction motors have high input surge currents, which are referred to as Magnetizing Inrush currents. This causes a reduction in voltage at the time of starting the motor.
- Due to poor starting torque, the motor cannot be used for applications which require high starting torque.

I.2.7- Fields of application of induction motors:

Three-phase AC induction motors have various uses in commercial and industrial applications. The two types of three-phase induction motors are- squirrel cage and slip ring motors. The features which make the squirrel cage motors widely applicable are mainly their simple design and rugged construction. With external resistors, the slip ring motors can have high starting torque. [19]

Three-phase induction motors are used extensively in domestic and industrial appliances because these are rugged in construction requiring little to no maintenance, comparatively cheaper, and require supply only to the stator. Applications of Three-phase motors have many uses, but we will suffice with some of them as follows: [18]

(Lifts, Cranes, Hoists, Large capacity exhaust fans, Driving lathe machines, Crushers, Oil extracting mills, Textile … etc.)

I.3.1- Simplifying assumptions:

Because the machine is a complex and nonlinear system, several phenomena are introduced during operation. This makes the machine difficult to control and model. To overcome this problem, we consider the following simplifying assumptions: [21] [22]

- Balanced rotor windings are assumed for all cases, and the three-phase machine equations, are derived upon the additional assumption that the stator windings are also balanced.
- It is assumed that the coefficient of mutual inductance between any stator winding and any rotor winding is a co-sinusoidal function of the electrical angle between the axes of the two windings.
- It IS further assumed that the rotor is smooth and that the self-inductances of all the windings are independent of the rotor position.
- No skin effect phenomenon.
- The effects of saturation, hysteresis, and eddy currents are neglected.
- Moreover, it will be considered that the magnetomotive force, created by the stator and rotor windings, has a sinusoidal distribution along the air-gap.

I.3.2- Modeling of the IM in the "a, b, c" three-phase plane:

Consider a three-phase induction machine with stator and rotor represented schematically by Fig (I-7), and whose phases are identified respectively by **SA**, **SB**, **SC**. The electrical angle $\boldsymbol{\theta}$ variable as a function of time defines the instantaneous relative position between the magnetic axes of the **SA** and **Ra** phases chosen as reference axes. [21] [23] [24]

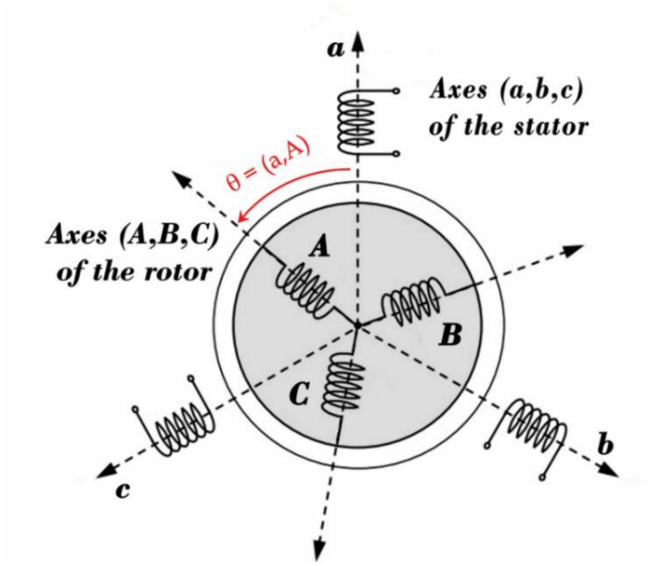


Fig I- 7 : Schematic representation of a three-phase asynchronous machine.

I.3.3- General equations of the three-phase induction machine in “a, b, c” plane:

Under these conditions, if it is considered that the induction motor is three-phase at the stator and at the rotor. The three types of equations reflecting the behavior of the motor are:

I.3.3.1- The electrical equations:

The voltage equations of the three stator phases and the three rotor phases are:

$$\begin{cases} V_{sa} = R_s i_{sa} + \frac{d}{dt}\varphi_{sa} \\ V_{sb} = R_s i_{sb} + \frac{d}{dt}\varphi_{sb} \\ V_{sc} = R_s i_{sc} + \frac{d}{dt}\varphi_{sc} \end{cases} \qquad \text{(I- 3)}$$

$$\begin{cases} V_{ra} = R_r i_{ra} + \frac{d}{dt}\varphi_{ra} \\ V_{rb} = R_r i_{rb} + \frac{d}{dt}\varphi_{rb} \\ V_{rc} = R_r i_{rc} + \frac{d}{dt}\varphi_{rc} \end{cases} \qquad \text{(I- 4)}$$

designating by:

V_{sa}, V_{sb}, V_{sc} : Voltages applied to the three stator phases.

i_{sa}, i_{sb}, i_{sc} : Currents which cross the three stator phases.

$\varphi_{sa}, \varphi_{sb}, \varphi_{sc}$: Flux through these windings.

V_{ra}, V_{rb}, V_{rc} : Rotor voltages.

i_{ra}, i_{rb}, i_{rc} : Rotor currents.

$\varphi_{ra}, \varphi_{rb}, \varphi_{rc}$: Rotor Flux.

R_s : Resistance of a stator phase.

R_r : Resistance of a rotor phase.

Equations (I-3) and (I-4) can be written in the following matrix form:

I.3.3.1.a - <u>For the stator:</u>

$$\begin{bmatrix} V_{sa} \\ V_{sb} \\ V_{sc} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \varphi_{sa} \\ \varphi_{sb} \\ \varphi_{sc} \end{bmatrix} \qquad \text{(I- 5)}$$

Or in condensed form as follows:

$$[V_{s(abc)}] = [R_s][i_{s(abc)}] + \frac{d}{dt}[\varphi_{s(abc)}] \qquad \text{(I- 6)}$$

I.3.3.1.b - <u>For the rotor:</u>

$$\begin{bmatrix} V_{ra} \\ V_{rb} \\ V_{rc} \end{bmatrix} = \begin{bmatrix} R_r & 0 & 0 \\ 0 & R_r & 0 \\ 0 & 0 & R_r \end{bmatrix} \begin{bmatrix} i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \varphi_{ra} \\ \varphi_{rb} \\ \varphi_{rc} \end{bmatrix} \qquad \text{(I- 7)}$$

Or in condensed form as follows:

$$[V_{r(abc)}] = [R_r][i_{r(abc)}] + \frac{d}{dt}[\varphi_{r(abc)}] \text{ ----- (I- 8)}$$

I.3.3.2- The magnetic equations:

The simplifying hypotheses cited above lead to linear relations between the fluxes and the currents of the asynchronous machine, these relations are written in matrix form as follows:

I.3.3.2.a - For the stator:

$$\begin{bmatrix} \varphi_{sa} \\ \varphi_{sb} \\ \varphi_{sc} \end{bmatrix} = [L_s]\begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} + [M_{sr}]\begin{bmatrix} i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix} \text{ ----- (I- 9)}$$

Or in condensed form as follows:

$$[\varphi_{s(abc)}] = [L_s][i_{s(abc)}] + [M_{sr}][i_{r(abc)}] \text{ ----- (I- 10)}$$

I.3.3.2.b - for the rotor:

$$\begin{bmatrix} \varphi_{ra} \\ \varphi_{rb} \\ \varphi_{rc} \end{bmatrix} = [L_r]\begin{bmatrix} i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix} + [M_{rs}]\begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} \text{ ----- (I- 11)}$$

Or in condensed form as follows:

$$[\varphi_{r(abc)}] = [L_r][i_{r(abc)}] + [M_{rs}][i_{s(abc)}] \text{ ----- (I- 12)}$$

Such as:

$$[M_{sr}] = [M_{rs}]^T \text{ ----- (I- 13)}$$

$[L_s]$: Stator inductance matrix.

$[L_r]$: Matrix of rotor inductors.

$[M_{sr}]$: Matrix of stator mutual inductances.

$[M_{rs}]$: Matrix of rotor mutual inductances.

Where:

$$[L_s] = \begin{bmatrix} l_s & M_s & M_s \\ M_s & l_s & M_s \\ M_s & M_s & l_s \end{bmatrix} \text{ ----- (I- 14)}$$

$$[L_s] = \begin{bmatrix} l_s & M_s & M_s \\ M_s & l_s & M_s \\ M_s & M_s & l_s \end{bmatrix} \text{ ----- (I- 15)}$$

also:

$$[M_{sr}] = [M_{rs}]^T = M_0 \begin{bmatrix} \cos(\theta) & \cos\left(\theta + +\frac{2\pi}{3}\right) & \cos\left(\theta - \frac{2\pi}{3}\right) \\ \cos\left(\theta - \frac{2\pi}{3}\right) & \cos(\theta) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ \cos\left(\theta + \frac{2\pi}{3}\right) & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos(\theta) \end{bmatrix} \quad \text{(I- 16)}$$

With:

l_s : Self-inductance of a stator phase.

l_r : Self-inductance of a rotor phase.

M_s : Mutual inductance between stator phases.

M_r : Mutual inductance between rotor phases.

θ : Electric angle defines the instantaneous relative position between the stator axes and the rotor axes which are chosen as reference axes.

M_0 : Maximum mutual inductance between stator phase and corresponding rotor phase

Finally, we get the three-phase asynchronous model as follows:

$$\begin{cases} [V_{s(abc)}] = [R_s][i_{s(abc)}] + \frac{d}{dt}\{[L_s][i_{s(abc)}] + [M_{sr}][i_{r(abc)}]\} \\ [V_{r(abc)}] = [R_r][i_{r(abc)}] + \frac{d}{dt}\{[L_r][i_{r(abc)}] + [M_{rs}][i_{r(abc)}]\} \end{cases} \text{------- (I- 17)}$$

I.3.3.3- Mechanical equations:

The study of the characteristics of the asynchronous machine introduces variation not only of the electrical parameters (voltage, current, flux) but also of the mechanical parameters (torque, speed):

$$C_e = p[i_{s(abc)}]^T \frac{d}{dt}[M_{sr}][i_{r(abc)}] \text{-- (I- 18)}$$

To have a complete model of the machine it is necessary to introduce the equation of motion of the machine is expressed as follows:

$$J\frac{d}{dt}\Omega_r = C_e - C_r - f_r \cdot \Omega_r \text{ --- (I- 19)}$$

J : Moment of inertia of rotating masses.

C_r : Resistive torque imposed on the machine shaft.

Ω_r : Rotor speed.

C_e : Electromagnetic torque.

f_r : Coefficient of viscous friction.

$\{f_r \cdot \Omega_r\}$: Viscous friction torque term.

The analytical resolution in this frame is very difficult, because the system of equations has variable coefficients as a function of θ (angle of rotation of the machine) This will lead to the use of the PARK transformation which will make it possible to make these parameters constant.

I. 3 - PARK model of induction machine:

Due to the existence of continuous trigonometric terms in the matrix of mutual inductances $[M_{sr}]$, the coefficients of the differential equations are variable and the analytical resolution of the system comes up against practically insurmountable difficulties to obtain a system of equations with constant coefficients, a three-phase machine can be represented by an equivalent two-phase machine using axis transformation. A three-phase machine and its equivalent two-phase machine is shown in Fig (I-8). Here d are direct and q quadrature axis of the rotor. The axis transformation relates current or voltages on (a, b, c) axes with the currents or voltages on (d, q, O) axes. This transformation, that relates the voltages, currents, and flux linkages associated with the stator winding, with variables associated with fictitious windings on d and q axis on the rotor, was first proposed by R. H. Park in 1920s. [24]

- Direct along the axis (d).
- Quadrature (transverse) along the axis (q).
- Homopolar (O).

The purpose of Park's transformation is to treat a wide range of machines in a unified way by reducing it to a single model. This conversion is often called axis transformation, a fact corresponding to the two windings of the original machine followed by a rotation, the electrically and magnetically equivalent windings. This transformation thus, for the purpose of making the mutual inductances of the model independent of the angle of rotation.

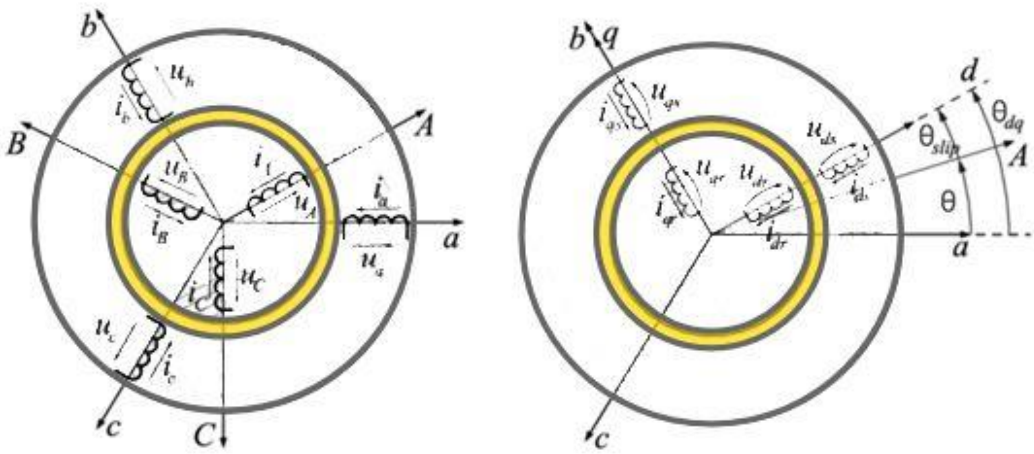


Fig I- 8 : Transition from a three-phase to a two-phase system

Where:

$\theta_d = \int \omega_d dt$: is any observation position between the two-phase axis systems with respect to the three-phase axis system.

I.4.1- Transformation of "PARK":

For the transition from the three-phase system to the two-phase system, we have the following equivalents: [25] [24]

The voltage equivalent : $[V_{(d,q,O)}] = [P(\theta_d)][V_{(a,b,c)}]$-------------------------------- (I- 20)

Current equivalent : $[i_{(d,q,O)}] = [P(\theta_d)][i_{(a,b,c)}]$ -------------------------------- (I- 21)

The flux equivalent : $[\varphi_{(d,q,O)}] = [P(\theta_d)][\varphi_{(a,b,c)}]$ ------------------------------ (I- 22)

Where:

$[P(\theta_d)]$: is the PARK matrix

In the case of an inverse passage, we have:

$$\begin{cases} [V_{(a,b,c)}] = [P(\theta_d)]^{-1}[V_{(d,q,O)}] \\ [i_{(a,b,c)}] = [P(\theta_d)]^{-1}[i_{(d,q,O)}] \\ [\varphi_{(a,b,c)}] = [P(\theta_d)]^{-1}[\varphi_{(d,q,O)}] \end{cases} \text{ -- (I- 23)}$$

The modified direct and inverse PARK transformation matrix is then written:

$$[P(\theta_d)] = \sqrt{\frac{2}{3}}\begin{bmatrix} \cos(\theta_d) & \cos(\theta_d - 2\pi/3) & \cos(\theta_d + 2\pi/3) \\ -\sin(\theta_d) & -\sin(\theta_d - 2\pi/3) & -\sin(\theta_d + 2\pi/3) \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \end{bmatrix} \text{----- (I- 24)}$$

The factor $\left(\sqrt{\frac{2}{3}}\right)$ is there to conserve the instantaneous electrical power also we can say it was chosen to give an invariant expression of the electromagnetic torque from the property of $[P]^{-1} = [P]^T$.

$$[P(\theta_d)]^T = \sqrt{\frac{2}{3}}\begin{bmatrix} \cos(\theta_d) & -\sin(\theta_d) & 1/\sqrt{2} \\ \cos(\theta_d - 2\pi/3) & -\sin(\theta_d - 2\pi/3) & 1/\sqrt{2} \\ \cos(\theta_d + 2\pi/3) & -\sin(\theta_d + 2\pi/3) & 1/\sqrt{2} \end{bmatrix} \text{----------- (I- 25)}$$

When the angle θ_d is assigned the value zero, the PARK transformation is called the Clarke transformation and the passage matrix is written as follows:

$$[C] = \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \\ 1/2 & 1/2 & 1/2 \end{bmatrix} \text{--- (I- 26)}$$

I.4.2- Application of the park transformation to the IM:

The application of the PARK transformation to the asynchronous machine corresponds to a transformation of the three coils (stator and rotor) to two equivalent coils taking up the same consideration or aspects in terms of flux, torque, current or at least an image which will be perfectly proportional to them. [24] [25]

Applying the PARK transformation to the electric, (I-3) and (I-4), and magnetic, (I-9), (I-11) matrix models, yields the following equations:

I.4.2.1 -Voltage equations:

In the PARK axis transformation (d, q) rotating at the angular velocity $\omega p = \frac{d\theta}{dt}$, the equations (I-3) and (I-4) are written:

$$\begin{cases} V_{sd} = R_s \cdot i_{sd} + \frac{d}{dt}\varphi_{sd} - \omega_s\varphi_{sq} \\ V_{sq} = R_s \cdot i_{sq} + \frac{d}{dt}\varphi_{sq} + \varphi_{sd} \\ 0 = R_r \cdot i_{rd} + \frac{d}{dt}\varphi_{rd} - (\omega_s - \omega) \cdot \varphi_{rq} \\ 0 = R_r \cdot i_{rq} + \frac{d}{dt}\varphi_{rq} + (\omega_s - \omega) \cdot \varphi_{rd} \end{cases} \quad \text{(I- 27)}$$

With: $\omega_s = \frac{d}{dt}\theta_s$ and $\omega = \frac{d}{dt}\theta$ and $\theta_r = \theta_s - \theta$

Equations (I-27) can be written in the form of the following matrices:

$$\begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} = \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \frac{d}{dt}\begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_s \\ \omega_s & 0 \end{bmatrix} \begin{bmatrix} \varphi_{sq} \\ \varphi_{sd} \end{bmatrix} \quad \text{(I- 28)}$$

$$\begin{bmatrix} V_{rd} \\ V_{rq} \end{bmatrix} = \begin{bmatrix} R_r & 0 \\ 0 & R_r \end{bmatrix} \begin{bmatrix} i_{rd} \\ i_{rq} \end{bmatrix} + \frac{d}{dt}\begin{bmatrix} \varphi_{rd} \\ \varphi_{rq} \end{bmatrix} + \begin{bmatrix} 0 & -(\omega_s - \omega) \\ (\omega_s - \omega) & 0 \end{bmatrix} \begin{bmatrix} \varphi_{rq} \\ \varphi_{rd} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \quad \text{(I- 29)}$$

I.4.2.2 -Flux equations:

$$\begin{cases} \varphi_{sd} = L_s \cdot i_{sd} + L_m \cdot i_{rd} \\ \varphi_{sq} = L_s \cdot i_{sq} + L_m \cdot i_{rq} \\ \varphi_{rd} = L_r \cdot i_{rd} + L_m \cdot i_{sd} \\ \varphi_{rq} = L_r \cdot i_{rq} + L_m \cdot i_{sq} \end{cases} \quad \text{(I- 30)}$$

Equations (I-30) can be written in the form of the following matrices:

$$\begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \end{bmatrix} = \begin{bmatrix} L_s & 0 \\ 0 & L_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \begin{bmatrix} L_m & 0 \\ 0 & L_m \end{bmatrix} \begin{bmatrix} i_{rd} \\ i_{rq} \end{bmatrix} \quad \text{(I- 31)}$$

$$\begin{bmatrix} \varphi_{rd} \\ \varphi_{rq} \end{bmatrix} = \begin{bmatrix} L_r & 0 \\ 0 & L_r \end{bmatrix} \begin{bmatrix} i_{rd} \\ i_{rq} \end{bmatrix} + \begin{bmatrix} L_m & 0 \\ 0 & L_m \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} \quad \text{(I- 32)}$$

With:

$L_s = l_s - M_s$, $L_r = l_r - M_r$: Self-cyclic inductance of stator and rotor respectively.

$L_m = \frac{3}{2} \cdot M_0$: Stator-rotor mutual cyclic inductance

I.4.2.3 -Mechanical equations:

The electromechanical torque represented in equation (I-18) becomes:

$$C_e = p \cdot \frac{L_m}{L_r} \cdot (\varphi_{rd} i_{sq} - \varphi_{rq} i_{sd}) \quad \text{(I- 33)}$$

The mechanical speed equation is represented by equation (I-34):

$$\begin{cases} J\frac{d}{dt}\Omega_r = C_e - C_r - f_r \cdot \Omega_r \\ J\frac{d}{dt}\Omega_r = p \cdot \frac{L_m}{L_r} \cdot (\varphi_{rd} i_{sq} - \varphi_{rq} i_{sd}) - C_r - f_r \cdot \Omega_r \\ \text{With: } \omega_r = p \cdot \Omega_r \end{cases} \quad \text{(I- 34)}$$

p : Number of pole pairs.

I. 4 - Choice of referential:

The isotropy of the asynchronous motor allows a great flexibility in the composition of the equations of the machine according to two axes (d, q) using the components of Park, this requires the use of a reference which makes it possible to simplify the expressions as much as possible analytical. [27] There are different possibilities for the choice of the axis reference, which is practically reduced to three orthogonal reference frames (two-phase systems), There are three important choices regarding the orientation of the axis coordinate system (d, q) which depend on the purpose of the application. [28]

In the following, the zero sequence components are assumed to be zero.

I. 5. 1 - Referential related to the stator:

In this frame of reference, the axes (d, q) are stationary with respect to the stator ($\omega_{coor} = 0$). This referential is best suited to work with instantaneous quantities and whose advantage does not require a transformation to the real system.

The use of this system makes it possible to study the starting and braking systems of induction Machines. [28] [29]

I. 5. 2 - Referential related to the rotor:

In this frame of reference, the axes (d, q) are immobile with respect to the rotor rotating at a speed ω therefore $(\omega_{coor} = \omega = p\omega)$ .The use of this frame of reference makes it possible to study transient systems in synchronous and asynchronous machines with non-symmetrical connection of the rotor circuits. [29]

I. 5. 3 - Referential related to the rotating field:

In this frame of reference, the axes (d, q) are stationary with respect to the electromagnetic field created by the stator windings, hence $(\omega_{coor} = \omega_s)$.

This frame of reference is generally used in order to be able to apply a command for speed, torque, etc. since the quantities in this frame of reference are of continuous form. [28] [29]

Where:

ω_{coor} : Angular speed of rotation of the two-phase axis system with respect to the three-phase axis system.

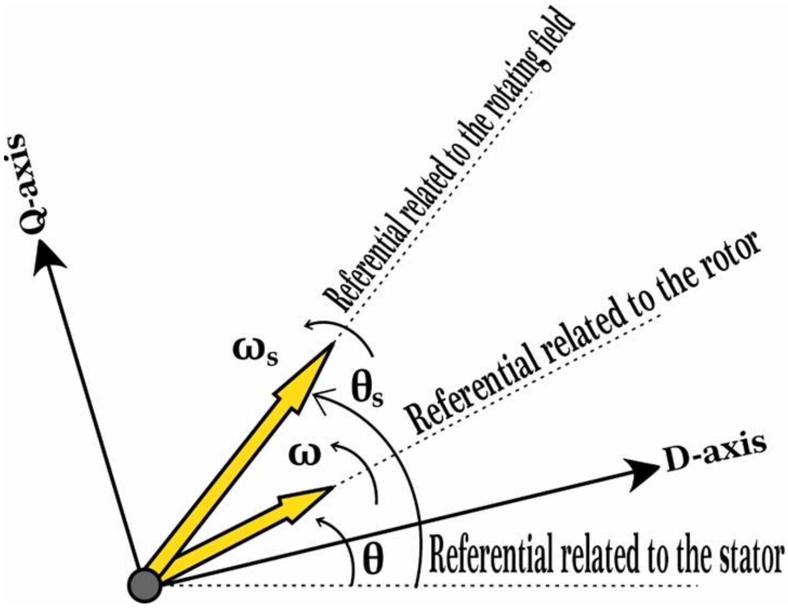


Fig I- 9 : Choice of repository

In my work, we use the reference frame linked to the rotating field $(\omega_{coor} = \omega_s)$ for the modeling and the control of the IM. In this case, the IM model becomes the same for voltage, magnetic field equations and mechanical equations, the equations respectively (I.35).

The electromagnetic torque can be derived from the co-energy expression or obtned using a power balance. This results in several expressions all of which are equal: [26] [29]

$$\begin{cases} C_e = p \cdot \frac{3}{2} \cdot \frac{M_0}{L_r} \cdot (i_{sq} i_{rd} - i_{rq} i_{sd}) \\ C_e = p \cdot \frac{3}{2} \cdot \frac{M_0}{L_r} \cdot (\varphi_{sd} i_{sq} - \varphi_{sq} i_{sd}) \\ C_e = p \cdot \frac{3}{2} \cdot \frac{M_0}{L_r} \cdot (\varphi_{rd} i_{sq} - \varphi_{rq} i_{sd}) \\ C_e = p \cdot \frac{3}{2} \cdot \frac{M_0}{L_r} \cdot (\varphi_{sq} i_{rq} - i_{rq} \varphi_{sd}) \end{cases} \quad \text{-- (I- 35)}$$

I. 5 - State space representation of the induction machine:

The state space model of a physical system is its mathematical representation using first order differential equations in the following format: [31]

$$[\dot{X}] = [A][X] + [B][U] \quad \text{--- (I- 36)}$$

With:

$[X]$: state vector representing all system variables.

$[U]$: input or control vector.

$[A]$: state matrix also called evolution matrix.

$[B]$: input matrix.

We consider the voltages (V_{sd}, V_{sq}) as control quantities, the stator currents (i_{sd}, i_{sq}), the rotor flux $(\varphi_{rd}, \varphi_{rq})$ and the mechanical speed Ω as state variables.

then:

The state vector: $[X] = [i_{sd}, i_{sq}, \varphi_{rd}, \varphi_{rq}]^T$.

The control vector: $[U] = [V_{sd}, V_{sq}]^T$.

From the systems of equations (I-27) to (I-30), we obtain the following system of equations: [30] [31]

$$\begin{cases} \frac{d}{dt} i_{sd} = -\lambda \cdot i_{sd} + \omega_s \cdot i_{sq} + \frac{k}{T_r} \cdot \varphi_{rd} + \omega \cdot k \cdot \varphi_{rq} + \frac{V_{sd}}{\sigma \cdot L_s} \\ \frac{d}{dt} i_{sq} = -\lambda \cdot i_{sq} - \omega_s \cdot i_{sd} + \frac{k}{T_r} \cdot \varphi_{rq} - \omega \cdot k \cdot \varphi_{rd} + \frac{V_{sq}}{\sigma \cdot L_s} \\ \frac{d}{dt} \varphi_{rd} = \frac{M}{T_r} \cdot i_{sd} - \frac{1}{T_r} \cdot \varphi_{rd} + (\omega_s - \omega_{slip}) \cdot \varphi_{rq} \\ \frac{d}{dt} \varphi_{rq} = \frac{M}{T_r} \cdot i_{sq} - \frac{1}{T_r} \cdot \varphi_{rq} - (\omega_s - \omega_{slip}) \cdot \varphi_{rq} \end{cases} \quad \text{------------ (I- 37)}$$

With:

$$\lambda = \frac{R_s}{\sigma \cdot L_s} + \frac{R_r \cdot M^2}{\sigma \cdot L_s \cdot {L_r}^2} \quad ; \; k = \frac{M}{\sigma \cdot L_s \cdot L_r} \quad ; \; T_r = \frac{L_r}{R_r} \quad ; \; \sigma = 1 - \frac{M^2}{L_s \cdot L_r} \quad ; \; M = L_m$$

σ : Blondel dispersion coefficient.

T_r : Rotor time constant.

state space form:

$$A = \begin{bmatrix} -\lambda & \omega_s & \frac{k}{T_r} & \omega_{slip} \cdot k \\ -\omega_s & -\lambda & -\omega_{slip} \cdot k & \frac{k}{T_r} \\ \frac{M}{T_r} & 0 & -\frac{1}{T_r} & (\omega_s - \omega) \\ 0 & \frac{M}{T_r} & -(\omega_s - \omega) & -\frac{1}{T_r} \end{bmatrix} \text{------------------------------------ (I- 38)}$$

$$B = \begin{bmatrix} \frac{1}{\sigma \cdot L_s} & 0 \\ 0 & \frac{1}{\sigma \cdot L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \text{--- (I- 39)}$$

The mechanical equation of motion and the electromagnetic torque equation are defined as follows:

$$\begin{cases} C_e = p \cdot \frac{M}{L_r} \cdot (\varphi_{rd} i_{sq} - \varphi_{rq} i_{sd}) \\ J \frac{d}{dt}\Omega_r = C_e - C_r - f_r \cdot \Omega_r \end{cases} \quad \text{(I- 40)}$$

I. 6 - Simulation of the induction machine model:

The asynchronous machine is normally supplied directly from the industrial network by a system of balanced three-phase voltages.

In certain applications for which speed variation is necessary, the motor will be supplied by a system of three-phase voltages or by a system of three-phase currents (injected) into the windings of the stator, via an electronic converter of power placed between the motor and the electrical industrial network. [32]

Fig (I-10) represents the block diagram of the model obtained, the latter will be simulated using the SIMULINK software under MATLAB.

The parameters of the IM used in this work are given in the Annex.

The simulation will be made in the (d, q) frame of reference for a test at rated load after a no-load start.

The supply voltages have been assumed to be perfectly sinusoidal with equal and constant amplitudes, they can be presented as follows:

$$\begin{cases} V_{sa} = \sqrt{2} \cdot V_S \sin(\omega_s t) \\ V_{sb} = \sqrt{2} \cdot V_S \sin\left(\omega_s t + \frac{2\pi}{3}\right) \\ V_{sc} = \sqrt{2} \cdot V_S \sin\left(\omega_s t - \frac{2\pi}{3}\right) \end{cases} \quad \text{(I- 41)}$$

With:

V_S : RMS voltage value

ω_s : Power pulse

I.7.1 - Asynchronous machine simulation block diagram:

The block diagram structure of this simulation is shown in Fig (I-10). The asynchronous motor presents as inputs the Park components of the supply voltage (V_{sd}, V_{sq}) as well as the resistive torque C_r and as output the speed Ω, the electromagnetic torque C_e , the stator currents and the rotor fluxes

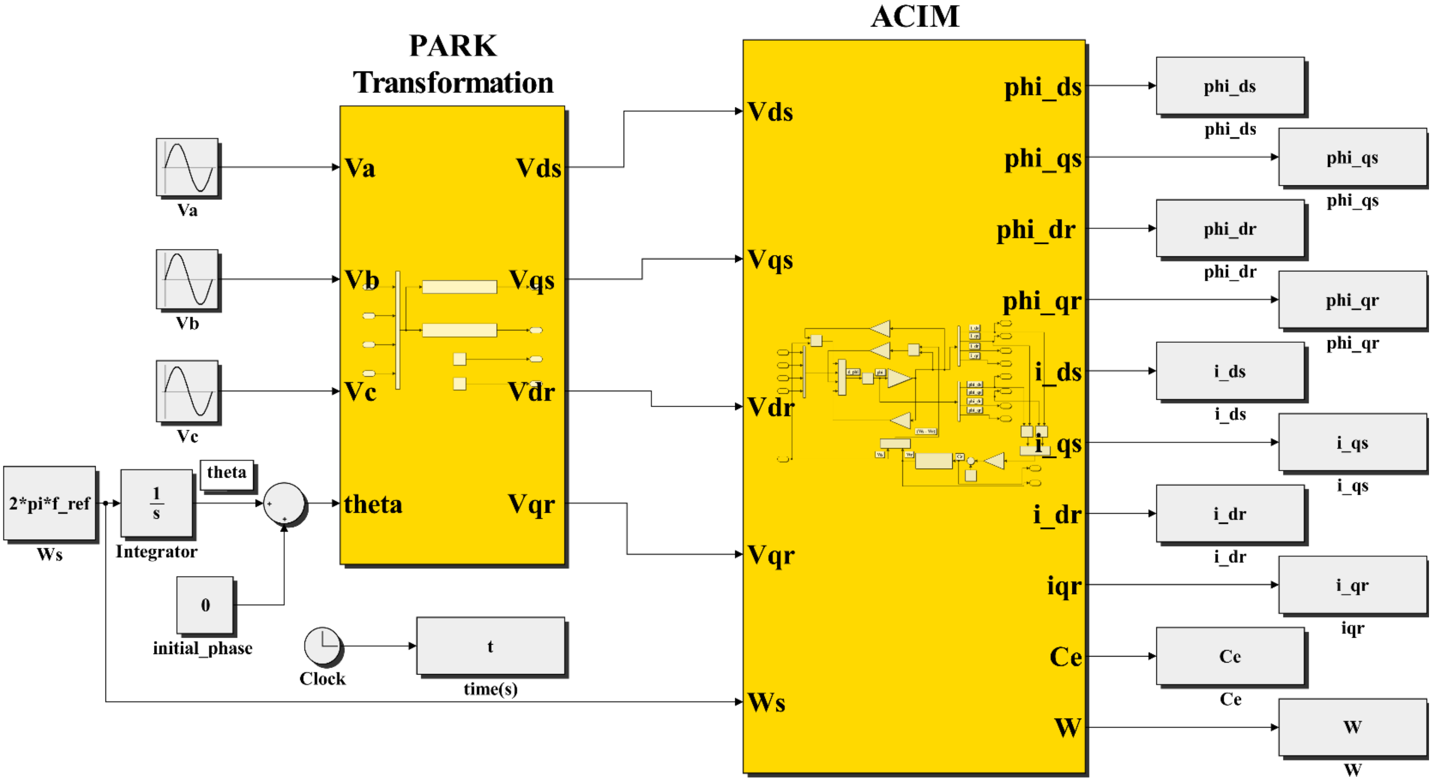


Fig I- 10 : Simulation diagram of an IM powered by an electrical network.

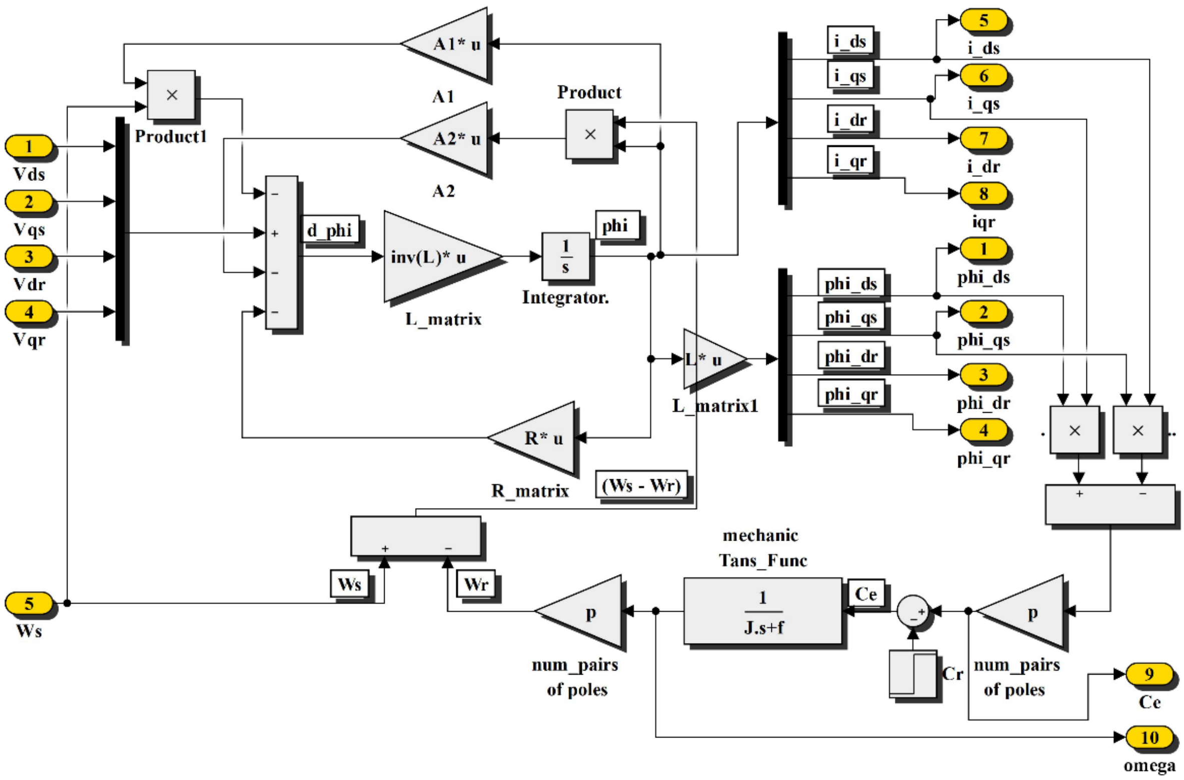


Fig I- 11 : Block diagram of the Induction machine model.

I.7.2 - Simulation and interpretation of results:

In the first step, I simulated the operation of the induction motor powered directly by the sine wave voltage source with $220/380V$ and $50Hz$ and with the application of the load $C_r = 25Nm$ at $t = 1s$, the simulation results are grouped in the Fig (I-12):

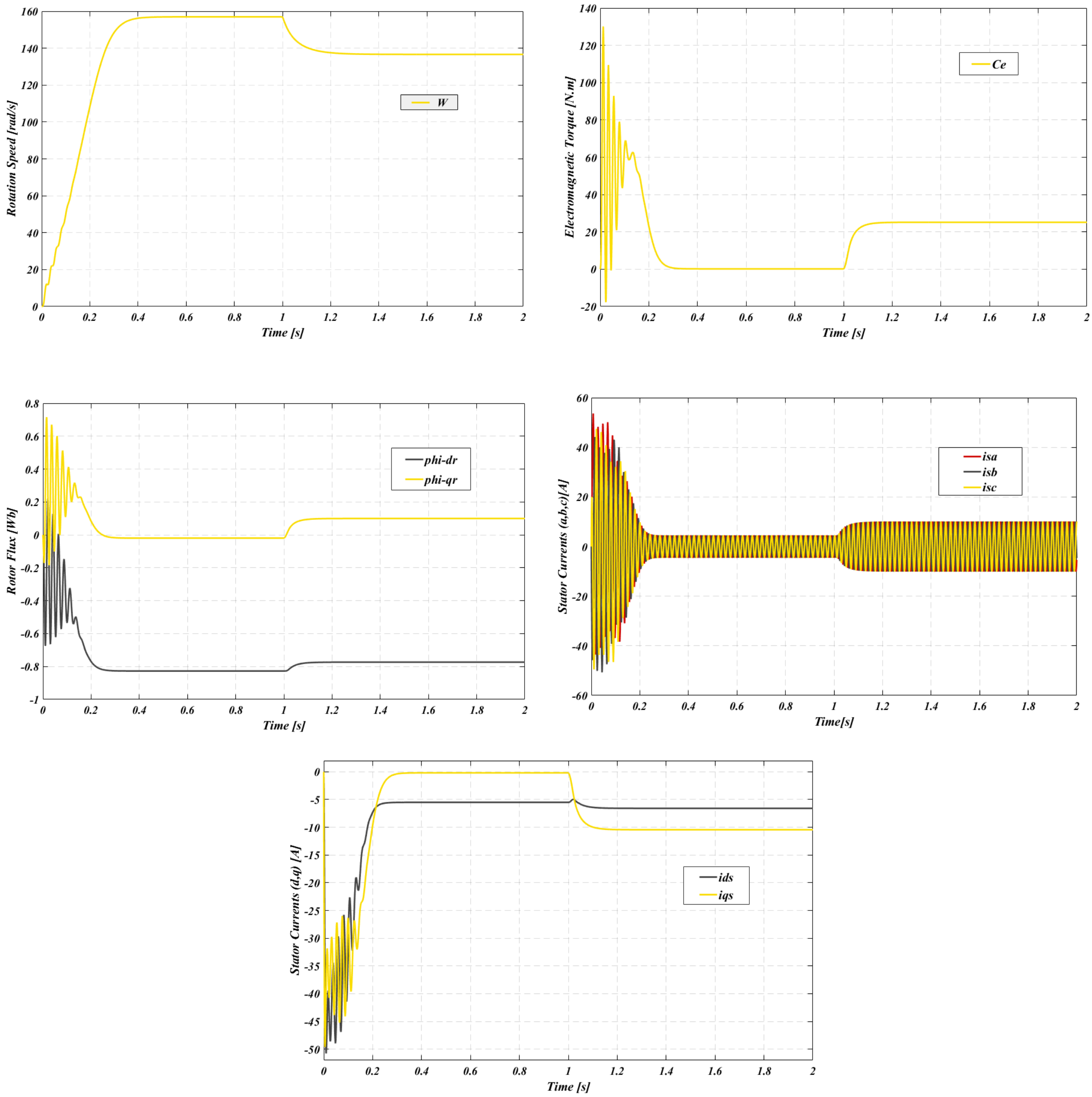


Fig I- 12 Simulation results of the IM powered by the sine wave voltage source with application of load $Cr = 25N.m$ at $t = 1s$

When the load is applied, the electromagnetic torque returns to its reference value to compensate for this stress with an almost instantaneous response. Before stabilizing at the resistive torque value, there is a decrease in rotor speed which results in very high slip. The stator currents changing according to the load applied to the motor shaft.

I. 8 - Three-phase voltage inverter:

The voltage inverter is a static converter which provides an alternating voltage of adjustable amplitude and frequency from a direct voltage source. It consists of switching cell generally transistor or thyristor for high powers. [33]

I.8.1 - Different types of inverter for supplying IM:

The inverter is put into operation through a direct current source that can be obtained by rectifying the voltage of the three-phase network. The direct current output of the rectifier and the direct current input of the inverter are connected by an intermediate circuit. Two types of connections are used: current source connections and voltage source connections. An inverter associated with a rectifier is then called a current or voltage inverter [34].

The inverter is the last part of the variable speed drive located before the motor. It provides variable electrical quantities to the motor. In all cases, the inverter is made up of semiconductors arranged in pairs in three arms. The semiconductors in the inverter switch on signals from the control circuit the switches can be made, depending on the power to be controlled, with MOS transistors, IGBTs or GTOs associated with an antiparallel diode to obtain current reversibility. [33] [34].

I.8.1.1 - Current inverter:

When operating as a current source, the rectifier supplies a constant current to the inverter; a smoothing inductor helps to keep the current constant. [34]

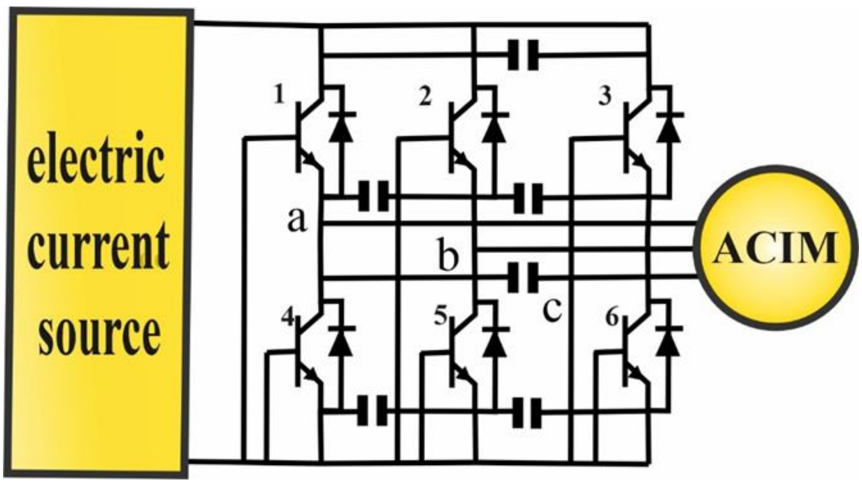


Fig I- 13 : Current inverter

I.8.1.2 - Voltage inverter:

When operating as a voltage source, the rectifier supplies a constant voltage to the inverter. The presence of a capacitor in the link circuit then helps to maintain a constant voltage at the input of the inverter. [34]

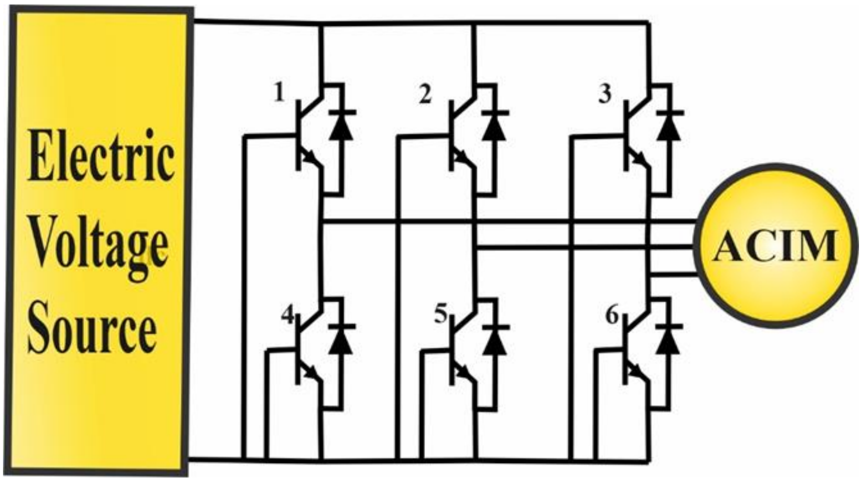


Fig I- 14 : Voltage inverter

I.8.2 - PWM voltage inverter with hysteresis: [33] [34]

To determine the closing and opening instants (switching instants) of the switches, the PWM technique (Pulse Width Modulation) is used, which consists of comparing the reference signal wave (modulating) of sinusoidal form at low frequency, to a triangular wave (carrier) signal of high frequency. The modulated signal is high when the modulator is higher than the carrier and is low when the modulator is lower than the carrier.

This type of inverter has the particularity of having a very good dynamic response, with a low level of torque ripple. Nowadays, it is the most coveted type of converter due to the improvement made on the electronic components and the innovations brought on the topologies of converters such as the multi-level inverters.

The instants of switching are determined by the points of intersection between the carrier and the modulating, Fig (I-15).

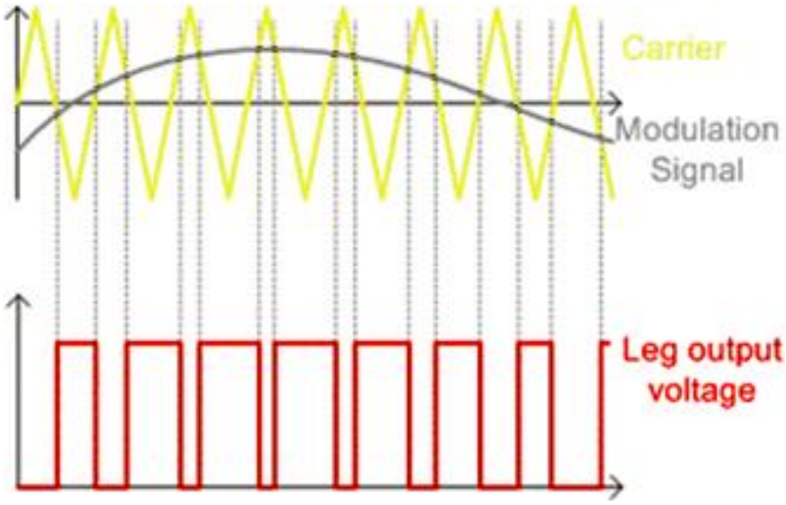


Fig I- 15 : Principle of PWM control

There are many reasons:

- The operation of the voltage inverter is greatly affected by the imperfections of the DC source, little by that of the load. For the current inverter, it is the opposite.
- The two inverters do not directly deliver a sinusoidal output voltage, so an output filter must be used. With the voltage inverter we know exactly what we have to filter
- the voltage to be filtered is imposed by the DC source. With the current inverter, the voltage to be filtered depends on the load.

I.8.2.1 - Principle:

The general principle consists in converting a modulating (reference voltage at the control level), generally sinusoidal, into a voltage in the form of successive slots of variable width. The commutation angles are calculated so as to eliminate a certain number of harmonics generated at the output of the inverter (power level). [34]

This technique is based on the comparison between two signals Fig (I-16)

The first, which is called the reference signal $\boldsymbol{V_{ref}}$, represents the image of the sinusoid that is desired at the output of the inverter. This signal is modulated in amplitude and frequency.

The second, which is called the $\boldsymbol{V_{por}}$ carrier signal, generally triangular, defines the rate of switching of the static switches of the inverter. It is a high frequency **HF** signal relative to the reference signal. [33] [34]

These two signals are compared, the comparison results are used to control the opening and closing of the switches of the power circuit.

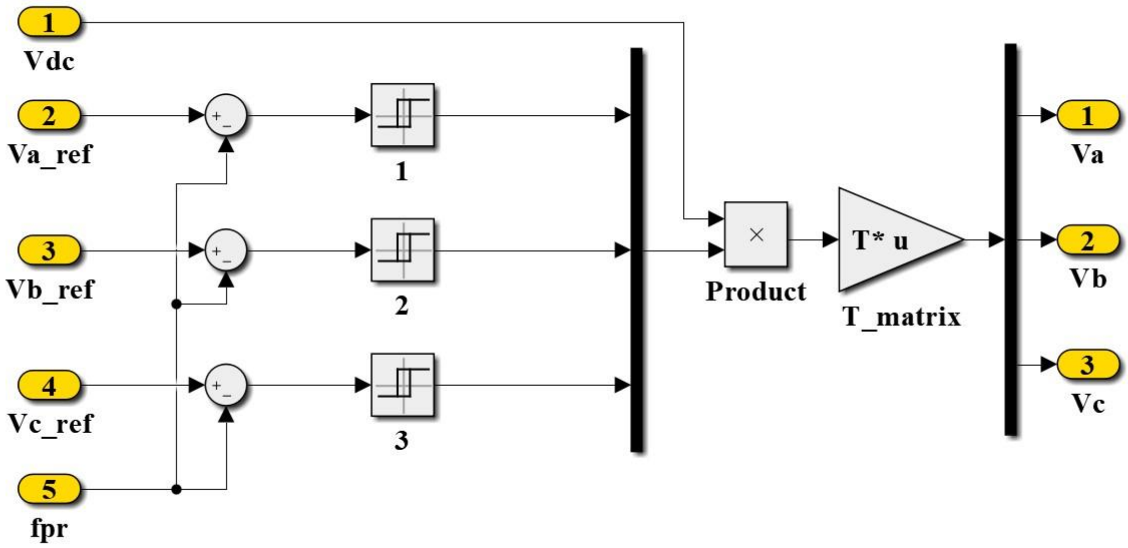


Fig I- 16 : PWM simulation diagram with hysteresis.

The intersection of these signals gives the instants of switching of the switches. The pulse wave is better than the rectangular wave if the frequencies:

$f_{por} > f_{ref}$

Fig (I-17) illustrates the simulation diagram of a voltage inverter controlled by the PWM with hysteresis:

We simulated our machine powered by hysteresis PWM voltage inverter. The simulations were carried out under MATLAB-Simulink. The parameters of the machine are given in the appendix. [34]

I. 9 - Modeling of the PWM inverter:

The diagram of a three-phase inverter supplying the IM. [34]

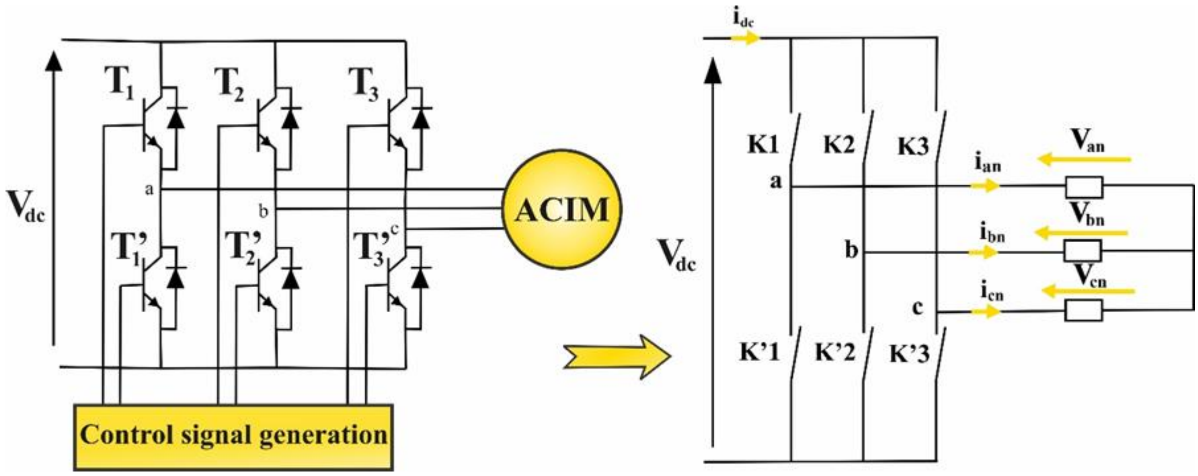


Fig I- 17 : "Diagram of the voltage inverter" and "Simplified diagram"

The simplified diagram of the voltage inverter associated with the machine is shown in Fig (1-17). In order to develop an operating model of the inverter, it is considered that each arm thereof consists of two switches $K1$ and $K2$ assumed to be perfect, that's to say the phenomena due to switching are neglected.

When the load neutral is isolated:

$$\begin{cases} V_{an} + V_{bn} + V_{cn} = 0 \\ i_{an} + i_{bn} + i_{cn} = 0 \end{cases} \quad \text{(I- 42)}$$

The inverter is modeled by associating with each arm a logic function F which determines its conduction states:

$$\begin{cases} F_1 = \begin{Bmatrix} 1: If \rightarrow \ K1\ closed\ , and \rightarrow \ K4\ open \\ 0: If \rightarrow \ K4\ closed\ , and \rightarrow \ K1\ open \end{Bmatrix} \\ F_2 = \begin{Bmatrix} 1: If \rightarrow \ K2\ closed\ , and \rightarrow \ K5\ open \\ 0: If \rightarrow \ K5\ closed\ , and \rightarrow \ K2\ open \end{Bmatrix} \\ F_3 = \begin{Bmatrix} 1: If \rightarrow \ K3\ closed\ , and \rightarrow \ K6\ open \\ 0: If \rightarrow \ K6\ closed\ , and \rightarrow \ K3\ open \end{Bmatrix} \end{cases} \quad \text{(I- 43)}$$

The potentials of the nodes (a, b, c) of the inverter with respect to the point N are given by the following relations:

$$\begin{cases} V_{an} = F_1 \cdot V_{dc} \\ V_{bn} = F_2 \cdot V_{dc} \\ V_{cn} = F_3 \cdot V_{dc} \end{cases} \quad \text{(I- 44)}$$

The phase-to-phase voltages of the inverter are deduced using the connection functions as follows:

$$\begin{cases} U_{ab} = V_{an} - V_{bn} = (F_1 - F_2) \cdot V_{dc} \\ U_{bc} = V_{bn} - V_{cn} = (F_2 - F_3) \cdot V_{dc} \\ U_{ca} = V_{cn} - V_{an} = (F_3 - F_1) \cdot V_{dc} \end{cases} \quad \text{(I- 45)}$$

We can also express simple voltages from phase-to-phase voltages as:

$$\begin{cases} V_{an} = \frac{U_{ab} - U_{ca}}{3} \\ V_{bn} = \frac{U_{bc} - U_{ab}}{3} \\ V_{cn} = \frac{U_{ca} - U_{bc}}{3} \end{cases} \quad \text{(I- 46)}$$

The expression of the phase-to-neutral voltages of the inverter by means of the connection logic functions is obtained from equations (I-46):

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} F_1 \\ F_2 \\ F_3 \end{bmatrix} \quad \text{(I- 47)}$$

V_{dc} : This is the DC supply voltage of the inverter.

$$\xrightarrow{then} [T] = \begin{bmatrix} 2/3 & -1/3 & -1/3 \\ -1/3 & 2/3 & -1/3 \\ -1/3 & -1/3 & 2/3 \end{bmatrix} \text{------------------------------------} \quad (I\text{- }48)$$

I. 10 - Simulation and interpretation of the results:

In the second step, the operation of the induction machine powered by a three-phase hysteresis inverter is simulated that's illustrates in the Fig (I-18), with the application of the same conditions as in the first step.

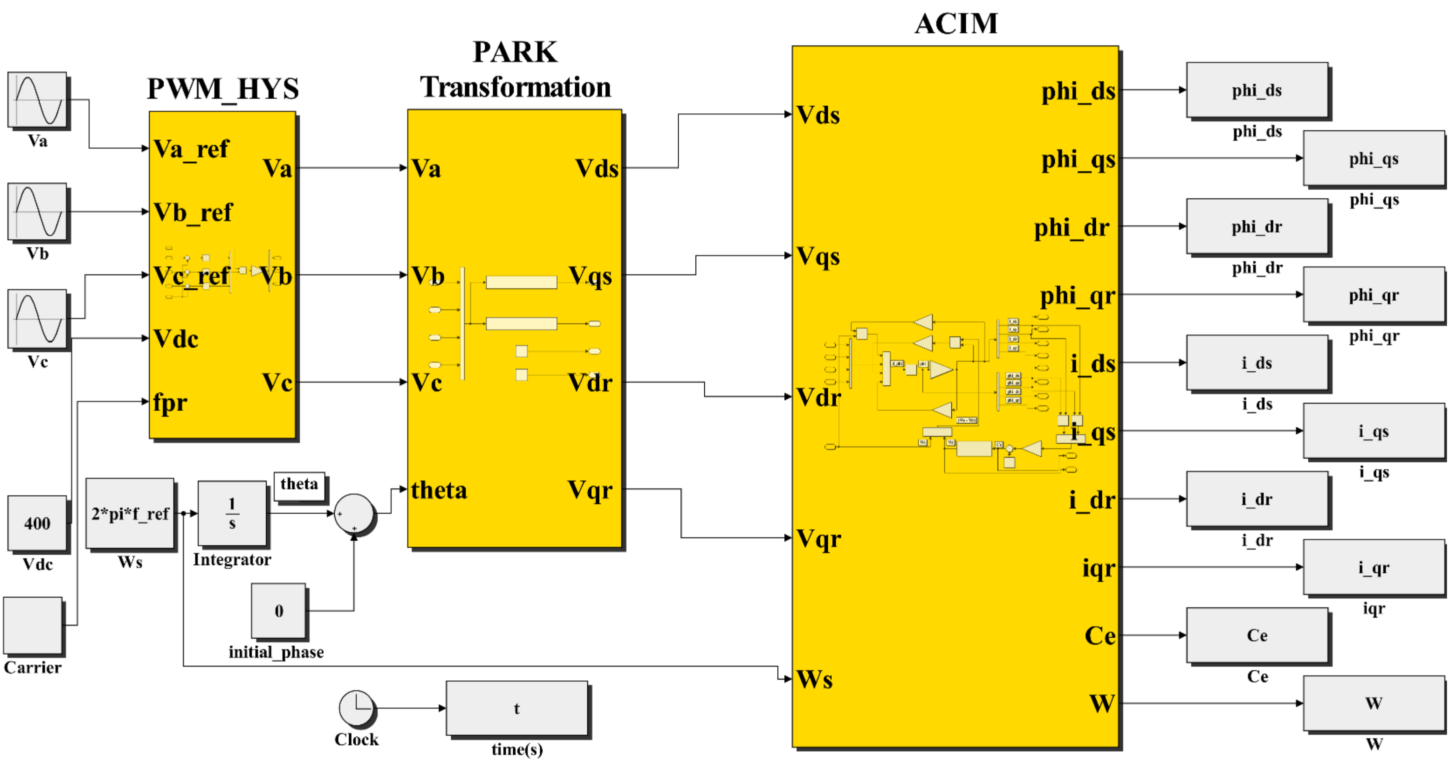


Fig I- 18 : IM simulation diagram powered by PWM hysteresis voltage inverter

The carrier frequency is $f_{carrier} = 5kHz$ with the load applied:

Fig (I-19) shows the similarity of the results obtained when the machine is powered by the hysteresis voltage inverter, it shows the evolution of the electromagnetic torque Ce, of the speed W, of the fluxes $\varphi dr, \varphi qr$ and of the currents ids, iqs. I notice:

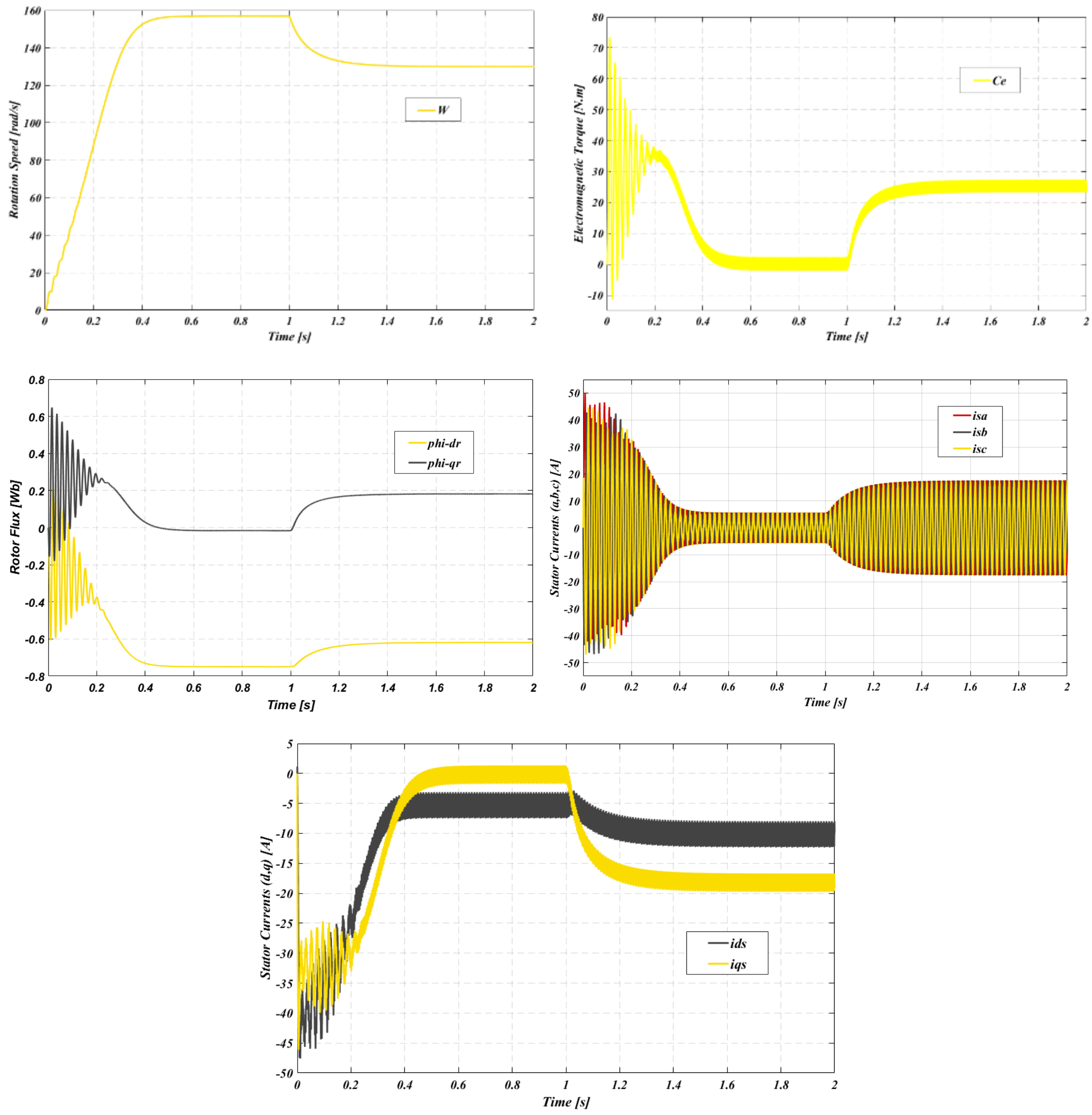


Fig I- 19 : Simulation results of the IM powered by Inverter with Hysteresis and with application of load $Cr = 25N.m$ at $t = 1s$

The torque oscillation is the highlight of this oscillator for a duration of $0.2\ s$, since the rated torque of the motor is much lower. It will therefore be necessary to take care with the dimensioning of the torque meter used if one does not want to destroy it. After disappearance of the transient state, the torque tends towards zero with many oscillations.

The torque oscillations are obviously felt on the evolution of the speed which in steady state stabilizes at 157 rad/s after 0.4 s with many oscillations.

I recognize the classic inrush current at startup equal to approximately 5 times the rated current (the current is measured in A). After its disappearance, the steady state is reached and there remains the.

According to this curve we see that the ripples of the current also affect the flux, also that the flux goes through a transient period then it reaches its final value following a sine wave

We applied a load of $25\ N \cdot m$ and a frequency of $5\ kHz$, the latter causes a decrease in speed and flux and an increase in the stator current and the torque keeps its permanent value with oscillations due to harmonic pollution.

The simulation results of the IM which is directly connected by the network are almost identical to the results of this part which is powered by a three-phase hysteresis inverter with small oscillations due to the switching frequency of the switches.

I. 11 - Conclusion:

In this chapter, I exposed the principle of operation as well as the modeling of the asynchronous machine. The development of the mathematical model of the machine is a necessary step for the simulation of their different operating and control systems. The model of the simulated machine was established by switching from the real three-phase system to a two-phase PARK system.

I also managed to model the frequency converter, which is represented by the voltage inverter, controlled by a PWM with hysteresis. The machine-converter association is simulated with voltage supply to highlight the impacts of this converter on the machine, with the use of different frequencies within the framework of the PWM with Hysteresis.

The results obtained clearly show the feasibility of the Park model used. The model of the machine as well as the inverter will be exploited for the establishment of the closed loop control which will be the subject of the second chapter (Rotor Flux Orientation Vector Control)

Chapter_II

“ Rotor Flux Orientation Vector Control”

II. 1 -Introduction:

The Field Orientation Control is denoted FOC.

The separately excited DC machine offers the main advantage of being easily controllable. The flux and the torque are decoupled and controlled independently and thanks to this property, high dynamic performance can be achieved. However, the presence of the brush-collector system limits its fields of use (power, speed). [35]

For several years, university and industrial research has been developed in order to achieve control of the asynchronous drive, equivalent to that of a DC motor. In this type of control, flux and torque, two essential adjustment variables, are decoupled and controlled independently. This principle of decoupled control, conditioning the stable operation of the asynchronous motor, is a characteristic principle of vector control. The latter leads to asynchronous drives with high industrial performance supporting the disturbances due to the load. [35] [37]

The absence of the brush-commutator system is one of the decisive advantages to replace the direct current machine by the alternating current one is precisely the asynchronous machine.

however, this machine has a difficulty in terms of control. That is to say that the torque and the flux are strongly coupled variables and that any action on one of them affects the other. [36]

The principles of this control were developed in 1972, by F. Blaschke, it brings the behavior of the asynchronous motor to that of a DC motor. It consists in placing the reference (d, q) such that the axis (d) coincides with the field to be oriented. [35] [36]

The aim is to eliminate the problem of coupling between the rotor and the stator by splitting the stator current into two components in quadrature, such that one of the components controls the flux and the other controls the torque. This makes it possible to reduce to operations comparable to those of a DC motor with separate excitation, where the field current controls the flux and the induced current controls the torque. [37]

This chapter consists in introducing the direct vector control method by orientation of the rotor flux. The methodology consists, first of all, in presenting the equations of the machine model, represented in the form of block diagrams, then in adding the command to this last formulation.

II. 2 -FOC historical review:

The first theoretical developments of the oriented field method were made in the early 70s by F. Blaschke and its effective applications were born thanks to Leonhard ten years later.

The publications of Leonhard (1983) show that the implantation of the CV offers to the IM supplied with current and voltage performances comparable to those of the direct current machine.

Recently, many developments and refinements have been introduced, thanks in large part to the hardware and software means allowing the perfect knowledge of the instantaneous position of the rotor flux. [38]

II. 3 -Advantages and disadvantages of vector control:

Vector control has the following advantages: [39]

- It is based on the transient model (to treat the transitory modes which the scalar command did not allow to do)
- It is precise and fast.
- There is torque control at standstill.
- The control of quantities is done in amplitude and in phase It has

Also, some disadvantages:

- Expensive (incremental encoder or speed estimator, DSP.). The processor must be able to calculate the algorithm approximately every millisecond.
- Weak robustness to parametric variations and in particular to those of the rotor time constant.
- Presence of coordinate transformations depending on an estimated angle θ_s.
- Wrong parameters lead to a torque error.

II. 4 -Field orientation control (FOC):

The examination of the expression of the torque of the asynchronous machine shows that it results from a difference of products of two components in quadrature, rotor fluxes and stator currents which presents a complex coupling between the quantities of the machine.

The aim of flux orientation control is to decouple the quantities responsible for the magnetization of the machine and the production of torque.

Mathematically, the control law consists in establishing all the transformations to pass from a system having a double structural non-linearity to a linear system which ensures independence between the creation of the field and the production of the torque as in a machine. with separately excited direct current. [39]

Flux orientation control consists in adjusting the flux by one component of the current and the torque by the other component. For this, it is necessary to choose a system of axis (d, q). A judicious choice of the orientation angle of the (d, q). mark leads to the alignment of the "d" axis on the resultant of the field, this alignment allows the cancellation of the transverse component of the field as indicated in Fig (II-1). [38]

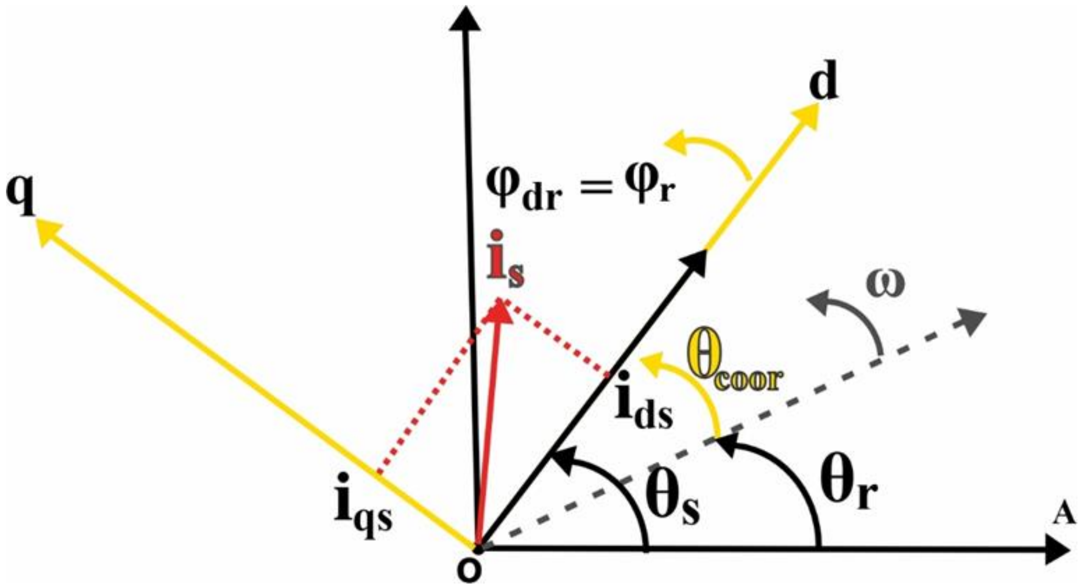


Fig II- 1 : Orientation of the rotor flux on the axis d

II. 5 -Choice of field orientation:

The choice of orientation axes can be made according to one of the machine field directions, namely the rotor, stator or air gap flow. [29]

- $\varphi_{rq} = 0 \; and \; \varphi_r = \varphi_{rd}$: it is the rotor flux which is oriented.
- $\varphi_{sd} = 0 \; and \; \varphi_s = \varphi_{sq}$: it is the stator flux which is oriented.
- $\varphi_{mq} = 0 \; and \; \varphi_m = \varphi_{md}$: it is the air gap flow that is oriented.

In the three cases the torque is proportional to the product of the flux by the component of the stator current in quadrature with the flux.

The orientation of the rotor flux makes it possible to obtain a high starting torque and requires knowledge of the rotor parameters.

In all that follows, the orientation of the rotor flux is the method that will be retained.

II. 6 -Principle of FOC by orientation of the rotor flux:

In this case the rotor flux is oriented on the axis " d " of a reference linked to the rotating field of speed (ω_s), so we can notice the following properties: [40] [41]

The transverse component of the rotor flux is zero. ($\varphi_{rq} = 0$)

The " d " axis is systematically aligned with the rotor flux vector. ($\varphi_r = \varphi_{rd}$)

The longitudinal component of the rotor current is zero if the rotor flux is kept constant. ($\varphi_{rq} = const \Leftrightarrow i_{rd} = 0$)

The vector model of the asynchronous machine is described by the following equations:

$$\begin{cases} \overline{V_s} = R_s\overline{\iota_s} + \frac{d}{dt}\overline{\varphi_s} + j\omega_s\overline{\varphi_s} \\ 0 = R_r\overline{\iota_r} + \frac{d}{dt}\overline{\varphi_r} + j\omega_r\overline{\varphi_r} \end{cases} \text{------(II- 1)}$$

$$\begin{cases} \overline{\varphi_s} = L_s\overline{\iota_s} + M\overline{\iota_r} \\ \overline{\varphi_r} = L_r\overline{\iota_r} + M\overline{\iota_s} \end{cases} \Rightarrow \overline{\iota_r} = \frac{\overline{\varphi_r}}{L_r} - \frac{M\overline{\iota_s}}{L_r} \text{------(II- 2)}$$

With:

$\bar{x} = x_d + jx_q$ ($\bar{x}$: represents flux, currents and voltages)

To write the IM model with the state ($\overline{\iota_s}, \overline{\varphi_r}$) we make the following changes:

$$\overline{\varphi_s} = L_s\overline{\iota_s} + \frac{M}{L_r}\overline{\varphi_r} - \frac{M^2}{L_r}\overline{\iota_s} = L_s\left(1 - \frac{M^2}{L_sL_r}\right)\overline{\iota_s} + \frac{M}{L_r}\overline{\varphi_r} \text{------(II- 3)}$$

$$\Longrightarrow \quad \overline{\varphi_s} = \sigma L_s\overline{\iota_s} + \frac{M}{L_r}\overline{\varphi_r} \text{------(II- 4)}$$

Equation (II-1) in the stator voltage equation and the rotor voltage equation gives:

$$\overline{\iota_r} = \frac{\overline{\varphi_r}}{L_r} - \frac{M\overline{\iota_s}}{L_r} \text{------(II- 5)}$$

$$\begin{cases} \overline{V_s} = R_s\overline{\iota_s} + \frac{d}{dt}\left(\sigma L_s\overline{\iota_s} + \frac{M}{L_r}\overline{\varphi_r}\right) + j\omega_s\left(\sigma L_s\overline{\iota_s} + \frac{M}{L_r}\overline{\varphi_r}\right) \\ 0 = R_r\left(\frac{\overline{\varphi_r}}{L_r} - \frac{M\overline{\iota_s}}{L_r}\right) + \frac{d}{dt}\overline{\varphi_r} + j\omega_r\overline{\varphi_r} \end{cases} \text{------(II- 6)}$$

$$0 = \frac{-M\,R_r}{L_r}\,\overline{\iota_s}\left(\frac{R_r}{L_r} + j\omega_r\right)\cdot\overline{\varphi_r} + \frac{d}{dt}\overline{\varphi_r} \implies 0 = \frac{-M\,R_r}{L_r}\,\overline{\iota_s}\left(\frac{R_r}{L_r} + j\omega_r + S\right)\cdot\overline{\varphi_r} \text{--(II- 7)}$$

$\overline{\varphi_r} = \varphi_{dr}$: (the orientation of the rotor flux)

We decompose the equation into real and imaginary parts, we will have:

$$\begin{cases} 0 = \frac{-M\,R_r}{L_r} i_{sd} + \left(\frac{R_r}{L_r} + S\right)\varphi_r \\ 0 = \frac{-M\,R_r}{L_r} i_{sq} + \omega_r \cdot \varphi_r \quad \Longrightarrow \quad i_{sq} = \frac{\omega_r \cdot L_r}{R_r} \end{cases} \text{---------------------------------(II- 8)}$$

$$\begin{cases} 0 = \frac{-M\,R_r}{L_r} i_{sd} + \left(\frac{R_r}{L_r} + S\right)\varphi_r \\ 0 = \frac{-M\,R_r}{L_r} i_{sq} + \omega_r \cdot \varphi_r \quad \Longrightarrow \quad i_{sq} = \frac{\omega_r \cdot L_r}{R_r} \end{cases} \text{---------------------------------(II- 9)}$$

The relation of i_{sd} means that in the model of the asynchronous machine in the frame (d, q) with oriented rotor flux, the modulus of this flux is linearly controlled by the direct component of the stator current i_{sd} with a first order dynamics with the constant of time $T_r + j\varphi_r\overline{\varphi_r}$

$$\begin{cases} V_{rd} = R_r i_{rd} + \frac{d}{dt}\varphi_{rd} - \omega_r \varphi_{rq} = 0 \\ V_{rq} = R_r i_{rq} + \frac{d}{dt}\varphi_{rq} + \omega_r \varphi_{rd} \end{cases} \text{--- } (\nabla)$$

$$(\nabla) \Longrightarrow \quad 0 = R_r i_{rd} + \frac{d}{dt}\varphi_{rd} \text{--- (II- 10)}$$

$$\varphi_{rd} = const \Longrightarrow \frac{d}{dt}\varphi_{rd} = 0 \text{--- (II- 11)}$$

From these properties we can write:

$$\begin{cases} \varphi_{rq} = 0 \\ \varphi_{rd} = \varphi_r = const \\ i_{rd} = 0 \end{cases} \text{-- (II- 12)}$$

We replace this system in the flux equations, we get:

$$\begin{cases} \varphi_{rd} = M i_{sq} \\ \quad then \\ \varphi_{rq} = L_r i_{rq} + M i_{sq} = 0 \end{cases} \text{-- (II- 13)}$$

From the last equation of this system we will have the following equation:

$$i_{rq} = -\frac{M}{L_r} i_{sq} \text{--- (II- 14)}$$

We replace the expression (II-12) in the electromagnetic torque equation we will have:

$$C_e = \frac{pM}{L_r} \cdot \varphi_{rd} \cdot \varphi_{sq} \text{-- (II- 15)}$$

Also, the torque becomes:

$$C_e = K \cdot i_{sd} \cdot i_{sq} \text{--- (II- 16)}$$

With: $$K = \frac{pM^2}{L_r} \text{--- (II- 17)}$$

This expression is analogous to that of the torque of a DC machine. The Fig (II-2) illustrates the equivalence between the expression of the torque that is achieved with the conventional uncoupled control of a DC machine and the vector control of an asynchronous machine. Thus, the brush-collector system in the DC machine is replaced, in the case of the asynchronous machine, by the autopilot system which makes it possible to achieve harmony between the frequency of rotation and that of the currents induced in the rotor, such that the following relationship: [26]

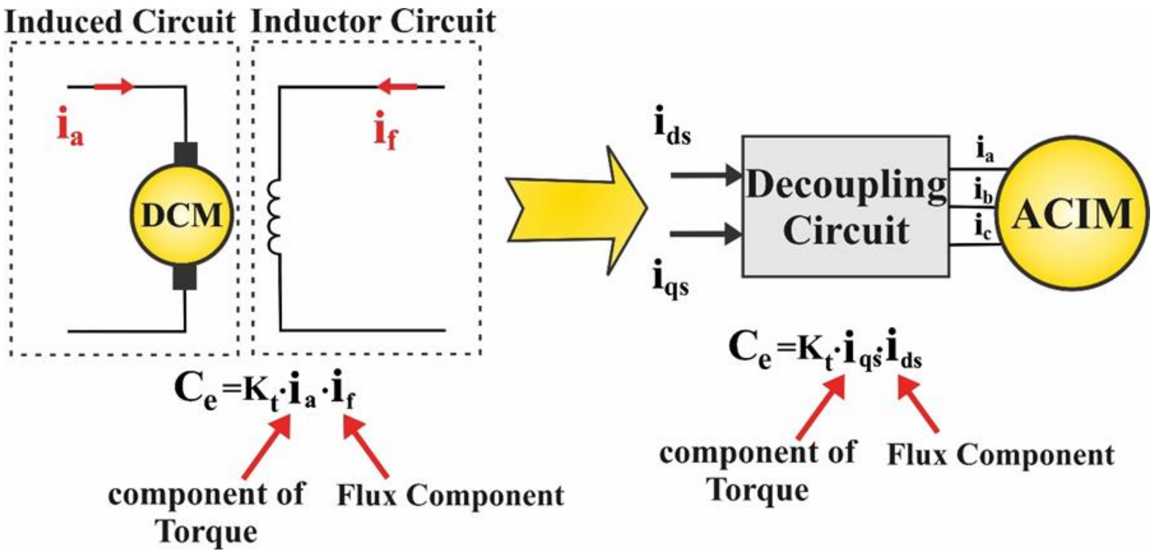


Fig II- 2 : Equivalence between the control of a separate-field motor and the vector control of an IM

II. 7 -Types of field orientation control:

All the research work carried out on this subject uses two main methods, the first called the direct method which was developed by F. Blaschke, the second known by the indirect method developed by K. Hasse.[42]

II.7. 1 - Direct vector control:

This method requires a good knowledge of the flux modulus and its position, and this must be verified regardless of the transient state performed. It is therefore necessary to carry out a series of measurements at the terminals of the system. Direct measurement makes it possible to know exactly the position of the flux. This control mode guarantees correct decoupling between flux and torque, whatever the operating point. However, it requires the use of a flux sensor, which considerably increases the cost of its manufacture and makes its use weaker. [42] [43]

The application of this method imposes several disadvantages of different natures:

- The unreliability of the flux measurement:
- measured signal filtering problem.

> mediocre accuracy of the measurement which varies according to the temperature (heating of the machine) and the saturation.

- High production cost (filter sensors).

Conventional direct field-oriented control (DFOC) algorithms provide more precision for torque control than scalar schemes, but require sensors for the speed control of the rotor and the magnetic flux to provide the data for the FOC algorithms. They also face challenges in the dynamic response and the dependence on measuring the parameters in the motor. [44]

We apply the direct vector control to the asynchronous machine supplied with voltage with converter.

II.7. 2 - Indirect vector control:

This method does not use the amplitude of the rotor flux but only its position. It does not require the use of a rotor flux sensor but requires the use of a sensor or a rotor position (speed) estimator. The latter can be developed by two main groups: [42] [43]

The rotor flux vector is obtained indirectly from the measured stator currents and voltages. In the second group, the rotor flux vector is estimated from the measurement of the stator currents and the rotor speed, based on the equations of the rotor circuit of the asynchronous motor in a reference system rotating in synchronism with the vector of rotor flux.

The major drawback of this method is the sensitivity of the estimate to the variation of the machine parameters due to the magnetic saturation and the temperature variation, especially the rotor time constant T_r .

indirect field-oriented control (IFOC) method estimates the phase angle of the rotor magnetic field flux, eliminating the need for additional sensors but adding to the complexity and the computation time of the control system. [44]

II. 8 -Structure of FOC of the rotor flux of IM supplied with voltage:

In this type of power supply, the control becomes more complicated because one must consider the dynamics of the stator in addition to that of the rotor. [42] The control quantities are the stator voltages (V_{sd}, V_{sq}) and the speed of the rotating field (ω_s). Considering the two stator voltages as control variables, the two stator currents $(i_{sd},\ i_{sq})$, the rotor flux (φ_r) and the mechanical speed (Ω_r) as state variables, we obtain the model of the three-phase asynchronous machine supplied with voltage by orientation of the rotor flux, [40][45]Taking into account that: $\varphi_{rq} = 0$ and $\overline{\varphi_{rd}} = 0 \longrightarrow \varphi_{rd} = \varphi_r$

$$\begin{cases} V_{sd} = \sigma L_s \frac{d}{dt} i_{sd} + \left(R_s + R_r \frac{M^2}{L_r{}^2}\right) \cdot i_{sd} - \omega_s \sigma L_s \, i_{sq} - \frac{M}{L_r{}^2} R_s \varphi_r \\ V_{sq} = \sigma L_s \frac{d}{dt} i_{sq} + \left(R_s + R_r \frac{M^2}{L_r{}^2}\right) \cdot i_{sq} - \omega_s \sigma L_s \, i_{sd} - \frac{M}{L_r{}^2} p \Omega_r \varphi_r \\ T_r \cdot \frac{d}{dt} \cdot \varphi_r + \varphi_r = M \cdot i_{sd} \\ \omega_s = p \cdot \Omega_r + \frac{M \cdot i_{sq}}{T_r \cdot \varphi_r} \end{cases} \quad \text{------ (II- 18)}$$

These expressions can be exploited as such to achieve flux-oriented vector control of asynchronous motors supplied with voltage but and influence both and therefore the flux and the torque Fig (II-3).

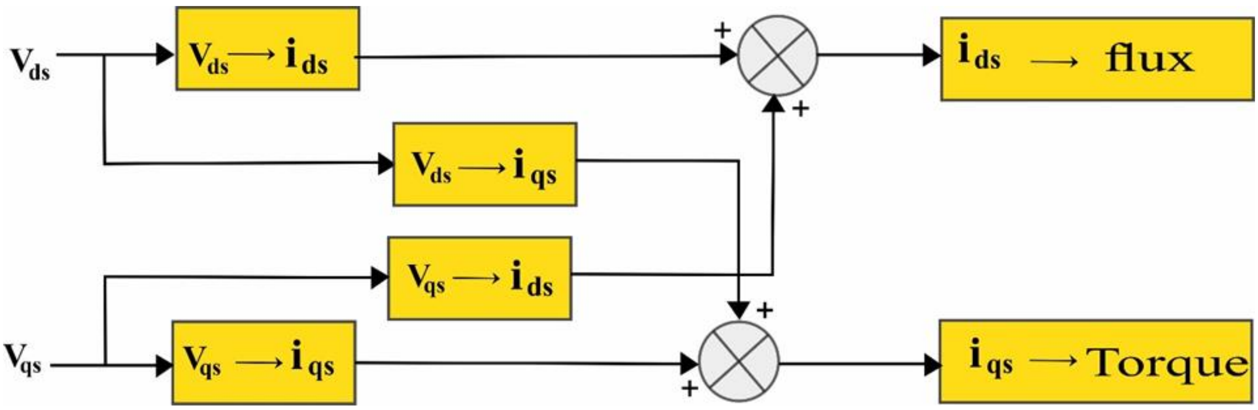


Fig II- 3 : The concept of decoupling.

II.8. 1 - System of equations related to rotor flux:

The equations (II-19), provided with the constraint ($\varphi_r = 0$) are simplified:

$$\begin{cases} V_{sd} = \sigma L_s \frac{d}{dt} i_{sd} + R_s i_{sd} - \omega_s \sigma L_s \, i_{sq} + \frac{M}{L_r} \frac{d}{dt} \varphi_r \\ V_{sq} = \sigma L_s \frac{d}{dt} i_{sq} + R_s i_{sq} + \omega_s \sigma L_s \, i_{sd} + \omega_s \frac{M}{L_r} \varphi_r \end{cases} \quad \text{------------------------ (II- 19)}$$

$$\frac{d}{dt} \varphi_r = \frac{M}{T_r} i_{sd} - \frac{1}{T_r} \varphi_r \text{-- (II- 20)}$$

$$C_e = \frac{P \cdot M}{L_r} \varphi_r i_{sq} \text{ -- (II- 21)}$$

$$\omega_r = \frac{M}{T_r} \cdot \frac{i_{sq}}{\varphi_r} \text{-- (II- 22)}$$

$$J \cdot \frac{d}{dt} \Omega_r = C_e - C_r - f \cdot \Omega_r \text{-- (II- 23)}$$

The equations (II-19) to (II-23) highlighting respectively the current producing the flux, and the current producing the torque. This offers the possibility of controlling the asynchronous machine by decoupling as in the direct current machine, the flux and the torque. The decoupling structure is defined by the equations defined previously (II-19) and (II-20). The block diagram of this structure is represented by the assembly of Fig (III-4). [45]

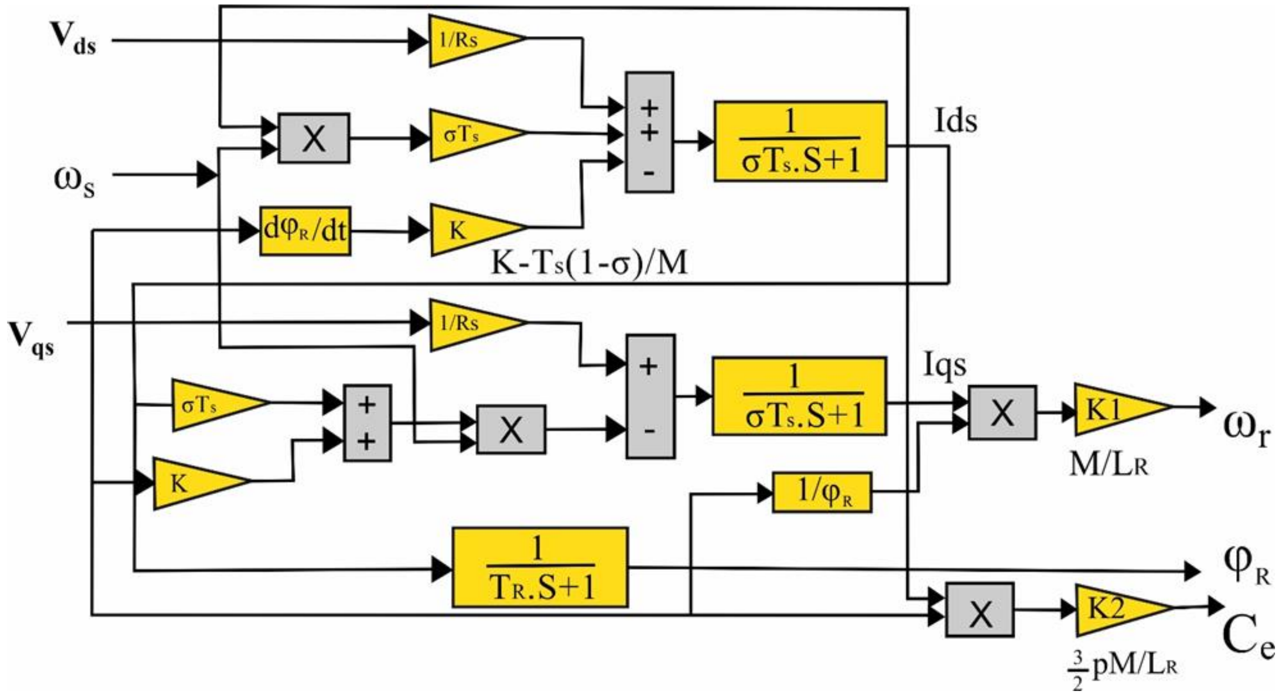


Fig II- 4 : Structure of the rotor flux orientation control

II.8. 2 - Input-Output decoupling:

The vector control laws of asynchronous machines supplied with voltage present couplings between the actions on the (d, q) axes. The flux and the torque depend simultaneously on the voltages (V_{sd}, V_{sq}), therefore a decoupling must be carried out.

The objective is, as far as possible, to limit the effect of an input to a single output, we can then model the process in the form of a set of single-variable systems evolving in parallel. The commands are then not interactive. Different techniques exist: decoupling using a regulator, decoupling by state feedback, decoupling by compensation, we present decoupling by compensation. [26] [46]

II.8. 3 - Decoupling by compensation method:

Definitions of two new control variables V_{sd}^* And V_{sq}^* such as: [46]

$$V_{sd} = V_{sd}^* emf_d \qquad and \qquad V_{sq} = V_{sq}^* emf_q$$

With:

$$\begin{cases} emf_d = \omega_s \sigma L_s i_{sq} + \frac{M}{L_r} R_r \varphi_r \\ emf_q = -\left(\omega_s \sigma L_s i_{sd} + \frac{M}{L_r} R_r \varphi_r\right) \end{cases} \text{--- (II- 24)}$$

"emf_d" and "emf_q" disturbances partially related to currents and introducing nonlinear coupling.

The voltages (V_{sd} , V_{sq}) are then reconstituted from the voltages $\left(V_{sd}{}^*, V_{sq}{}^* \right)$ Fig (II-5)

Such as:

$$\begin{cases} V_{sd}{}^* = \sigma L_s \frac{d}{dt} i_{sd} + \left(R_s + \frac{M^2}{L_r{}^2} R_r\right) \cdot i_{sd} \\ V_{sd}{}^* = \sigma L_s \frac{d}{dt} i_{sq} + \left(R_s + \frac{M^2}{L_r{}^2} R_r\right) \cdot i_{sq} \end{cases} \quad \text{------------------------------------- (II- 25)}$$

The block diagram in the following figure shows the compensation method for cross terms and nonlinear terms.

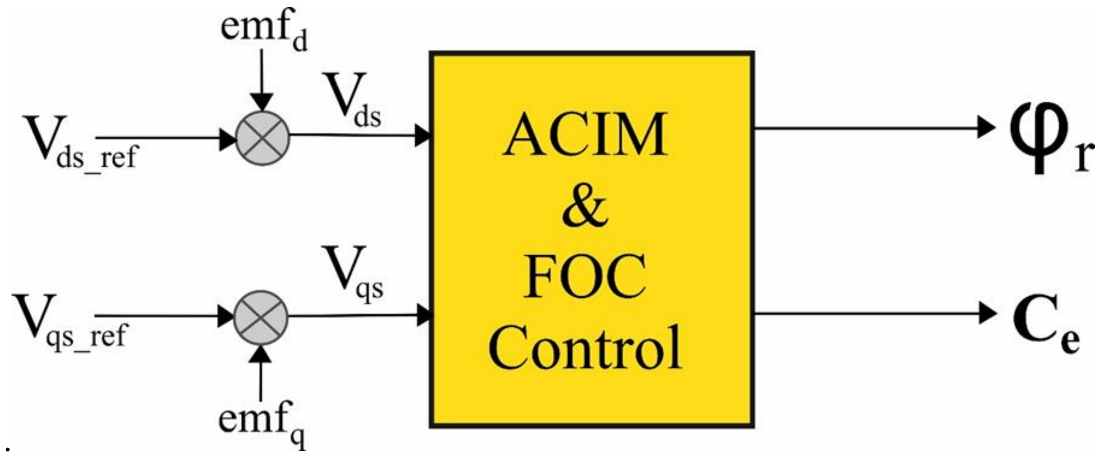


Fig II- 5 : Reconstruction of voltages V_{sd} and V_{sq}

II.8. 4 - Field weakening block:

The field weakening allows optimal exploitation of the magnetic capacities of the machine, allows operation at constant torque if the speed is lower than the rated speed on the one hand this block also makes it possible to weaken the flux inversely proportional to the speed, for operation at constant power when the speed exceeds the rated speed. It is defined by the following nonlinear function: [47]

Under-speed: $\varphi_r = \varphi_{r\_rated} \xrightarrow{for} |\Omega_r| \leq \Omega_{r\_rated}$ ------------------------------- (II- 26)

Over-speed: $\varphi_r = \frac{\Omega_{r\_rated}}{|\Omega_r|} \cdot \varphi_{r\_rated} \xrightarrow{for} |\Omega_r| \geq \Omega_{r\_rated}$ --------------------- (II- 27)

With:

Ω_{r_rated} : Rated rotational speed.

φ_{r_rated} : rated rotor flux.

The principle of flux-weakening consists of keeping the rotor flux constant and equal to the rated flux and by varying it over a range for speeds above the rated speed. [40]

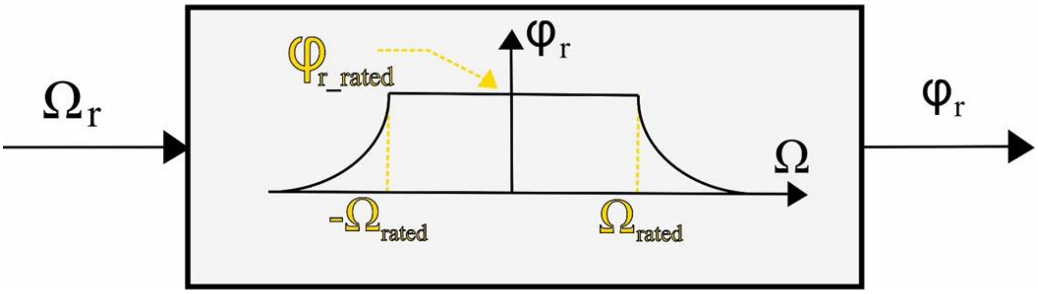


Fig II- 6 : Field weakening block.

II.8. 5 - Principle of operation:

The block diagram of the structure of direct vector control by orientation of the rotor flux of an IM supplied with voltage is represented by the Fig (II-7). [26]

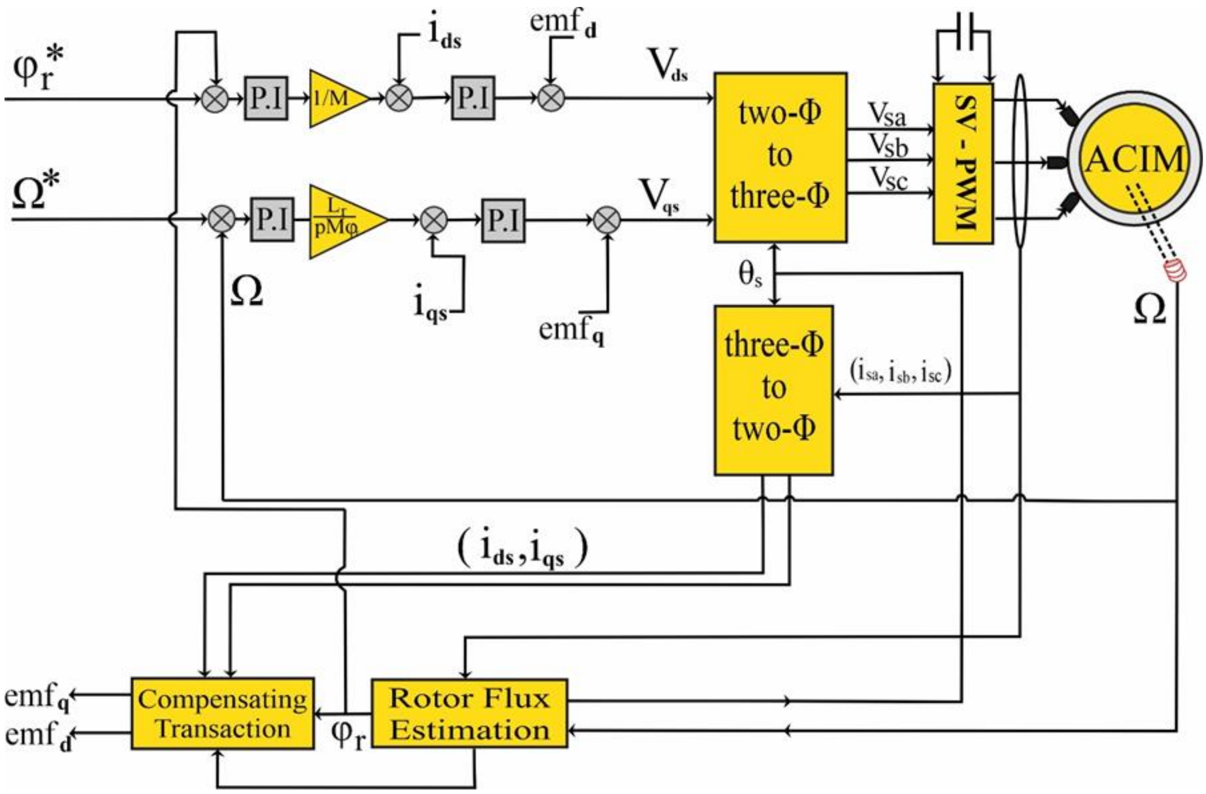


Fig II- 7 : Schematic diagram of direct vector control with oriented rotor flux of the IM

II.8. 6 - Rotor flux estimation:

Only the stator quantities are accessible, the rotor quantities are not, it is therefore necessary to be able to estimate them from the stator quantities. [48]

In general, rotor flux estimators come in four forms:

- estimator based on a current model.
- estimator based on an elimination method.

- estimator based on a voltage model.
- full order estimator.

All these estimators come from equations modeling the Asynchronous Machine. From the system of equations (II-19) to (II-13) with " $\omega_{coor} = 0$ " (Stator-related frame of reference), we deduce that the rotor fluxes " $\varphi_{rd,} \varphi_{rq}$ " can be estimated from the stator currents and the speed of the rotor. This open-loop estimator, which is presented in Fig (II-8), is based on a so-called current model. [48]

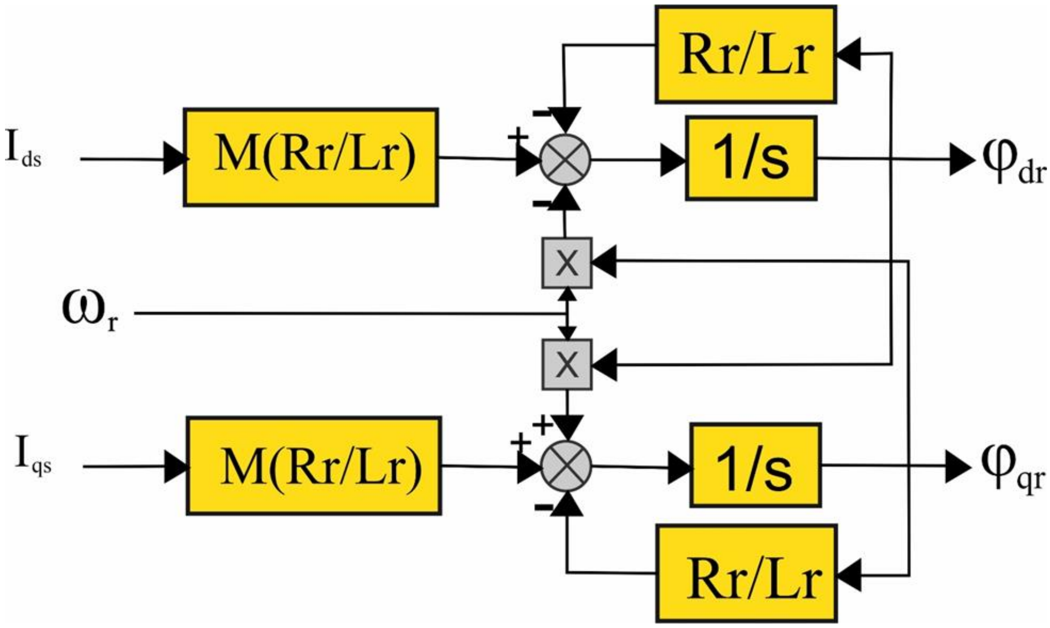


Fig II- 8 : Diagram of the open-loop flux estimator

The flux modulus is given by:

$$\sqrt{\varphi_{rd}{}^2 + \varphi_{rq}{}^2} \xrightarrow{and} \theta_s = \tan^{-1}\left(\frac{\varphi_{rq}}{\varphi_{rd}}\right) \text{ ---------------- (II- 28)}$$

II.8. 7 - Regulation of the system:

In the case of my study, we limit ourselves to the control technique (PI).

II.8.7. 1 - Design of regulators:

$y(t)^*$ is the signal to follow, and $y(t)$ the output signal of the system to control. [49]

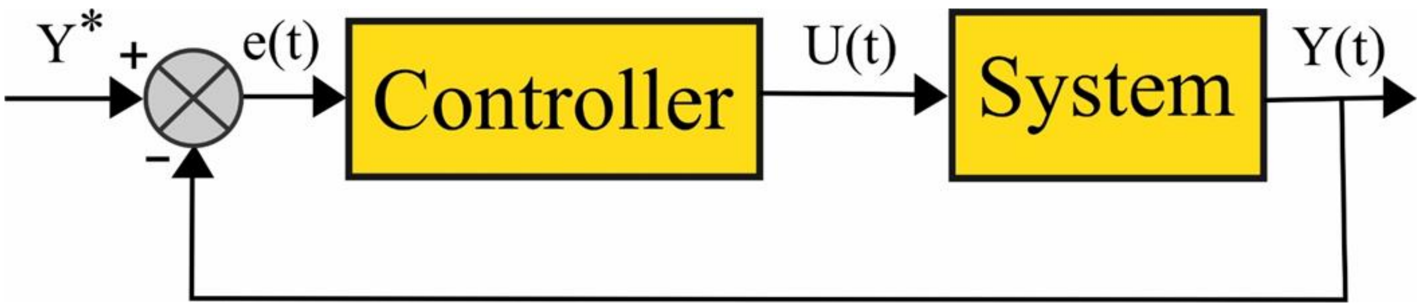


Fig II- 9 : Representation of PI control

The control law is:

$$U(t) = K_p \cdot e(t) + K_i \int e(t)\, dt \text{ -- (II- 29)}$$

II.8.7.1. a - Proportional action:

If $\boldsymbol{K_p}$ is great, the correction is fast. The risk of overshoot and oscillation in the output increases.

If $\boldsymbol{K_i}$ is small, the correction is slow, there is less risk of oscillations. [50]

II.8.7.1. b - Integral action:

The integral action reacts slowly to the variation of the error and ensures progressive catching up of the setpoint.

As long as the positive (or negative) error remains the action $U\ (t)$ increases (or decreases) until the error cancels out. [50]

II.8.8 - Regulators characteristics: [50] [51]

II.8.8.1 - Stability:

A looped system must be stable. If only if the reactions of the regulation system are energetic without being disproportionate with the error to be corrected.

A correction that is too strong or too late risks leading the system to instability.

II.8.8.2 - Accuracy:

In regulation, the accuracy obtained by the implementation of integration in the loop.

II.8.8.3 - Rapidity:

In general, a looped system must respond quickly to the variation of its setpoint (tracking) and quickly clear disturbances (regulation).

The reaction time is of course closely related to the inertia of the process.

II.9- Calculation of regulators: [26] [40] [41] [51]

II.9.1 - Direct stator current i_{sd} regulation:

The block diagram of the regulation of the direct component of the stator current " i_{sd}" is shown in the Fig (II-10):

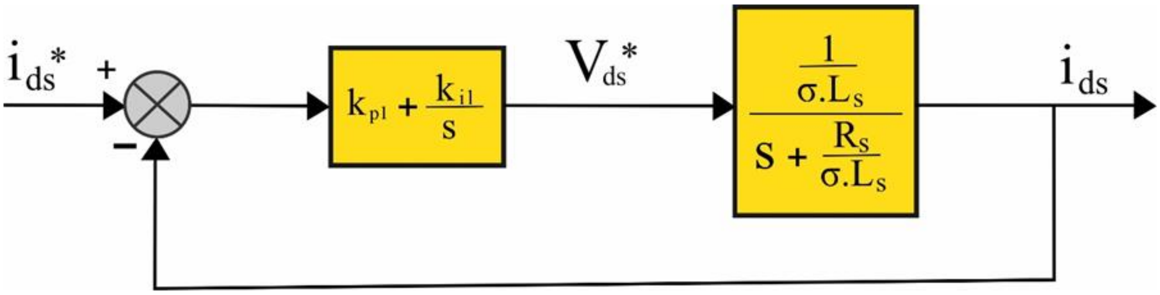


Fig II- 10 : Block diagram of " i_{sd} " stator current regulation

The open loop transfer function is written:

$$Fi_{sd}(s) = \frac{i_{sd}}{i_{sd}^*} = K_{p1} \cdot \frac{s + (K_{i1}/K_{p1})}{s} \cdot \frac{(1/\sigma L_s)}{s + (R_s/L_s)} \text{ ------------------------------- (II- 30)}$$

We compensate the pole " $s + (K_{i1}/K_{p1})$ " by $s + (\alpha/\sigma L_s)$ which results in the condition

$$\frac{K_{i1}}{K_{p1}} = \frac{R_s}{\sigma L_s} \text{ ------------------------------- (II- 31)}$$

The open loop transfer function is now written:

$$Fi_{sd}(s) = \frac{K_{p1}}{s \cdot \sigma L_s} \text{ ------------------------------- (II- 32)}$$

The closed loop transfer function is given by:)

$$Gi_{sd}(s) = \frac{Fi_{sd}}{1 + Fi_{sd}} = \frac{1}{1 + (\sigma L_s / K_{p1})} \Longrightarrow \tau_1 = \frac{\sigma L_s}{K_{p1}} \text{ ------------------------------- (II- 33)}$$

We obtain a response of the first order type of time constant τ_1 .

For an imposed response time of “**5**%”, we obtain the following condition:

$$3 \cdot \tau_1 = t_{resp1(5\%)} \text{ ------------------------------- (II- 34)}$$

$$3 \cdot \frac{\sigma L_s}{K_{p1}} = t_{resp1(5\%)} \text{ ------------------------------- (II- 35)}$$

Thus:

$$K_{p1} = \frac{3 \cdot \sigma L_s}{t_{resp1(5\%)}} \text{ ------------------------------- (II- 36)}$$

According to the equation (II-35), we can write:

$$K_{i1} = \frac{3 \cdot R_s}{t_{resp1(5\%)}} \text{ ------------------------------- (II- 37)}$$

II.9.2 - Quadrature stator current i_{sq} regulation:

The block diagram of the regulation of the inverse component of the stator current is shown in the Fig (II-11):

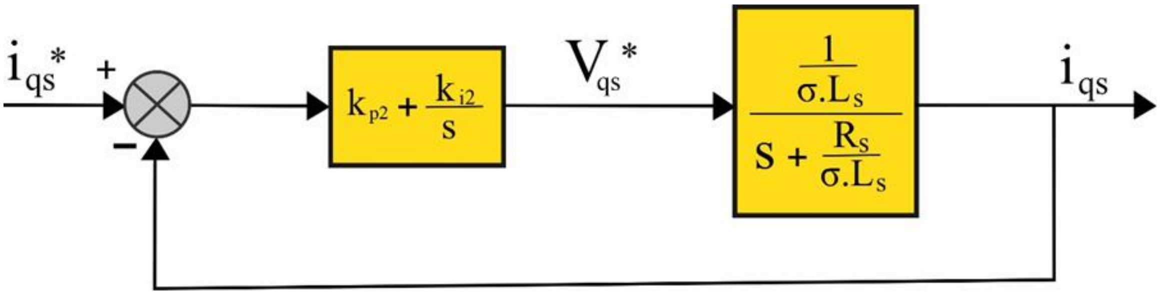


Fig II- 11 : Block diagram of i_{sq} stator current regulation

We notice that the current i_{sq} has the same dynamics as the current i_{sd} , we then find the same parameters as before

$$K_{p2} = \frac{3 \cdot \sigma L_s}{t_{resp2(5\%)}} \quad \text{(II- 38)}$$

$$K_{i2} = \frac{3 \cdot R_s}{t_{resp2(5\%)}} \quad \text{(II- 39)}$$

II.9.3 - Velocity regulation Ω_r :

The block diagram of velocity regulation is shown in the Fig (II-12):

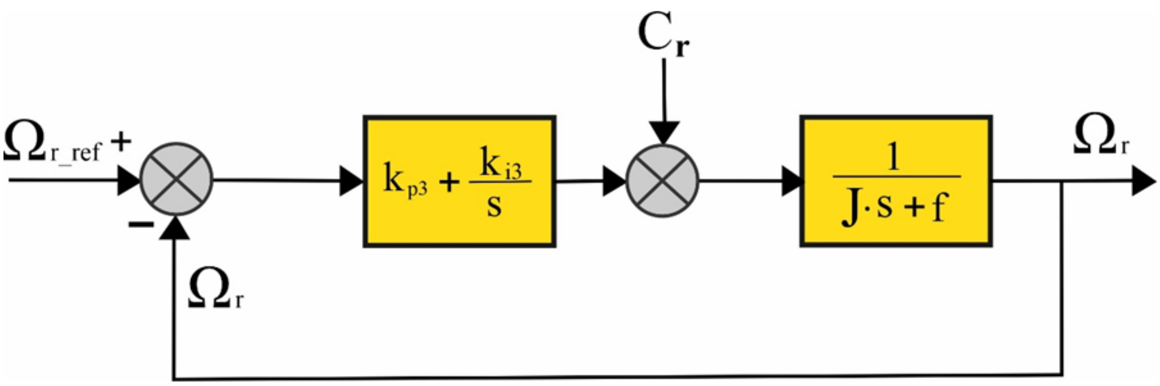


Fig II- 12 : Diagram of speed regulation block

The open loop transfer function with no-resistive torque is given by:

$$F\Omega_r(s) = \frac{\Omega_r}{\Omega_r^*} = \frac{s \cdot K_{p3} + K_{i3}}{s \cdot (s \cdot J + f)} \quad \text{(II- 40)}$$

The closed loop transfer function is written:

$$G\Omega_r(s) = \frac{s \cdot K_{p3} + K_{i3}}{J \cdot s^2 + (K_{p3} + f) \cdot s + K_{i3}} \quad \text{(II- 41)}$$

This transfer function has a second order dynamics, whose canonical form:

$$G(s) = \frac{\omega_n}{s^2 + 2\xi\omega_n \cdot s + \omega_n^2} \quad \Longrightarrow \quad G(s) = \frac{1}{\frac{1}{\omega_n^2} s^2 + \frac{2\xi}{\omega_n} \cdot s + 1} \quad \text{(II- 42)}$$

By comparison, we then get:

$$\frac{J}{K_{i3}} = \frac{1}{\omega_n^2} \quad and \quad \frac{K_{p3} + f}{K_{i3}} = \frac{2\xi}{\omega_n} \quad \text{(II- 43)}$$

For a damping coefficient $\xi = 1$ and a given pulsation ω_n , we obtain:

$$K_{i3} = J\omega_n^2 \quad \text{(II- 44)}$$

$$K_{p3} = 2J\omega_n - f \quad \text{(II- 45)}$$

II.9.4 - Flux regulation φ_r:

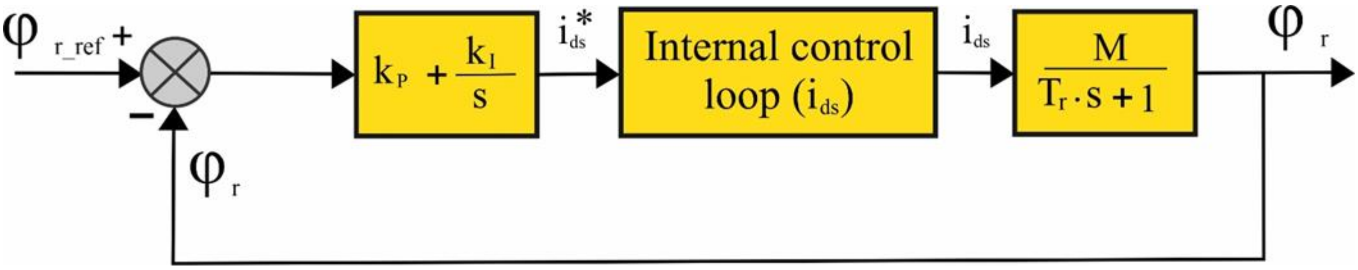


Fig II- 13 : Scheme of flux regulation block φ_r

For the outer loop of the flux, the calculation will be done in the same way. Knowing that the faster inner loop has already reached the desired reference value, the open loop transfer function is, according to the Fig (II-13), given by:

$$F\varphi_r(s) = K_{pf} \cdot \frac{s + (K_{if}/K_{pf})}{s} \cdot \frac{(M/T_r)}{s + (1/T_r)} \quad \text{(II- 46)}$$

After compensation of the dominance $\frac{1}{T_r} = \frac{K_{if}}{K_{pf}}$, the closed-loop response time will be $\frac{T_r}{K_{pf} \cdot M}$, and this will be chosen ten times greater than the response time of the internal loop to allow the internal current loop to reach the reference value $\tau_f = 10 \cdot \tau$.

$$K_{pf} = \frac{T_r}{10 \cdot \tau \cdot M} \quad \text{(II- 47)}$$

$$K_{if} = \frac{K_{pf}}{T_r} \quad \text{(II- 48)}$$

II. 10 - Simulation and interpretation:

From the theoretical study of the structure of the direct field-oriented vector control, in this chapter, we can elaborate the different blocks necessary for a simulation of the process. The overall diagram is given by Fig (II-14)

The simulation results of the machine control assembly are defined by the imposition of the following reference variables: $\varphi ref\ = 1Wb, \Omega ref\ = 157(rad/sec)$.

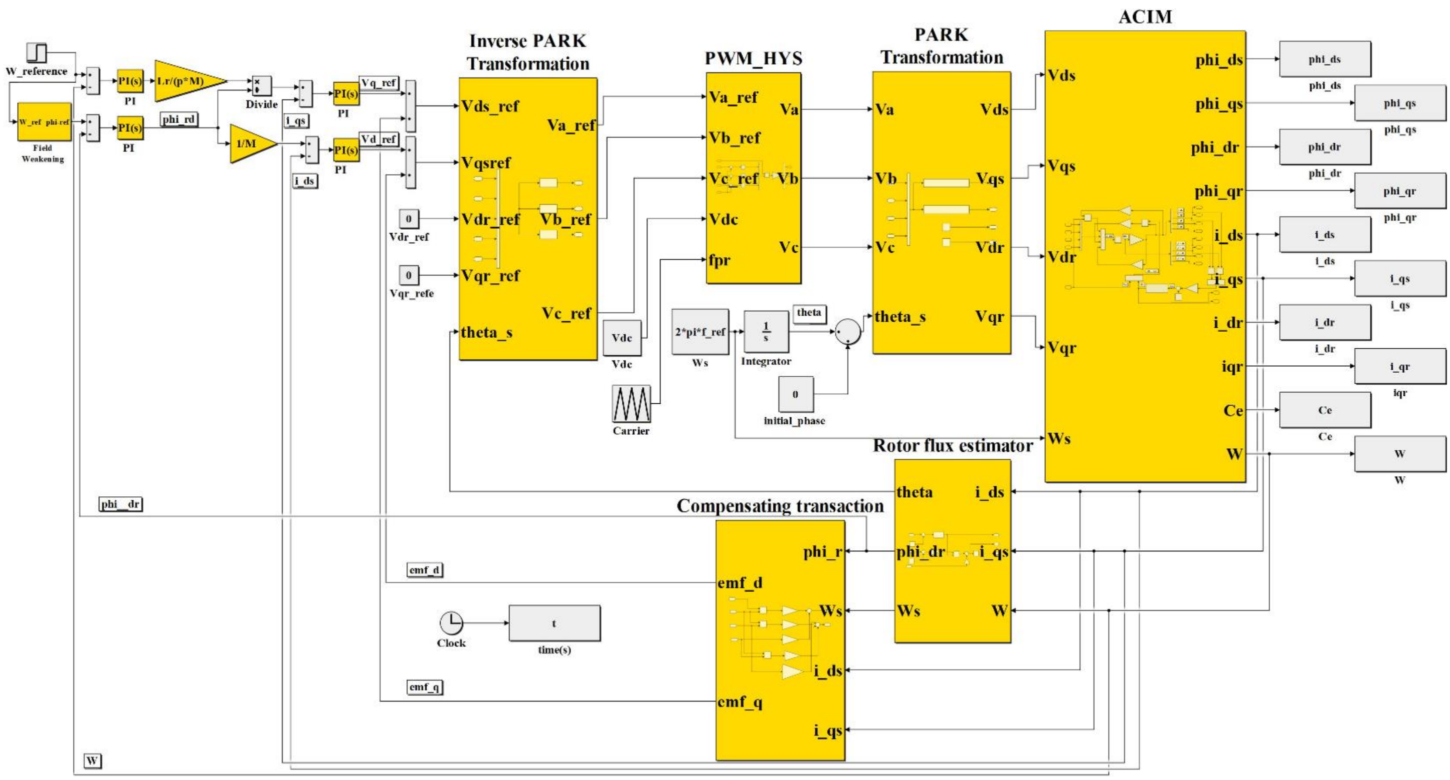


Fig II- 14 : Block diagram of direct vector control

Using the diagram of the general structure of the direct vector control Fig. (II-14), the tests are carried out from the simulation of the following operating modes:

In order to show the results obtained by simulation of direct voltage vector control. We simulated the system in different operating cases such as the speed, load variation and reversal of the direction of rotation and the parametric variation in this case the rotor resistance.

II.10 .1 - Simulation of the speed variation:

The simulation results obtained for the speed variation with application of a resistive torque $Cr\ = 25\ N.m$ of Fig (II-15) are explained in the following table:

$Time\ (sec)$	0	1	1.4
$\Omega_{ref}\ (rad/sec)$	157	90	157

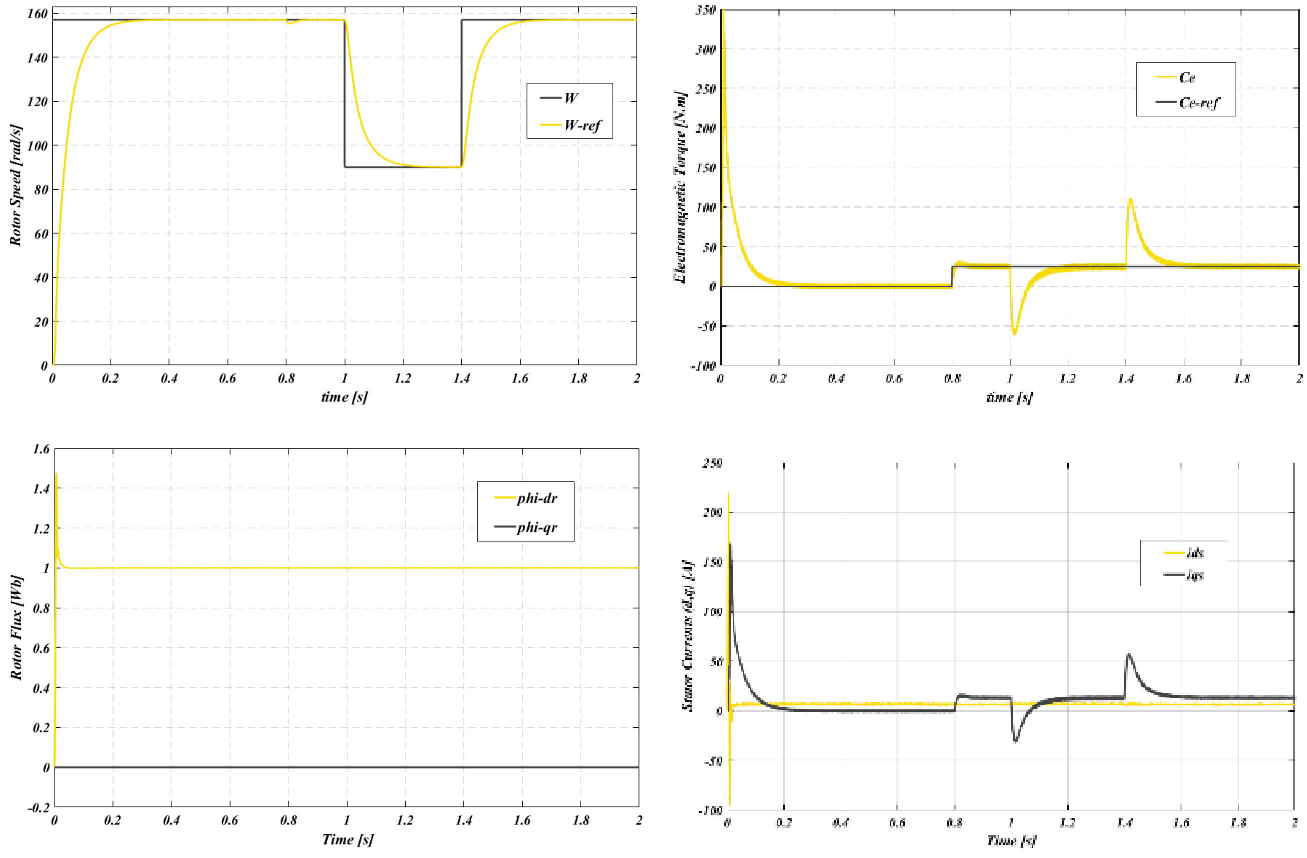


Fig II- 15 : DFOC simulation results during IM shaft speed variation

The simulation results obtained for the speed variation of Fig (II-15) show that this variation leads to a variation of the stator frequency which influences the currents, the fluxes, and the electromagnetic torque. It is noted that the system responds positively to this test. The speed follows its new reference, this means that the regulation is robust. also, the torque undergoes a transaction peak when switching from one mode to another, then regains its value without error, while the decoupling still exists, so the regulation is robust from a speed control point of view.

II.10 .2 - Simulation of the inversion of the rotationnel direction :

To perform this test, the sign of the speed was reversed from $\left(+157\ rad/s \xrightarrow{to} -157\ rad/s\right)$

from $t = 1.2\ s$. with application of a resistive torque $Cr = 25\ N.m.$ in $t = 0.8s$ The simulation results are illustrated by Fig (II-16):

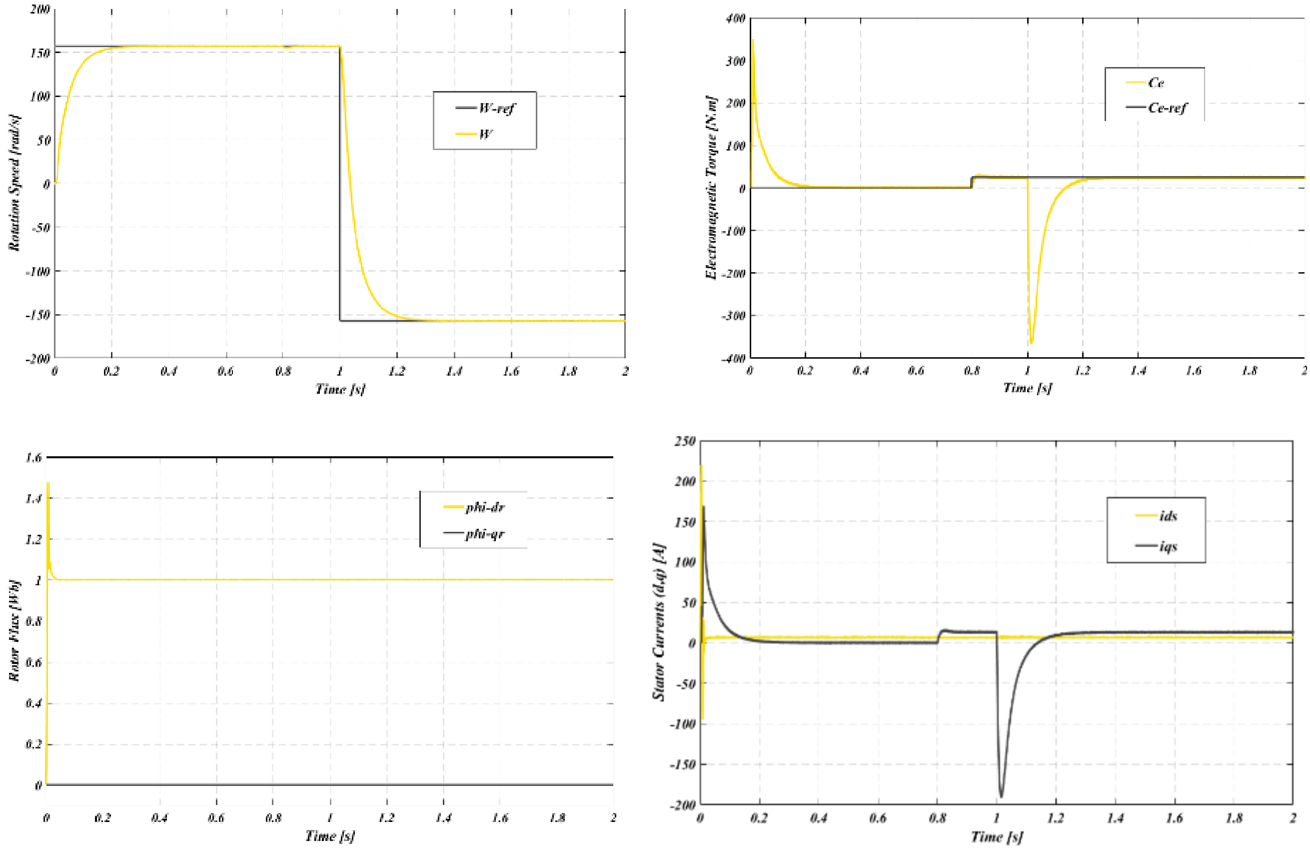


Fig II- 16 : DFOC simulation results during IM shaft speed inversion

I notice from the response shown that the speed is obtained without overshoot despite the dynamics of the flux. This shows that the analytical approach proposed for the design of the PI regulator is quite strict. During start-up, I also observe an overrun of the electromagnetic torque due to the initialization of the flux.

It can be seen that the system responds successfully to this type of test and the decoupling between the flux and the torque is verified. So, I can say that my control is robust with respect to variations in load and direction of rotation.

II.10 .3 - Simulation with load torque variation:

To test the robustness of the regulation, we simulated a no-load start for a reference speed of 157 rad/s, then a cyclic change of different levels of load torque which are applied to the IM by time is shows in the following table:

$Time\ (sec)$	0	0.8	1.1	1.4	1.6	1.8
$C_{resist}\ (N\cdot m)$	0	25	15	−20	−10	0

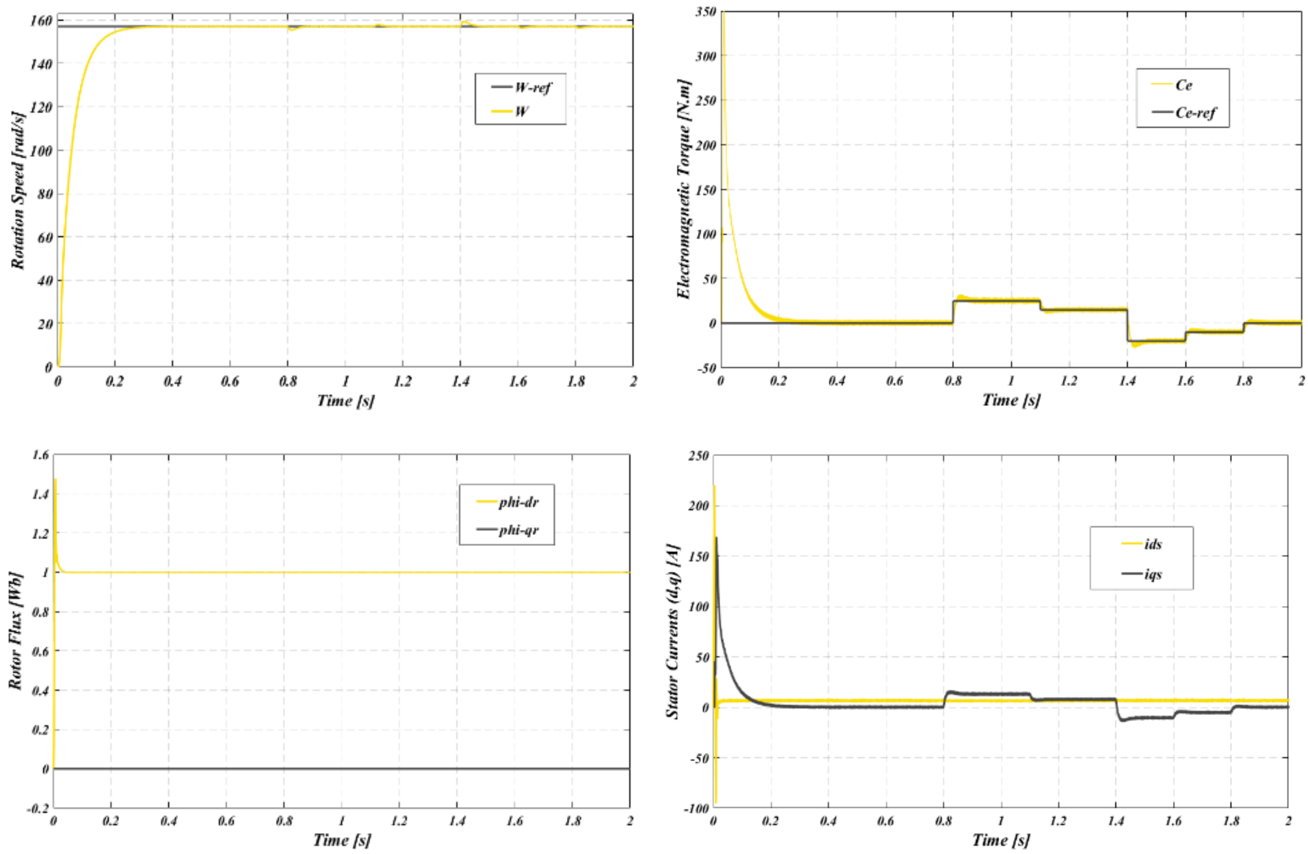


Fig II- 17 : DFOC simulation results during no-load starting followed by a load torque variation

The results of simulation of the direct vector control (DFOC) of the Induction Motor in voltage are illustrated by the Fig (II-17), the system is subjected to the test of follow-up of the instruction to the variation of the load, the quantities such that speed, torque, fluxes and currents are influenced by this variation from which the system is perfectly controlled. The flux curve also shows a decoupling between the electromagnetic torque Ce and the rotor flux, the electromagnetic torque has the same shape as the current isq except for a coefficient which proves that the decoupling is perfectly achieved ($\varphi rq\ =\ 0$) I also note that the electromagnetic torque follows the setpoint, the stator phase current perfectly follows the load variation, the rejection of the disturbance is also achieved with a return to the speed setpoint.

II.10 .4 - Robustness test for the variation of the rotor resistance:

The performances of the DFOC of the induction motor against parametric drifts are tested for a variation of the rotor resistance. However, a 50% increase in resistance Rr from $t = 1.4\,s$ with application of a resistive torque $Cr = 10\,N.m.$ in $t = 0.8s$ the results are shown in Fig (II-18).

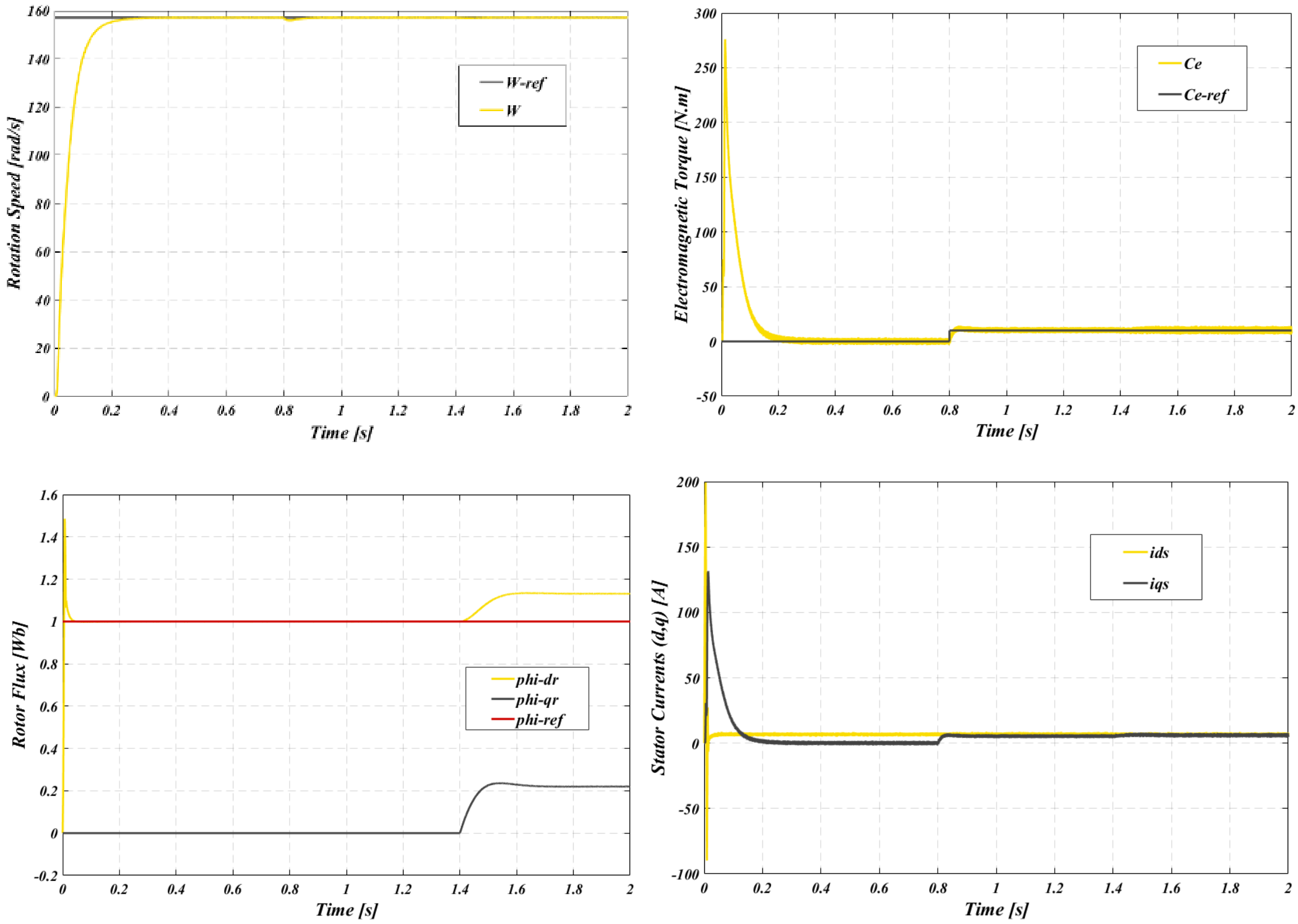


Fig II- 18 : DFOC simulation results of robustness test for rotor resistance variation.

After the results presented in Fig (II-18), we notice at the moment of variation of the rotor resistance of 50% of the rated resistance due to the heating of the machine. which stabilize at their steady state.

The results show that before the moment (t = 1.4s) that is to say at the moment of variation of the resistance of 50%, no variation on the curves of speed, torque, currents and flux.

From the moment of variation of the resistance, the flux curves are divergent to their desired values. The figure confirms the dependence of the control law of the DFOC on the rotor resistance, so I can say that this technique is not robust with respect to internal parametric variations.

II. 11 - Conclusion:

The direct oriented field method applied for several years to the induction motors (IM) remains the most popular method. Indeed, this allows us not only to simplify the model of the machine but also to decouple the regulation of the torque and the flux. It makes it possible to make the shape of the torque of the induction motor similar to that of the Direct Current motor (DCM).

The DFOC of the IM that we have developed presents a satisfactory tracking of the reference, but the classic PI regulator does not allow in all cases to control the transitional systems, and in general, the uncertainties of uncertain systems.

Moreover, after the robustness tests, I did for the variation of speed, the load and the variation of rotor resistance. The results show that the direct vector control is sensitive to the parametric variation (rotor resistance).

These are the reasons why I devote a chapter to the application of a Fuzzy Logic Controller (FLC).

:Chapter_III:

“Application of Fuzzy Logic Control on DFOC”

III.1 - Introduction:

Fuzzy logic is a mathematical description of a process based on fuzzy set theory. This theory introduced in 1965 by Professor Lotfi Zadeh. At that time the theory of fuzzy logic was not taken seriously. Indeed, computers, with their exact operation by all or nothing (1 or 0), have started to spread on a large scale. On the other hand, fuzzy logic made it possible to deal with non-exact variables whose value can vary between 1 and 0. Initially, its goal is, as in classic automatic, to deal with process control problems, that is to say to manage a process according to a given instruction, by acting on the variables which describe the process, but its approach is different from that of traditional automatic control. It most often uses the knowledge of experts or qualified operators working on the process. [53]

The first applications of fuzzy logic were confined to non-technical fields, such as commerce and management, and it was not until 1974 that it was applied automatically by E. H. Mamdani who began by making the first fuzzy controller. In this first section, I deal with two fundamental notions: fuzzy and fuzzy logic and subsets. [54]

The purpose of this chapter is to first represent a brief reminder of fuzzy sets and a general overview of fuzzy logic, as well as its application for adjusting the speed of the three-phase asynchronous machine by replacing the conventional speed regulator. control by linearization input output by a fuzzy regulator.

The performance of this setting will be shown by simulation results, and robustness tests will also be performed.

III.2 - Overview of the history and applications of fuzzy logic:

The roots of fuzzy logic are found in Heisenberg's uncertainty principle. In the 1920s, physicists introduced the third value ½ into the bivalent logic system {0,1}. Why ½? Because it is the truth value of all paradoxes. Classical logic forbids all paradoxes by its axioms. [54]

In the early 1930s, the Polish logician Jan Lukasiewicz developed the logical system with three values and then extended it to all rational numbers between 0 and 1. He defined fuzzy logic as logic that uses the general truth function which can take any value between 0 (false) and 1 (true). [55]

In the 1930s, Max Black applied fuzzy logic to sets of elements or symbols. He called the uncertainty of these sets' imprecision. He drew the first membership function of a fuzzy set. [55]

In 1965 Lotfi Zadeh published the article "fuzzy sets" in which he developed the theory of fuzzy sets and introduced the term fuzzy in the technical literature. This is the beginning of attempts to model systems by fuzzy relations. Zadeh's first investigations were the use of logic to represent an "expert system" approach to automatic tuning, where the control rule is replaced by fuzzy rules.

The first results in fuzzy control were published by Mamdani and Assilian in 1975. This encouraged different activities in England, Denmark and France. [55]

After 1980, research stopped in Europe but the Japanese resumed it. Their industry has launched many products based on fuzzy logic, including household appliances and audio-visual equipment. Currently, fuzzy logic is considered a basic tool in Japan. [53] [55]

Since then, fuzzy logic has experienced a real boom in Japan due to the fact that Japanese companies have quickly understood its advantages, both technical and commercial: [56]

- Ease of installation.
- Solutions of complex multivariate problems.
- Robustness against uncertainties.
- Possibility of integrating the know-how of the expert.

We then see a real panoply of products stamped “Fuzzy logic inside”,

of which we will cite the following examples:

- Household appliances (washing machine, vacuum cleaner, pressure cooker, etc.)
- Audio-visual systems (autofocus camera, camcorder with image stabilizer, photocopier, etc.)
- Embedded automotive systems (suspension, air conditioning, etc.)
- Transport systems (train, metro, lift, etc.)

Finally, it is interesting to note that in recent years fuzzy processors have appeared on the market, and are true processors dedicated to fuzzy logic tuning applications.

III.3 - Advantages and disadvantages of fuzzy logic control:

Fuzzy logic control combines a number of advantages and disadvantages. [57] [58]

III.3.1 - The advantages of fuzzy logic control are:

- The non-necessity of modeling (however, it may be useful to have a suitable model).
- The possibility of implanting (linguistic) knowledge of the process operator.

- Control of the process with complex behavior (highly non-linear and difficult to model).
- Frequently obtaining better dynamic performance (non-linear regulator).
- Two solutions are possible: software solution (by microprocessor, DSP and PC) or hardware solution (by fuzzy processors).

III.3.2 - The disadvantages of fuzzy logic control are:

- Lack of precise guidelines for designing a setting (choice of quantities to be measured, determination of fuzzification, inferences and defuzzification).
- The artisanal and non-systematic approach (implementation of operators' knowledge is often difficult).
- The impossibility of demonstrating the stability of the control circuit in all generality (in the absence of a valid model).
- The possibility of occurrence of limit cycles due to non-linear operation.
- Consistency of inferences not guaranteed a priori (appearance of contradictory rules of inference possible).

III.4 - Fuzzy logic in industry:

In recent years, we have seen the emergence, particularly in the United States, of several applications using control systems based on the theory of fuzzy sets. The main areas of research and application of fuzzy logic are: [57] [59]

- Automation of iron and steel production, water purification, assembly line and manufacturing robots.
- Instrument control (sensor and measuring instruments), and voice and character recognition.
- Design, judgment and decision (consultation, investment and development of train schedules).
- Control of arithmetic units, microcomputers, and realization of operators.
- Processing of information such as data, research of information, modeling of systems, etc.

III.5 - Fuzzy set theory:

III.5.1 - Definition:

The term “fuzzy logic” has two aspects: [53] [60]

- The first corresponds to all the developments concerning the theory of fuzzy sets.

- The second represents an extension of classical logic in order to reason about imperfect knowledge.

In order to know the fundamental principle of fuzzy logic, we introduce a simple example, that of the classification of people into three sets “young”, “middle-aged” and “old”. For the case of classical logic (loop logic) which admits two values 0 or 1, the classification could be done as in the Fig (III-1). All persons aged under 30 belong to the young set and all persons aged over 50 are considered to belong to the “old” set. However, such a classification logic is not even logical because the question that arises:

why should a 50-year-old person be considered as belonging to the “old” group? In reality such a passage is done gradually and individually. In addition, during the classification by classical logic, people located in the "between two ages" zone are not taken into consideration.

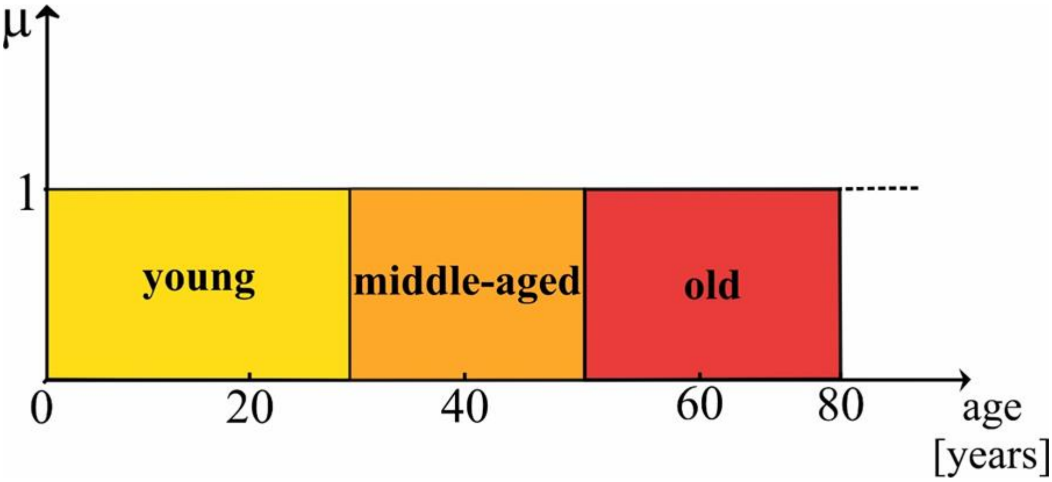


Fig III- 1: Classification of people into three sets according to classical logic.

Fuzzy logic, whose membership function can take any value between 0 and 1, makes it possible to take this reality into account. It is therefore possible to find another classification for the previous example using fuzzy logic. The limits do not vary suddenly but gradually as shown in the Fig (III-2).

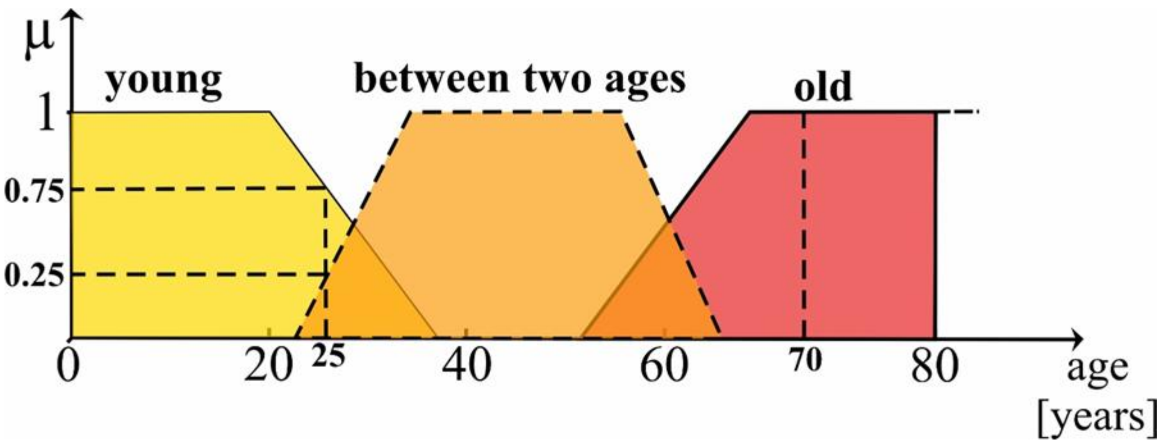


Fig III- 2 : Classification of people into three sets according to fuzzy logic.

A 25-year-old person belongs to the “young” set with a value of the membership function $\mu = 0.75$ and to the “between two ages” set with $\mu = 0.25$, on the other hand, a 70-year-old person belongs with a value $\mu = 1$ to the “old” set.

III.5.2 - Fuzzy sets:

In conventional set theory, a thing either belongs or does not belong to a certain set. However, in reality, it is rare to encounter things whose status is precisely defined. For example, where exactly is the difference between a tall person and an average tall person? It is from this kind of observation that Zadeh developed his theory. He defined fuzzy sets as Linguistic terms of the kind: zero, large, negative, small... These terms also exist in conventional sets. [61]

However, what differentiates these two set theories comes from the limits of sets. In fuzzy sets, it is allowed for a thing to partially belong to a certain set; this is called the membership degree. In conventional sets, the degree of membership is 0 or 1 then in fuzzy logic, the degree of membership becomes a function which can take a real value between 0 and 1 (we then speak of a membership function μ). [54] [61]

III.5.3 - Different forms of membership functions:

We can use for the membership functions different forms Fig (III-3) [61] [62]

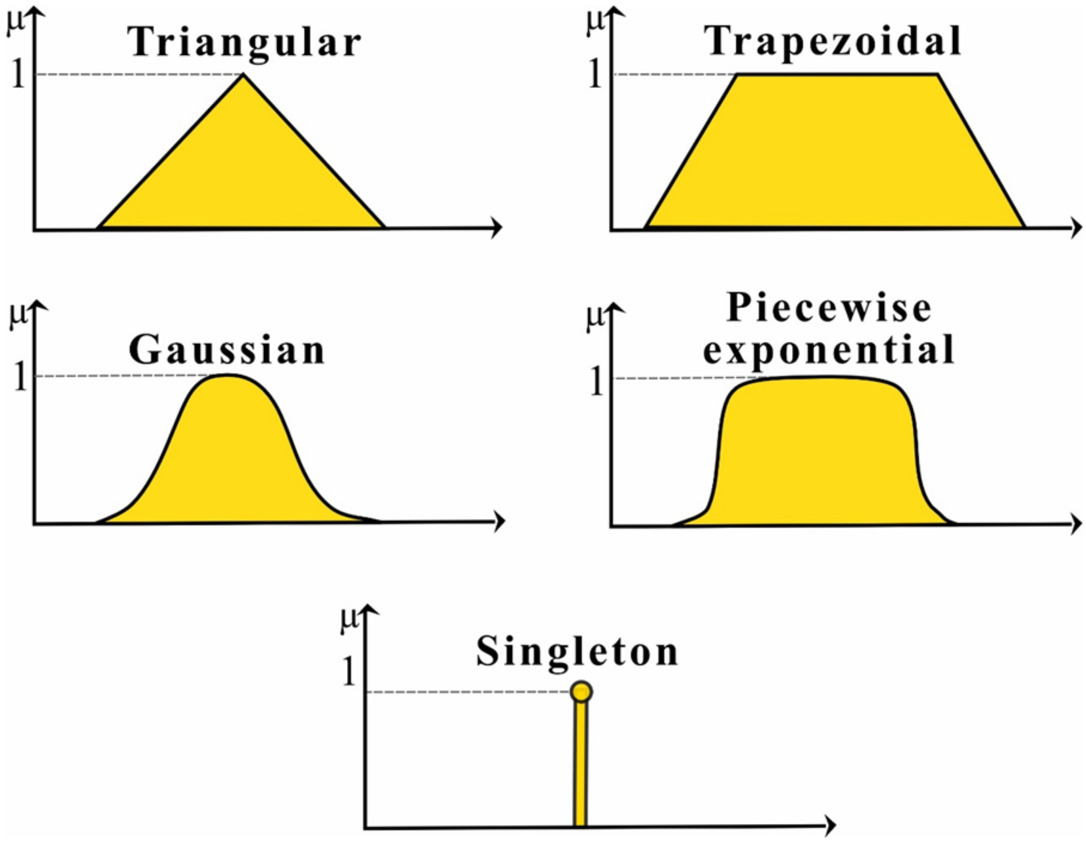


Fig III- 3 : Different forms for membership functions.

In a discrete $X = \{x_i \longrightarrow i = 1,2,3 \cdots n\}$ or continuous X domain, a fuzzy set A can be defined by a set of peers: (degree of membership/element):

$$A = \frac{\mu_A(x_1)}{x_1} + \frac{\mu_A(x_2)}{x_2} + \cdots + \frac{\mu_A(x_n)}{x_n} = \sum_{i=1}^{n} \frac{\mu_A(x_i)}{x_i} \implies Discrete\ case \text{----(III- 1)}$$

$$A = \int_x \frac{\mu_A(x)}{x} \quad \Longrightarrow \quad case\ continued \text{------------(III- 2)}$$

In a continuous domain, fuzzy sets are defined analytically by their membership functions. We find in the literature various forms for fuzzy membership functions.

III.5.3.1 - Triangular and trapezoidal membership function:

$$\mu(x; a, b, c, d) = \max\left\{0; \min\left(\frac{x-a}{b-a}, 1, \frac{d-x}{d-c}\right)\right\} \text{------------(III- 3)}$$

Where $a, b, c\ and\ d$ are the coordinates of the apexes of the trapezium If $b < x \leq c$, we get a triangular membership function.

$$\boldsymbol{\mu(x; a, b, c, d)} = \begin{cases} \mathbf{0} & \xrightarrow{if} \quad x \leq a \\ \frac{x-a}{b-a} & \xrightarrow{if} \quad a \leq x \leq b \\ \frac{c-x}{c-b} & \xrightarrow{if} \quad b \leq x \leq c \\ \mathbf{0} & \xrightarrow{if} \quad x \geq c \end{cases} \text{------------(III- 4)}$$

III.5.3.2 - Piecewise exponential membership function:

$$\boldsymbol{\mu(x; c_g, c_d, w_g, w_d)} = \begin{cases} e^{-\left(\frac{x-c_g}{2w_g}\right)^2} & \xrightarrow{if} \quad x < c_g \\ e^{-\left(\frac{x-c_d}{2w_d}\right)^2} & \xrightarrow{if} \quad x > c_d \\ \mathbf{1} & otherwise \end{cases} \text{------------(III- 5)}$$

Where c_g and c_d are the left and right limits, respectively, and w_g , w_g are the left and right widths, respectively. For " $c_g = c_d$ " and " $w_g = w_d$ ".

III.5.3.3 - The Gaussian membership function:

$$\boldsymbol{\mu(x; c, \sigma)} = e^{-\left(\frac{x-c}{2\sigma}\right)} \text{------------(III- 6)}$$

where c is the center of the Gaussian and σ its width.

III.5.3.4 - Singleton membership function:

$$\boldsymbol{\mu(x)} = \begin{cases} \mathbf{1} & \xrightarrow{if} \quad x = x_0 \\ \mathbf{0} & otherwise \end{cases} \text{------------(III- 7)}$$

In general, three geometric shapes are used for membership functions: trapezoidal, triangular and bell. The first two forms are most often used because of their simplicity.

III.5.4 - Fuzzy logic operators:

The characterization of fuzzy sets by membership functions has allowed an extension of certain operations defined on classical sets to the fuzzy case.

Let A and B be two fuzzy sets defined in the universe of discourse X having respectively μ_A and μ_A as membership function.

The usual operators, namely the addition, subtraction, division and multiplication of two or more fuzzy sets also exist. However, these are the two union and intersection operators most often used in fuzzy logic control. [54] [63]

III.5.4.1 - "NON" Operator:

According to set theory, the complementary set:

$C = \bar{A} = NON(A)$ -- (III- 8)

Is defined by the elements of x that do not belong to the set A. In the case of fuzzy logic, this definition can be expressed by the membership functions in the following way:

$\mu_C(x) = 1 - \mu_A(x)$ -- (III- 9)

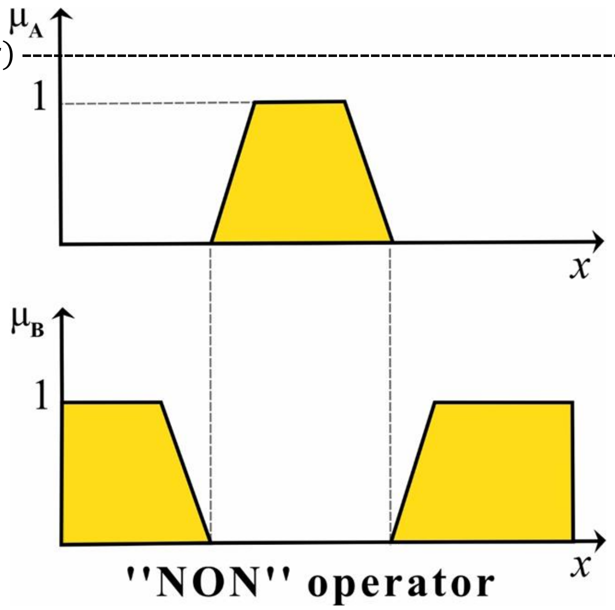


Fig III- 4 : NON-Operator.

III.5.4.2 - "AND" Operator:

The "AND" operator corresponds to the intersection of two sets A and B, we write:

$C = A \cap B = A \; and \; B$ --- (III- 10)

In fuzzy logic, the "AND" operator is realized in most cases by the formulation of the minimum, applied to the membership functions $\mu_A(x)$ and $\mu_B(x)$ of the two sets A and B, that is:

$\mu_C(x) = min[\mu_A(x), \mu_B(x)]$ -- (III- 11)

We then speak of the minimum operator; this operation is represented in the Fig (III-5).

III.5.4.3 - "OR" Operator:

The "OR" operator corresponds to the union of two sets A and B, so we have:

$$C = A \cup B = A \; Or \; B \text{--} \quad \text{(III- 12)}$$

The realization of the operator 'OR' at the level of fuzzy logic is generally done by the formulation of the maximum, applied to the membership functions $\mu_A(x)$ and $\mu_B(x)$ of the two sets A and B . So, we have the maximum operator:

$$\mu_C(x) = max[\mu_A(x), \mu_B(x)] \text{--} \quad \text{(III- 13)}$$

This operation is represented in Fig (III-5).

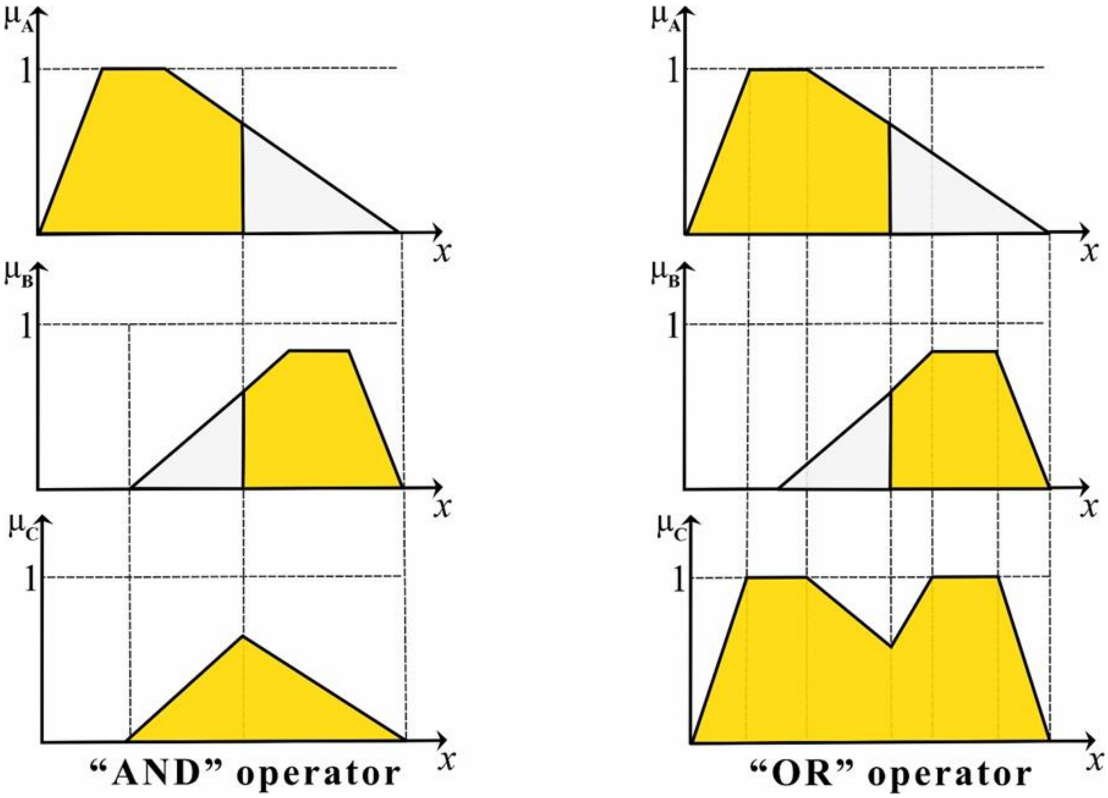


Fig III- 5 : AND, OR Operator

III.5.5 - Fuzzy reasoning:

The natural mode is too complex. These measurements are subject to inaccuracies, vague and possibly uncertain. The reasoning of this knowledge in classical logic is not enough, we use fuzzy reasoning. [64] [65]

III.5.5.1 - Fuzzy implication:

A conditional expression of the type "if x is A then y is B", where A and B are fuzzy sets on the universes U and V respectively a fuzzy relation R on the Cartesian product $U \times V$ is called "Fuzzy rule".

x and y being the linguistic variables described respectively by A and B, R is characterized by a membership function $\mu_R(x, y)$.

A fuzzy rule is based on the notion of fuzzy implication. Thus, the rule "if x is A then y is B" can be written as (x, y) is $A \rightarrow B$, where $A \rightarrow B$ is a fuzzy implication characterized by the truth vector $\mu_{A\rightarrow B}(x, y)$ which is only $\mu_R(x, y)$, that is:

$$\mu_R(x, y) = \mu_{A\rightarrow B}(x, y) = \gamma[\, \mu_A(x), \mu_B(y)\,] \quad \text{(III- 14)}$$

γ is a specific fuzzy implication operator. There are many operators in fuzzy logic.

The most used operators in fuzzy control are the implications of "Mamdani and Larsen":

III.5.5.1.a - <u>The implication of mamdani:</u>

$$\gamma[\, \mu_A(x), \mu_B(y)] = min[\, \mu_A(x), \mu_B(y)] \quad \text{(III- 15)}$$

III.5.5.1.b - <u>The implication of larsen:</u>

$$\gamma[\, \mu_A(x), \mu_B(y)] = \mu_A(x) \times \mu_B(y) \quad \text{(III- 16)}$$

III.5.6 - The generalized modus ponens:

The goal of approximate reasoning is to build a deductive process with the objective of determining a precise conclusion from imprecise facts and a set of fuzzy rules. Such a process is very suitable for the qualitative description of the behavior of systems. A special case of reasoning is the MPG whose inference mechanism is: [65]

$$\frac{\begin{array}{c} x \;\; is \;\; A' \\ if\ x\ is\ A\ then\ u\ is\ B \end{array}}{y\ is\ B'} \quad \text{(III- 17)}$$

The values of y on V are given by the projection of $R = A' \cap (A \rightarrow B)$ on V, that is. Passing to the membership functions:

$$\forall y \in V,\ \mu_{B'}(y) = sup_{x\in U} min[\, \mu_{A'}(x), \mu_{A\rightarrow B}(x, y)] \quad \text{(III- 18)}$$

If A' is a fuzzy singleton, i.e., $A' = \{x_0\}$ with $\mu_{A'}(x) = 1$ for $x = x_0$ and 0 elsewhere, then B' is given by the reduced expression:

$$\mu_{B''}(y) = \mu_{A\rightarrow B}(x_0, y) \quad \text{(III- 19)}$$

III.6 - Control using fuzzy logic:

Experiments have shown that, in many cases, the results obtained with a fuzzy controller (unconventional technique) are better than those obtained with conventional control algorithms. Thus, fuzzy logic control can be seen as a step towards a rapprochement between precise mathematical control and human decision-making. [66]

Fuzzy logic control is on the rise. Indeed, this method makes it possible to obtain a tuning law that is often very effective without having to carry out in-depth modeling. As opposed to a standard regulator or a state feedback regulator, the fuzzy logic (FLC) regulator does not deal with a well-defined mathematical relation, but uses inferences with several rules, based on linguistic variables. By inferences with several rules, it is possible to consider the experiences acquired by the operators of a technical process. [61] [66]

In the following, we will introduce the general basics of fuzzy logic control and the general procedure for designing a fuzzy logic control. We will detail the steps of designing an FLC to control the speed of the induction machine.

III.6.1 - Structure of a fuzzy logic control:

A fuzzy regulator is a particular knowledge-based system composed of four main modules, namely: (the rule base, the fuzzification, the inference engine and the defuzzification) as shown in Fig (III-6): [54]

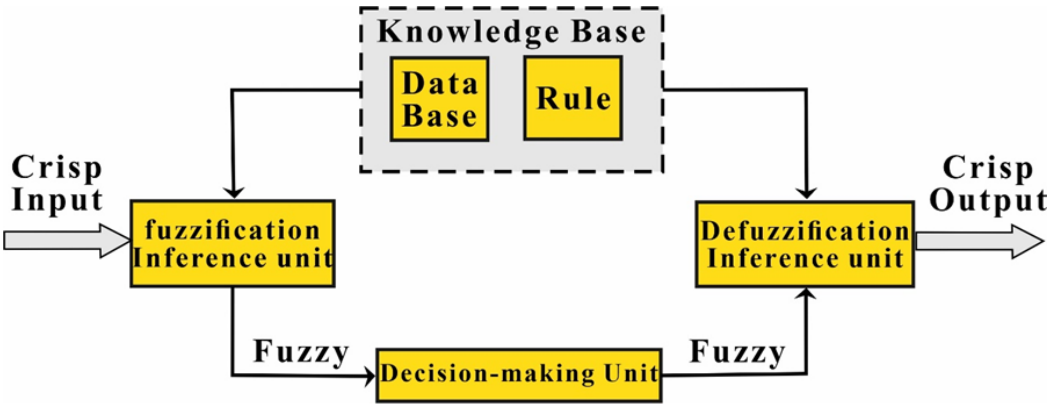


Fig III- 6 : Structure of a fuzzy controller

III.6.1.1 - The fuzzification interface:

This interface performs the following functions: [61]

- The definition of membership functions for input variables.
- The passage of physical quantities into linguistic variables which can thus be treated by the rules of inference.

There are two fuzzification techniques:

- singleton fuzzification.
- non-singleton fuzzification.

Singleton fuzzification is the most used in control, it interprets a digital point μ_0 as a fuzzy set A in X, having membership function $\mu_A(x)$

With:

$$\mu_A(x) = \begin{cases} 1 & \xrightarrow{if} \quad x = x_0 \\ 0 & \xrightarrow{if} \quad x \neq x_0 \end{cases} \quad \text{--} \quad \text{(III- 20)}$$

III.6.1.2 - Rules base:

The fuzzy rule base, or knowledge base, contains fuzzy rules describing the behavior of the system, it is the heart of the entire system in the sense that all other components are used to interpret and combine these rules to form the final system. it is composed of: [56]

1. A database providing the information necessary for standardization functions

2. the rule base is a set of linguistic expressions structured around expert knowledge, and represented in the form of rules:

$$\boldsymbol{If\ condition} \xRightarrow{\boldsymbol{Then}} \boldsymbol{consequence}$$

III.6.1.3 - Standardization:

This first step allows the processing of the input variables of the fuzzy controller. For example, calculation of errors (difference between quantities measured and setpoints) and variation of errors. The use of normalized domain (Universe of speech between $[-1, 1]$) requires a transformation of scale, this one is carried out via scale factors of transformation of the physical sizes of the inputs into values normalized belonging to the interval $[-1\ ,1]$.[67]

III.6.1.4 - Inference engine:

Inference or decision making is the core of the fuzzy controller. It has the ability to simulate human decision-making based on fuzzy concepts and expertise. Inference can be described explicitly by linguistic description using a number of rules. Each rule has a condition preceded by the symbol IF and a conclusion, action or operation, preceded by the symbol $THEN$. Depending on the tuning strategy adopted. For the presentation of the different possibilities of expressing the inferences, we choose an example of a system to be regulated with two fuzzy variables x_{in1} and x_{in2} which form the input variables of the inference, and an output variable x_{out} also expressed as variable blurry. Inference rules can be described in several ways. [54]

The linguistic description of the inferences can be written as follows:

IF (x_{in1} is large negative **AND** x_{in2} is approximately zero) **THEN** (x_{out} is large negative)

IF (x_{in1} is large negative **AND** x_{in2} is medium positive) **THEN** (x_{out} is medium positive)

A rule's condition can also contain **OR** and **NOT** operators, and the rules are determined according to the adopted tuning strategy.

- Symbolically it is in fact a linguistic description where the designation of fuzzy sets is replaced by abbreviations.

IF (x_{in1} NG **AND** x_{in2} EZ) **THEN** $x_{out} = NG$

IF (x_{in1} NG **AND** x_{in2} PM) **THEN** $x_{out} = PM$

And so on.

- By inference matrix it gathers all the inference rules in the form of a table.

In the case of a two-dimensional matrix, the entries of the table III.2 represent the fuzzy sets of the input variables ($x_{in1} and\ x_{in2}$). The intersection of a column and a row gives the fuzzy set of the output variable (x_{out}) defined by the rule. There are as many boxes as there are rules.

If all the cells of the matrix are filled, then we speak of complete inference rules.

The different assemblies are characterized by standard designations shown in table III-1:

Fuzzy Sets	Abbreviations
Big Negative	**BN**
Average Negative	**AN**
Small Negative	**SN**
Approximately Zero	**AZ**
Small Positive	**SP**
Average Positive	**AP**
Big Positive	**BP**

Tab III- 1 : Abbreviations of fuzzy sets

x_{out}		x_{in1}						
		BN	AN	SN	AZ	SP	AP	BP
x_{in2}	BN	BN	BN	BN	AN	SN	SN	AZ
	AN	BN	AN	AN	AN	SN	AZ	SP
	SN	BN	AN	SN	SN	AZ	SP	AP
	AZ	BN	AN	SN	AZ	SP	AP	BP
	SP	AN	SN	AZ	SP	AP	AP	BP
	AP	SN	AZ	SP	AP	AP	AP	BP
	BP	AZ	SP	SP	AP	BP	BP	BP

Tab III- 2 : Matrix of complete inferences

For tuning by fuzzy logic, one of the following methods is generally used. [56] [68]

III.6.1.4.a - Max-Min inference method (mamdani):

The max-min inference method is carried out, at the level of the condition of the operator "**AND**" by the formulation of the minimum. The conclusion in each rule, introduced by "**THEN**", links the membership factor of the premise with the membership function of the output variable is achieved by the formation of the minimum. Finally, the "**OR**" operator which links the different rules is realized by the formation of the maximum

III.6.1.4.b - Max-product inference method (larsen):

The max-product inference method is performed, at the condition level, the "**AND**" operator by forming the product. The condition in each rule, introduced by "**THEN**" is realized by the formation of the product. The "**OR**" operator, which links the different rules, is realized by the formation of the maximum.

III.6.1.4.c - Sugeno's method:

The "**AND**" operator is realized by the formation of the minimum, the conclusion of each fuzzy rule has a polynomial form. The output is equal to the weighted average of the output of each fuzzy rule.

III.6.1.5 - Defuzzification interface:

The role of defuzzification is to provide a physical control action from a fuzzy control action, there are several defuzzification strategies, the most used are:

III.6.1.5.a - Maxima method:

Is the simplest, it consists of considering for each output only the rule with the maximum validity. This technique is rarely used because it represents disadvantages when there are several values for which the resulting membership function is maximum.

III.6.1.5.b - Weighted average method:

The defuzzifier examines the fuzzy set which determines the values for which the membership function is maximal, then calculates the average of these values as a defuzzification result.

III.6.1.5.c - Center of gravity (COG) / centroid of area (COA) method:

Is more efficient and gives the best results. It consists in determining the center of gravity of the output membership function using the following relationship:

$$x^* = \frac{\sum_{i=1}^{n} x_i \cdot \mu(x_i)}{\sum_{i=1}^{n} \mu(x_i)} \text{ ---------- (III- 21)}$$

the membership function, and n represents the number of elements in the sample. For continuous membership function, $x^*$is defined as:

$$x^* = \frac{\int x \cdot \mu_A(x) dx}{\int \mu_A(x) dx} \text{ ---------- (III- 22)}$$

The integral in the denominator gives the area, while the integral in the numerator is the moment of the area.

III.6.1.5.d - Center of sums method (COS):

This is the most commonly used defuzzification technique. In this method, the overlapping area is counted twice.

The defuzzified value x^* is defined as:

$$x^* = \frac{\sum_{i=1}^{N} x_i \cdot \sum_{i=1}^{n} \mu_{A_k}(x_i)}{\sum_{i=1}^{N} \sum_{i=1}^{n} \mu_{A_k}(x_i)} \text{ ---------- (III- 23)}$$

Here, n is the number of fuzzy sets, N is the number of fuzzy variables, $\sum_{i=1}^{n} \mu_{A_k}(x_i)$ is the membership function for the $k - th$ fuzzy set.

III.7 - Application of fuzzy logic to IM speed control:

In this part, we will follow the steps shown in the diagram in Fig (III-7) to design a Mamdani type fuzzy logic controller for the speed control loop.

III.7.1 - Designing an FLC:

determined and the dynamic behavior of the process analyzed with respect to the variation of the control quantity, the description may make use of linguistic variables which may be incorporated into knowledge of control theory **(AND/OR**) operational experience. [54]

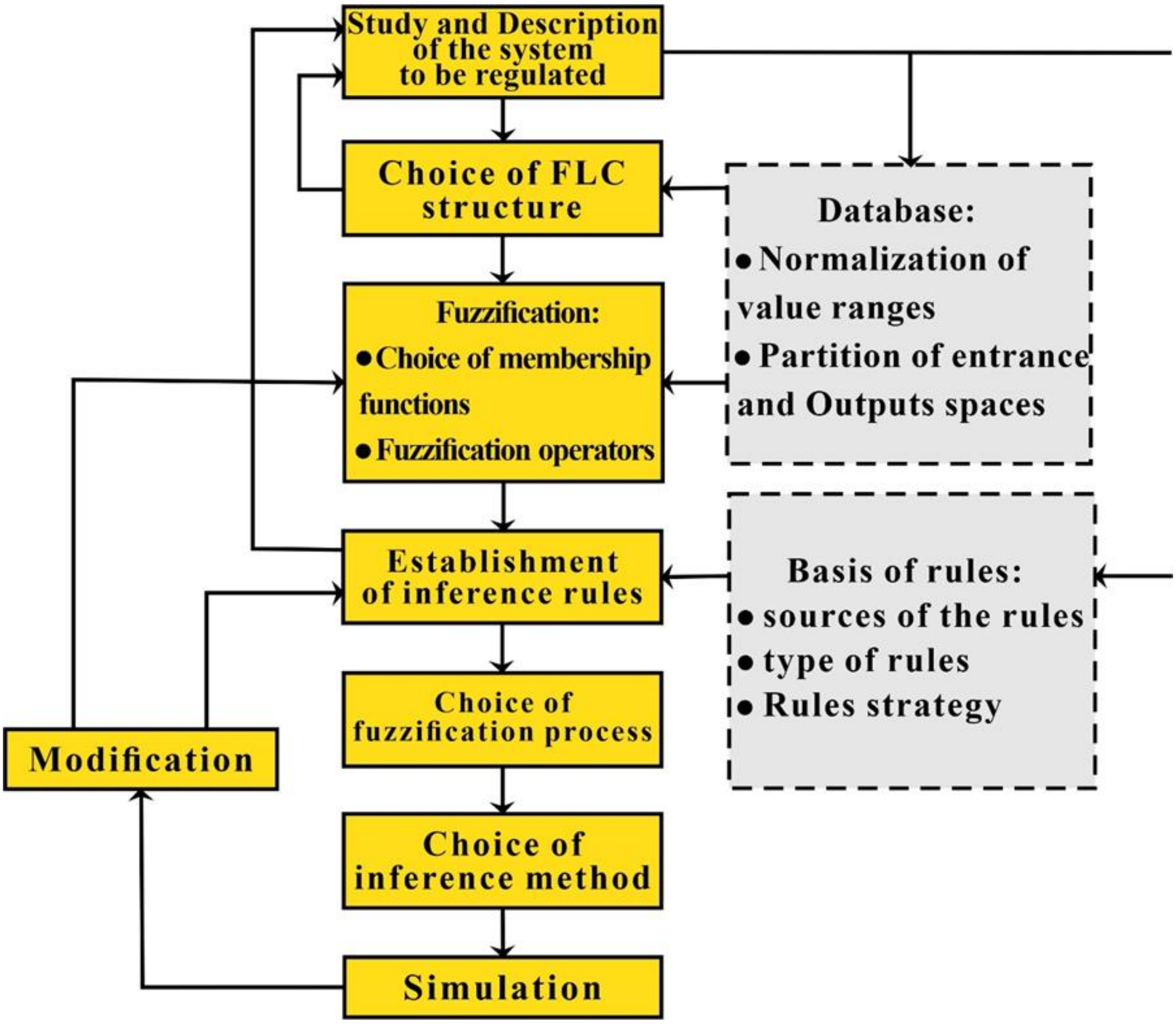


Fig III- 7 : Main steps when designing an FLC.

III.7.2 - Development of a fuzzy controller: [69] [70]

After having stated basic concepts and the linguistic terms used in fuzzy logic, we present the structure of a fuzzy controller. In what follows, we are mainly interested in the speed regulator within a control by linearization input output of the asynchronous machine.

The reference speed can be controlled by an external operator. The output quantity of this speed regulator is the electromagnetic torque.

The developed controller uses the scheme proposed by Mamdani. This diagram is presented by Fig (III-8) it is composed:

- Standardization factors associate with the error (e), its variation (Δe) and the variation of the command (Δu).
- An error fuzzification block and its variation.
- Fuzzy control rules.
- The control strategy is presented by an inference matrix of the same type as that presented in the table III-4.
- A defuzzification block used to convert the fuzzy control variation into a numeric value.
- An integrator.

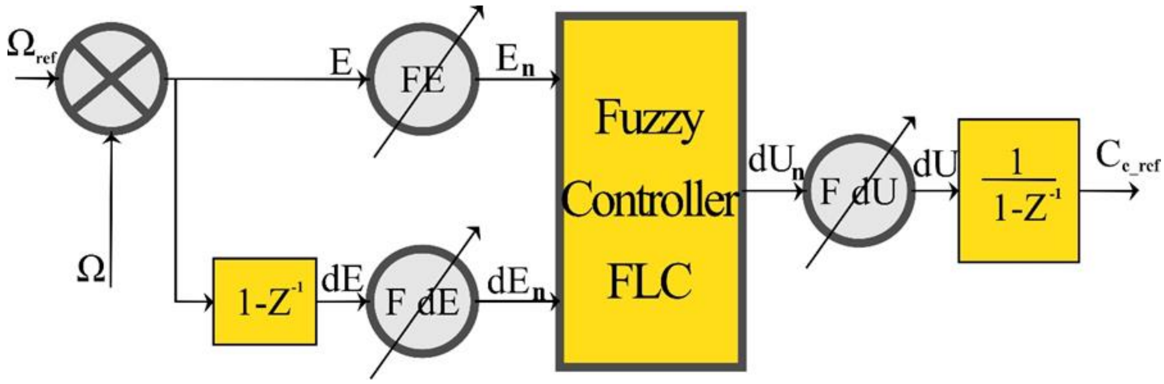


Fig III- 8 : Synoptic diagram of a speed regulator

In the diagram above as in the following, we note:

E: The error is defined by:

$$E(k) = \Omega_{ref}(k) - \Omega(\mathrm{k}) \quad \text{(III- 24)}$$

dE: The derivative of the error, it is approximated by

$$dE(k) = \frac{E(k)-E(k-1)}{T_e} \quad \text{(III- 25)}$$

Te: Being the sampling period.

The regulator output is given by:

$$C_{ref}(k) = C_{ref}(K-1) + dU(k) \quad \text{(III- 26)}$$

Gains called “scale factors” are found at the input and output of the fuzzy controller, which make it possible to change the sensitivity of the fuzzy controller without changing its structure. The quantities indexed “n” are therefore the standardized quantities at the input and at the output of the fuzzy controller.

Inference rules are used to determine the behavior of the fuzzy controller. It must therefore include intermediate steps that allow it to pass from real quantities to fuzzy quantities and vice versa; these are the fuzzification and defuzzification stages Fig (III-8). The fuzzy controller inputs En and dEn are normalized by using the following expressions:

$$E_n = K_e \cdot E \quad \text{(III- 27)}$$

$$dE_n = K_e \cdot dE \quad \text{(III- 28)}$$

Similarly, the output of Un from the controller is denormalized to dU using the following relationship:

$$dU_n = K_e \cdot dU \quad \text{(III- 29)}$$

It is important to choose the ranges of values carefully. A good choice of ranges with a good distribution can guarantee a successful conception. On the other hand, a bad choice leads to long corrections in the following steps, it is often even necessary to redefine the ranges of values in order to avoid failure in the design. A good choice requires experience and knowledge of the system to be controlled.

III.7.2.1 - Fuzzification stage:

The fuzzy sets of input variables (En, dEn) and output variables dUn are defined by membership functions with 7 or 5 or 3 sets Fig (III-9). A very fine subdivision of the universe of discourse on more than seven fuzzy sets does not generally bring any improvement in the dynamic behavior of the system to be regulated.

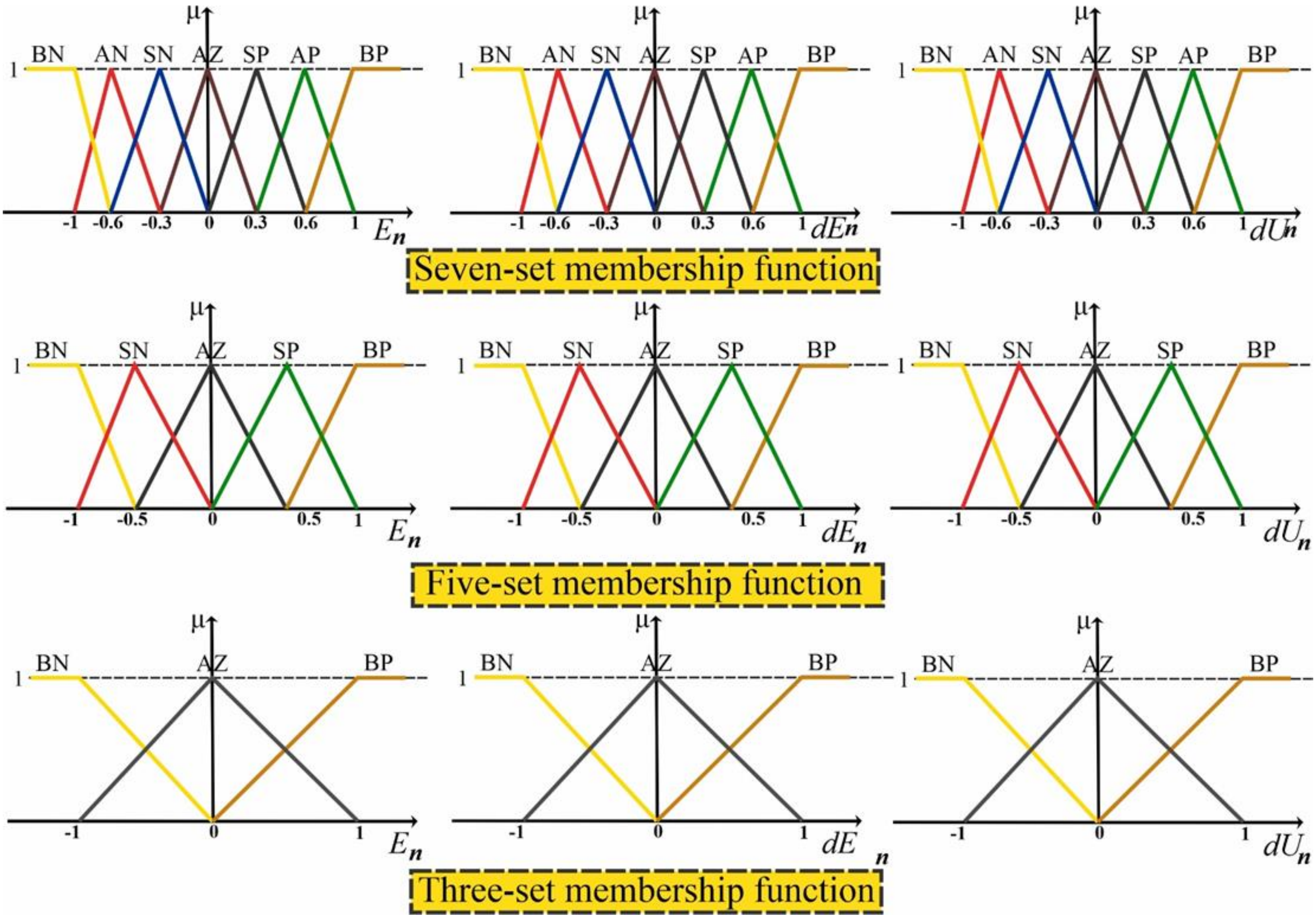


Fig III- 9 : The membership functions

III.7.2.2 - Stage of establishing inference rules:

From the study of the closed-loop system behavior of the velocity based on the experiments, we can establish the control rules, which connect the output with the inputs. As we have seen, there are seven fuzzy sets, which implies forty-nine possible combinations of these inputs, hence forty-nine rules.

The rules are in the following table:

Rules Number	IF	AND	THEN
Number : 1	BN	BN	BN
Number : 5	BN	SP	AN
Number : 39	AP	AZ	AP

Tab III- 3 : Inferences rules table

The inference matrix is shown in the following table, where the 49 rules are illustrated:

dE_w \ E_w		x_{in1}						
		BN	**AN**	**SN**	**AZ**	**SP**	**AP**	**BP**
x_{in2}	**BN**	BN	BN	BN	AN	SN	SN	AZ
	AN	BN	AN	AN	AN	SN	AZ	SP
	SN	BN	AN	SN	SN	AZ	SP	AP
	AZ	BN	AN	SN	AZ	SP	AP	BP
	SP	AN	SN	AZ	SP	AP	AP	BP
	AP	SN	AZ	SP	AP	AP	AP	BP
	BP	AZ	SP	SP	AP	BP	BP	BP

Tab III- 4 : Rules table for speed FLC

For example: $rule45\!: IF\ E_w = \boldsymbol{BP}\ AND\ dE_w = SN\ THEN\ Ci_{qs}{}^* = AP$

III.7.2.3 - Defuzzification:

When the fuzzy output is calculated, it must be transformed into a numerical value. There are several methods to achieve this transformation. The most used is the center of gravity method, which we have adopted in my work.

The abscissa of the center of gravity corresponding to the output of the regulator is given by the following relationship:

$$x^* = \frac{\int x \cdot \mu_A(x)dx}{\int \mu_A(x)dx} \quad \text{(III- 30)}$$

III.8 - Simulation and interpretation:

In this study, we will use the following reference [81] [82].

The block diagram of this simulation is shown in Fig (III-10). The gains of the fuzzy regulator are adjusted by trial and error to achieve the desired performance. Fuzzy Sets of controller inputs and outputs are divided into seven membership functions (BN, AN, SN, AZ, SP, AP, BP) in triangular and trapezoidal shape.

These performances were established from the simulation of the following operating modes: a no-load start followed by the introduction of a load torque, a reversal of the direction of rotation, speed variation test with an application of a torque load, the robustness of the control with parametric variations (rotor resistance) and on the behavior of the system.

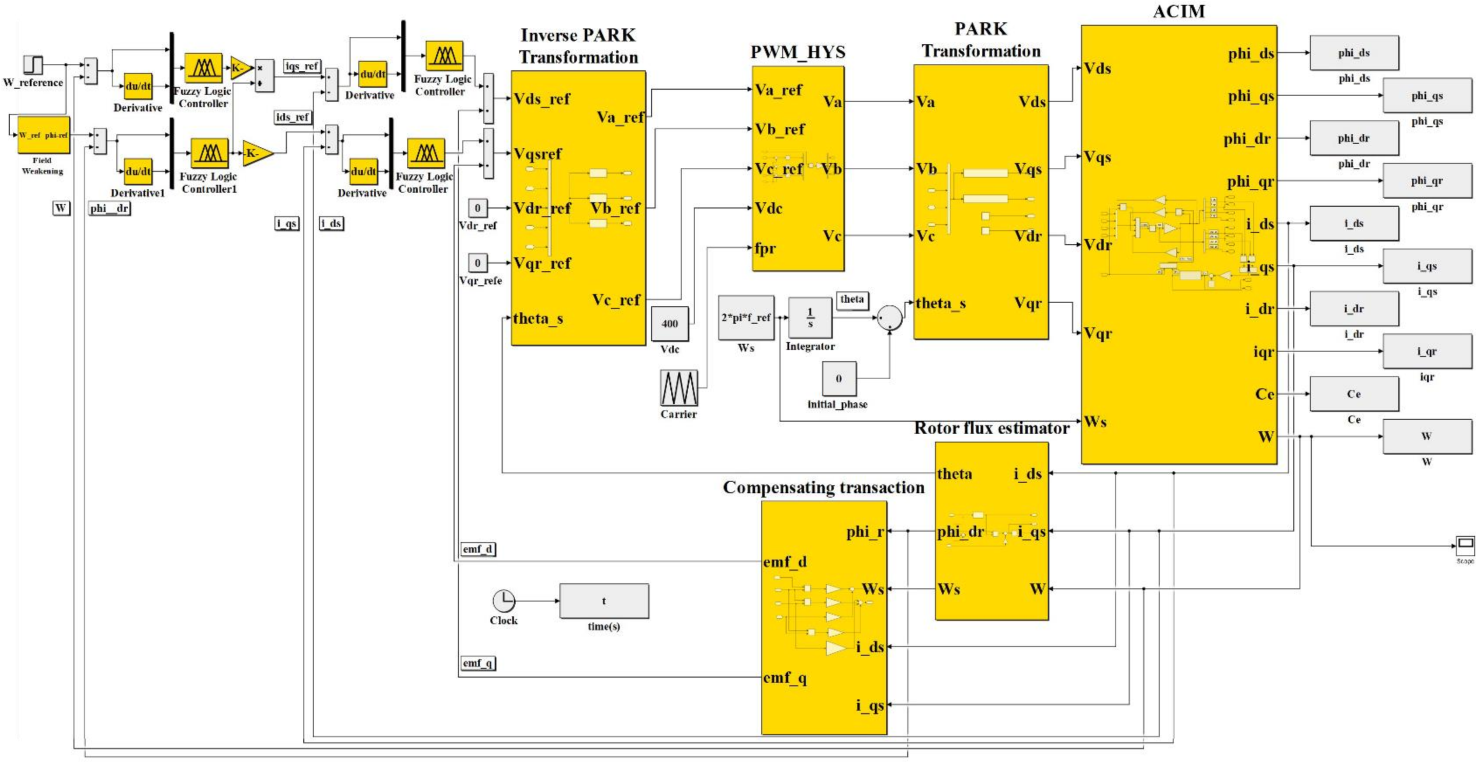


Fig III- 10 : Diagram of a DFOC Based on FLC to Controlling IM.

III.8 .1 - Simulation of the speed variation:

The simulation results obtained for the speed variation of Fig (III-11) are explained in the following table:

$Time\ (sec)$	0	1	1.4
$\Omega_{ref}\ (rad/sec)$	157	90	157

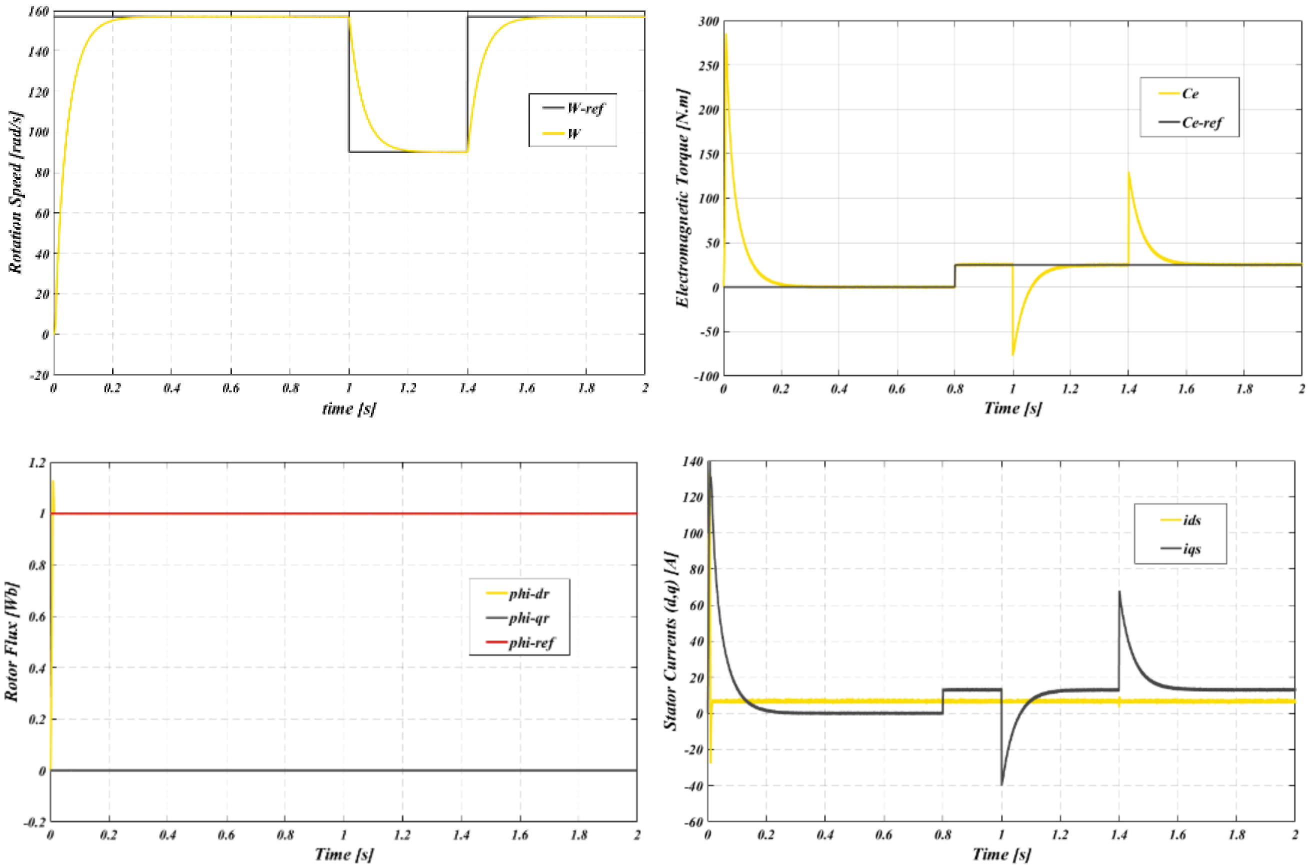


Fig III- 11 : DFOC with an FLC simulation results during IM shaft speed variation

after no loaded start Fig (III-11) shows that the speed regulation response is very satisfactory in all operating intervals. Whether the motor is running at rated speed or at reduced speed, load impacts have Approximately no influence on its value. Also, the torque undergoes a transaction peak when switching from one mode to another.

As I have a separation between the flux and the variation of the speed. The load torque Approximately has no effect on the rotor flux, I find that low speed operation almost has no effect on the magnetic state of the motor. So, let to say that the control is robust to load and speed variations.

III.8 .2 - Simulation of the inversion of the rotational direction:

To perform this test, the sign of the speed was reversed from $\left(+157\ rad/s \xrightarrow{to} -157\, rad/s\right)$

from $t = 1.2\ s$. with application of a resistive torque $Cr\ =\ 25\ N.m$. The simulation results are illustrated by Fig (III-12):

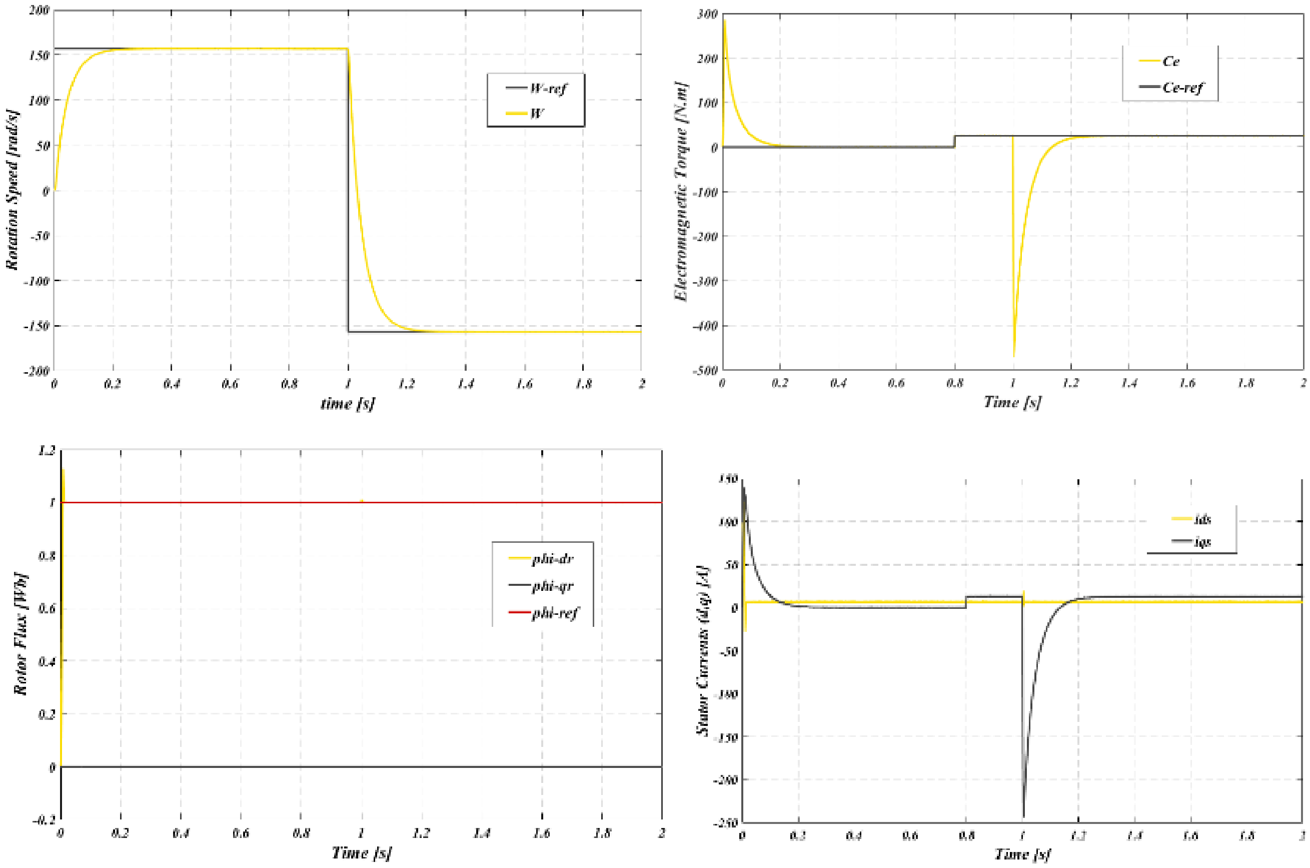


Fig III- 12 : DFOC with an FLC simulation results during IM shaft speed inversion.

I noticed that the speed response is very satisfactory in both areas of fast and precise operation. That the machine rotates at the speed of 157rad/s, from the moment t=1.2s, we noticed that the speed response is very satisfactory in the two fast and precise operating zones. Whether the machine is rotating at a speed of +157rad/s or at a speed in the opposite direction of –157rad/s, the rotation inversion of the speed makes it possible to deduce that the system responds successfully to this type of test and the decoupling between the flux and the torque is verified. So, I can say that my control is robust with respect to variations in load and direction of rotation.

III.8 .3 - Simulation with load torque variation:

To test the robustness of the regulation, we simulated a no-load start for a reference speed of 157 rad/s, then a cyclic change of different levels of load torque which are applied to the IM by time is shows in the following table:

$Time\ (sec)$	0	0.8	1.1	1.4	1.6	1.8
$C_{resist}\ (N \cdot m)$	0	25	15	−20	−10	0

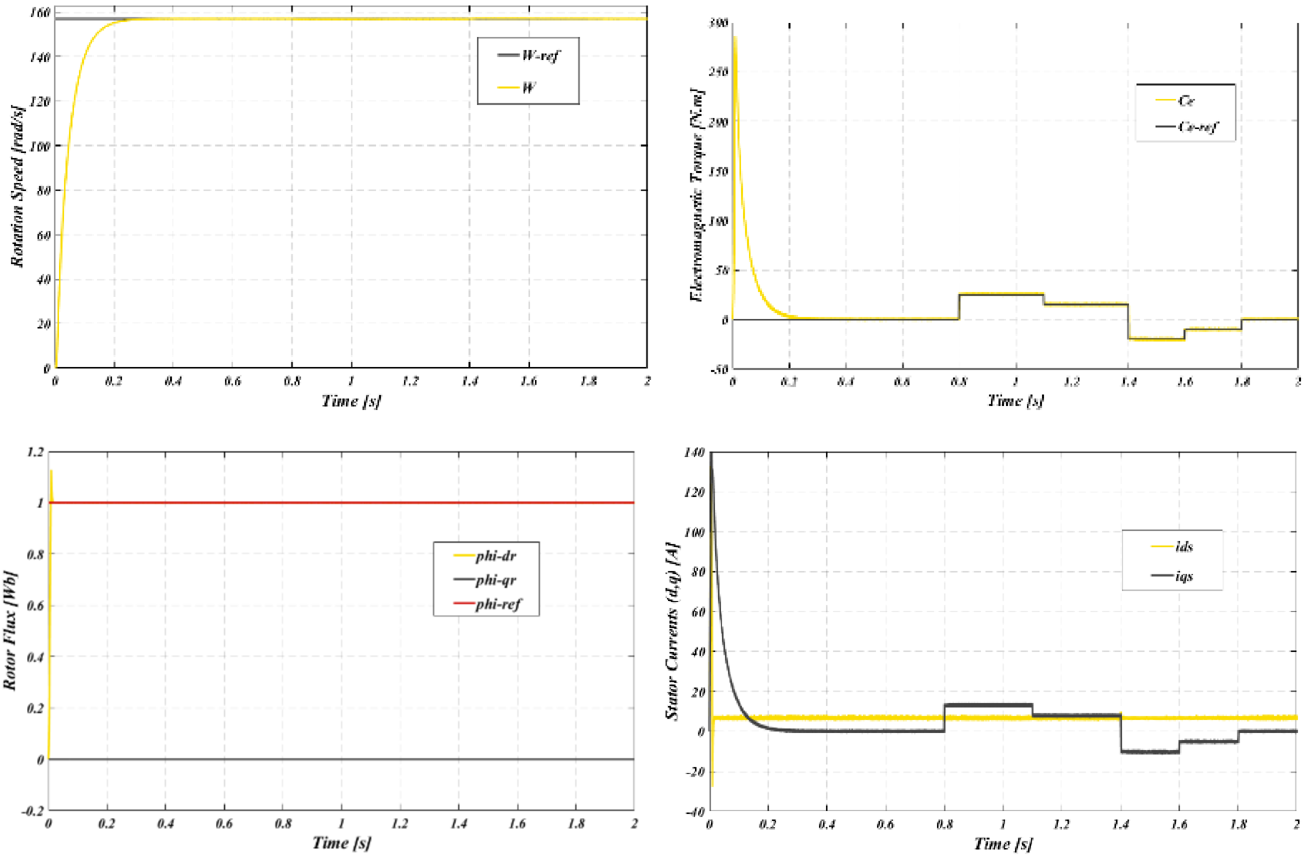


Fig III- 13 : DFOC with an FLC simulation results during no-load starting followed by a load torque variation.

I got a very satisfying speed response with load variation impacts approximately have no influence on its values. I also note that the electromagnetic torque follows the setpoint, the stator phase current perfectly follows the load variation, the rejection of the disturbance is also achieved with a return to the speed setpoint.

These results show the decoupling between the electromagnetic torque and the flux during the application of the load which is quickly rejected by the fuzzy speed controller.

III.8 .4 - Robustness test for the variation of the rotor resistance:

The performances of the DFOC of the induction motor against parametric drifts are tested for a variation of the rotor resistance. However, a 50% increase in resistance Rr from $t = 1.4\ s$ with application of a resistive torque $Cr = 10\ N.m.$ in $t = 0.8s$ the results are shown in Fig (III-14).

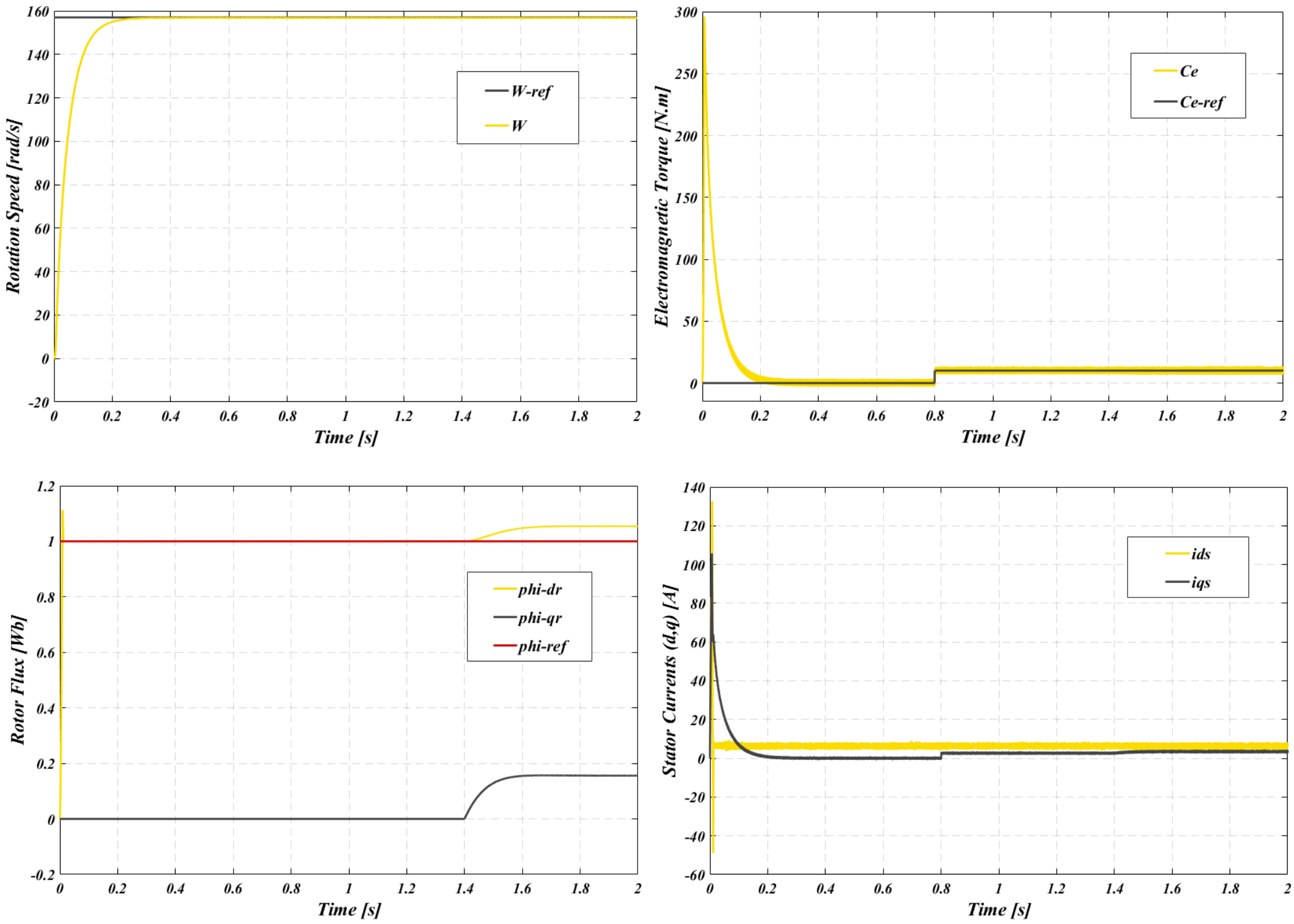


Fig III- 14 : DFOC with an FLC simulation results of Robustness test for rotor resistance variation.

After the results presented in Fig (III-14), we notice at the moment of variation of the rotor resistance of 50% of the rated resistance due to the heating of the machine. which stabilize at their steady state.

The results show that before the moment ($t = 1.4s$) that is to say at the moment of variation of the resistance of 50%, no variation on the curves of speed, torque, currents and flux.

The simulation results also showed an acceptable robustness of this regulator, especially from the point of view of Torque-Flux decoupling during large parametric variations and low speeds.

III.9 - Conclusion:

The basic notions of fuzzy logic were presented at the beginning of this chapter. Aspects of fuzzy logic control, as well as the design of a fuzzy controller have been introduced while justifying our choice of this type of control which lies in its ability to deal with the imprecise, the uncertain and the vague and its simplicity of design. A simulation based on a fuzzy controller was carried out to adjust the speed of an asynchronous induction machine.

The simulation results show an acceptable decoupling between the two subsystems (flux and electromagnetic torque). also, the simulation results showed an acceptable robustness of this regulator, especially from the point of view of Torque-Flux decoupling during large parametric variations and low speeds. The performance of this command is satisfactory, it is clear that for the same operation condition the induction motor speed control using fuzzy controller technique had better performance than the PI control.

General Conclusion

In this thesis, we have presented the control of the asynchronous induction machine via three different control structures: the direct vector control and the control by fuzzy logic

First, we established the mathematical model of the machine according to the modeling with the transformation of PARK, in order to simplify the equations of the asynchronous induction machine in transient state.

Then, we gave the basic principles of the DFOC by orientation of rotor flux, which makes it possible to impose on the asynchronous induction machine a behavior similar to that of the DC machine with separate excitation where the flux is not affected by the variation of the electromagnetic torque. On the other hand, there is a greater complexity of the control.

The results obtained by DFOC clearly show perfect decoupling, but the latter is affected by variations in machine parameters, which is the major drawback of vector control. an alternative to the latter in order to solve this problem, I tried to apply fuzzy logic controlling.

In order to have a better rating of the results obtained by the conventional PI controllers and the fuzzy logic regulator based on vector control, this study was carried out by a comparative study of the performances between the two. We conclude that fuzzy logic Controlling is more robust than PI controllers.

But with all these satisfactory results achieved, it is still tainted by little effect of machine parameters changes.

According to the simulation results presented in chapters III, it was concluded that the regulation by FLC have a good decoupling between the flux and the electromagnetic torque.

Finally, to improve the control performance, we recommend the two control methods, PID and FLC, in terms of separation of the rotor flux from the electromagnetic torque, but in terms of stiffness against the variation of internal parameters of the machine, based on the good results it got.

ANNEX

Parameters and characteristics of the IM used:	
Parameters	R_s (Stator resistance) = 1.2 Ω
	R_r (Rotor Resistance) = 1.8 Ω
	L_s (Stator Inductance) = 0.1554 H
	L_r (Rotor Inductance) = 0.1564 H
	M_{sr} (mutual inductances) = 0.15 H
Mechanical constants	J (Rotor inertia) = 0.0700 $[Kg.m^2]$.
	f (Coefficient of friction) = 0.001 $I.S$
Characteristics	Resistant torque = 25 $N.m$
	Frequency = 50HZ
	Power = 4kW
	Voltage = 220/380V
	Speed = 1440 rpm
	Pole pair = 2
	Power factor cos(φ_{rated})= 0.8
	The carrier frequency f_{pro} = 5000

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