

M^{ed} Boudiaf M'sila University
Electrical Engineering Department

Specialty: **renewable energy in electrical engineering**

Level : MASTER II

Title :

Control Techniques in renewable energy

Techniques de commande des systèmes à énergie renouvelable

Realised by :

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شهادة إدارية
بخصوص مطبوعة الدروس الخاصة بالأستاذ
غضبان إسماعيل

بناءً على محضر اللجنة العلمية لقسم الهندسة الكهربائية تحت رقم: 212/ق.ه.ك/ 2023 المنعقد بتاريخ 08 ماي 2023 والمتضمن تعيين الخبراء: الأستاذ شودار عيسى أستاذ بجامعة المسيلة الأستاذ معيوف المسعود أستاذ بجامعة المسيلة، والأستاذ بن شوية محمد التوفيق أستاذ بجامعة بسكرة وذلك لتقييم مطبوعة الدروس الخاصة بالأستاذ غضبان إسماعيل أستاذ محاضر "أ" بقسم الهندسة الكهربائية لجامعة المسيلة تحت عنوان :

" Techniques de commande des systèmes à énergie renouvelables "

مطبوعة دروس مكتوبة باللغة الإنجليزية تحت عنوان:

" Control Techniques in Renewable Energy "

وبعد إطلاع رئيس اللجنة العلمية ورئيس القسم على التقارير الواردة و التي كانت كلها ايجابية، وعليه فإن اللجنة لا ترى مانعا أن تتخذه سنداً في تدريس طلبة السنة الثانية ماستر طاقات متجددة في الكهروتقني ، شعبة الكهروتقني ميدان علوم و تكنولوجيا و أن تعتمد في أي تقييم للمسار العلمي للأستاذ المعني.

رئيس القسم

رئيس اللجنة العلمية

أ. د. بوقرة عبد الرحمان



د. داف المبروك



ملاحظة: سلمت هذه الشهادة للمعني(ة) لاستعمالها في حدود ما يسمح به القانون.

Abstract :

The integration of renewable energy sources such as solar, wind, hydro, and biomass into the power grid presents unique challenges due to their intermittent and variable nature. Effective control techniques are essential to optimize the performance, stability, and efficiency of renewable energy systems. The course highlights the importance of real-time monitoring, fault detection, and grid synchronization to ensure reliable energy production and distribution. Ultimately, effective control of renewable energy systems is crucial for the sustainable and efficient utilization of clean energy. Teaching objectives, Know the different electrical systems of electric actuators (motor + mechanical loads and static converters) , Be able to establish a simulation model of an electrical system including motor, power electronics and control - Be able to simulate a model in the Matlab/Simulink environment , Be able to size the correctors present in the controls of energy production chains

key words : renewable energy , wind ,solar , control technique , matlab simulink

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Realised by :

GHADBANE Ismail

University year: 2023/2024

Semestre3 Master : Energies renouvelables en Electrotechnique

Semestre: 3

UE Fondamentale Code : UEM2.1.1

Matière : Commande des systèmes à énergies renouvelables

VHS:45h00 (Cours: 1h30, TD 1H30)

Crédits: 4

Coefficient: 2

Objectifs de l'enseignement :

- Connaître les différents systèmes électriques d'actionneurs électriques (moteur+ Charges mécanique et convertisseurs statiques)
- Être capable d'établir un modèle de simulation d'un système électrique comprenant moteur, électronique de puissance et commande
- Être capable de simuler un modèle dans l'environnement Matlab/Simulink
- Être capable de dimensionner les correcteurs présents dans les asservissements des chaînes de production d'énergie

Connaissances préalables recommandées :

- Electronique de puissance
- Modélisation des machines synchrones et asynchrones

Contenu de la matière :

- **Chapitre1** : Rappels sur la modélisation des machines à courant alternatif (asynchrones et synchrones) dans les repères de Clark, Concordia et Park.
- **Chapitre2** : Variateurs de vitesse basés sur des machines asynchrones et synchrones.
- **Chapitre 3**: Commande vectorielle des machines synchrones à aimants permanents
- **Chapitre 4** : Contrôle direct du couple des moteurs asynchrones (DTC)
- **Chapitre 5** : Applications aux systèmes à ER (commandes vectorielle et DTC d'une éolienne basée sur une machine asynchrone, ...)

Mode d'évaluation :

Contrôle continu : 40%, Examen : 60%.

Références bibliographiques:

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Teaching objectives:

- Know the different electrical systems of electric actuators (motor + mechanical loads and static converters)
- Be able to establish a simulation model of an electrical system including motor, power electronics and control
- Be able to simulate a model in the Matlab/Simulink environment
- Be able to size the correctors present in the controls of energy production chains

Recommended prior knowledge:

- Power electronics
 - Modeling of synchronous and asynchronous machines
-
- **Chapter 1** : Reminders on a modeling of alternating current machines (asynchronous and synchronous) in the Clark, Concordia and Park benchmarks.
 - **Chapter 2** : Variable speed drives based on asynchronous and synchronous machines.
 - **Chapter 3**: Vector control of permanent magnet synchronous machines
 - **Chapter 4** : Direct torque control of asynchronous motors (DTC)
 - **Chapter 5** :Applications to RE systems (vector and DTC controls of a wind turbine based on an asynchronous machine, etc.)

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1.Introduction to Modeling of Alternating Current (AC) Machines

The modeling of alternating current (AC) machines is a fundamental aspect of electrical engineering that enables engineers and designers to understand, predict, and optimize the performance of these crucial components in electrical systems. AC machines, including induction motors, synchronous machines, and other types, are integral to various applications, from industrial drives and transportation to power generation and household appliances.

What is AC Machine Modeling?

1.2. Modeling of AC machines :

refers to the development of mathematical or computational representations that describe the behavior of these machines under different operating conditions. These models help in analyzing how AC machines perform, how they interact with their environment, and how they can be controlled efficiently.

1.2.1.Key Objectives of AC Machine Modeling

Performance Analysis: Models provide insights into the operational characteristics of AC machines, such as speed, torque, efficiency, and power factor. This information is essential for designing and optimizing machinery and systems.

Control System Design: Accurate models are crucial for designing control systems, including speed control, torque control, and other advanced control strategies. They allow for the development of algorithms that can regulate machine performance precisely.

System Integration: Models help in integrating AC machines with other system components, such as power converters and mechanical systems, ensuring compatibility and optimal performance.

Predictive Maintenance: Modeling enables the prediction of potential issues and the scheduling of maintenance before problems arise, thus reducing downtime and extending the lifespan of the machines.

Energy Efficiency: By simulating different operating scenarios, engineers can optimize the energy consumption of AC machines, leading to cost savings and improved efficiency.

1.2.2.Types of AC Machine Models

Equivalent Circuit Models: These models represent the AC machine's electrical behavior using an equivalent circuit. The circuit includes components such as resistances, reactances, and mutual inductances that help in analyzing steady-state conditions and performance.

Induction Motor Equivalent Circuit: Includes stator and rotor resistances, stator and rotor reactances, and mutual inductance.

Synchronous Machine Equivalent Circuit: Includes stator windings, rotor windings, and excitation systems.

Dynamic Models: These models describe the behavior of AC machines under transient conditions, such as startup, sudden load changes, or fault conditions. They involve differential equations that capture the time-dependent responses of the machine.

Park's Transformation: A common technique to simplify the analysis of dynamic models by converting three-phase systems into a two-axis (d-q) coordinate system.

Vector Control Models: These models, also known as field-oriented control, decouple the control of torque and flux in AC machines. This allows for precise control of machine performance, particularly useful in applications requiring variable speed and torque.

Finite Element Analysis (FEA) Models: These models provide detailed insights into the magnetic fields, thermal effects, and structural characteristics of AC machines. FEA is used for in-depth analysis and optimization of machine design.

1.2.3.Key Parameters in AC Machine Modeling

Stator and Rotor Resistances (R_s , R_r): Electrical resistances of the stator and rotor windings.

Stator and Rotor Reactances (X_s , X_r): Inductive reactances associated with the stator and rotor.

Mutual Inductance (M): The inductive coupling between the stator and rotor windings.

Magnetizing Reactance (X_m): Reactance representing the magnetic field created in the machine.

Inertia (J): The rotational inertia of the rotor, affecting its dynamics.

Friction and Damping: Mechanical losses due to friction and other damping effects.

1.2.4.Applications of AC Machine Modeling

Design and Optimization: Helps in selecting machine sizes and specifications to meet specific performance criteria and optimize designs.

Control System Development: Essential for creating effective control strategies and algorithms.

Energy Management: Allows for the optimization of energy consumption and efficiency in various applications.

Educational Purposes: Provides a foundation for learning and understanding the principles of AC machine operation and control.

The modeling of AC machines is a critical process that provides a detailed understanding of their performance and behavior. By developing accurate models, engineers can effectively analyze, control, and optimize AC machines for a wide range of applications. These models are indispensable tools for ensuring the reliable and efficient operation of AC machines in various electrical systems and industrial processes.

1.3.Description of induction motor

An induction motor is presented (Figure.1) in the form of a casing (2) surrounding the magnetic, ferromagnetic, stator circuit which accommodates in slots the polyphase stator winding (generally three-phase) wound in wire insulated copper (1). Inside this magnetic circuit, which appears as a hollow cylinder, separated by an air gap, rotates the rotor magnetic circuit (3) which accommodates in its notches the bars of the rotor cage, made of cast aluminum or copper, short -circuited at each end by rings made of the same material. The rotor magnetic circuit is crossed by the shaft which rests on bearings mounted in the flanges (5), (6) fixed to the casing [2].

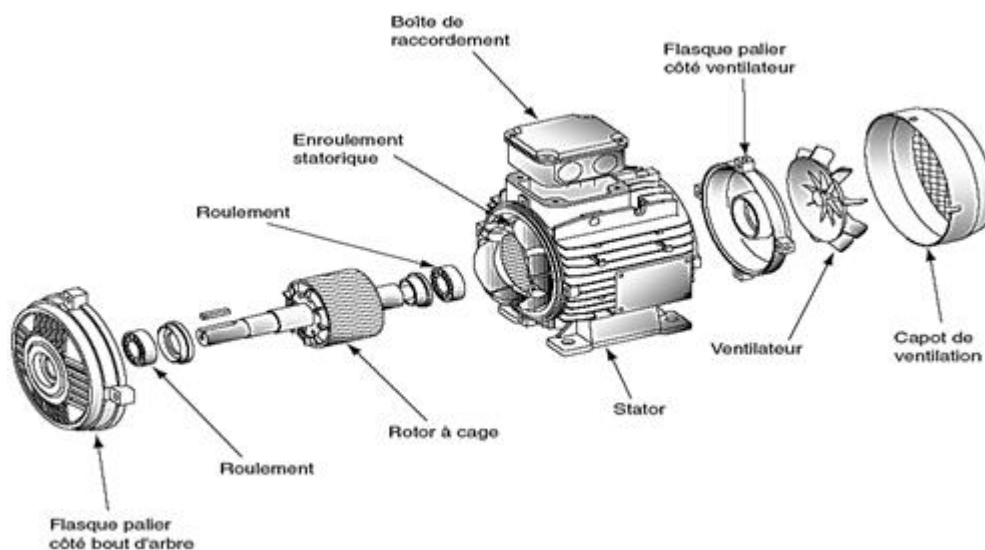


Figure (1.1) Leroy-Somer induction motor

1.3.1 Principle of operation

The pulsating stator currents create a rotating field at the so-called Synchronism speed N_s which has the same frequency as the stator current. This field passes through the rotor winding and induces electromotive forces "e.m.f.". These e.m.f.s produce currents because the rotor winding is short-circuited. The action of the currents on the rotating field which induced them (according to Lenz's law the induced currents oppose the cause which gave rise to them) creates the torque.

This motor is often called an induction machine.

If the rotor rotates at the synchronous speed N_s , i.e. as quickly as the rotating field, the field through each rotor coil is constant. The induced e.m.f. no longer exists at the rotor, therefore no more current or torque. It is therefore necessary that N_r rotor rotation speed differs from N_s synchronism speed for there to be a torque, hence the name asynchronous machine. The operation of the engine is characterized by the relative difference between these two speeds; this is the slip g defined by:

$$g = \frac{\Omega_s - \Omega_r}{\Omega_s}$$

1.3.1. Advantages of the induction motor

The induction motor is the most common motor in industry: it is robust, reliable, economical. It is also appreciated for its very good standardization. With its assumptions, if we consider that the asynchronous motor has three coils in the stator and three coils in the rotor (Figure I.01).

The three types of equations reflecting the behavior of the engine are:

- Electrical equations.
- Magnetic equations.
- Mechanical equations.

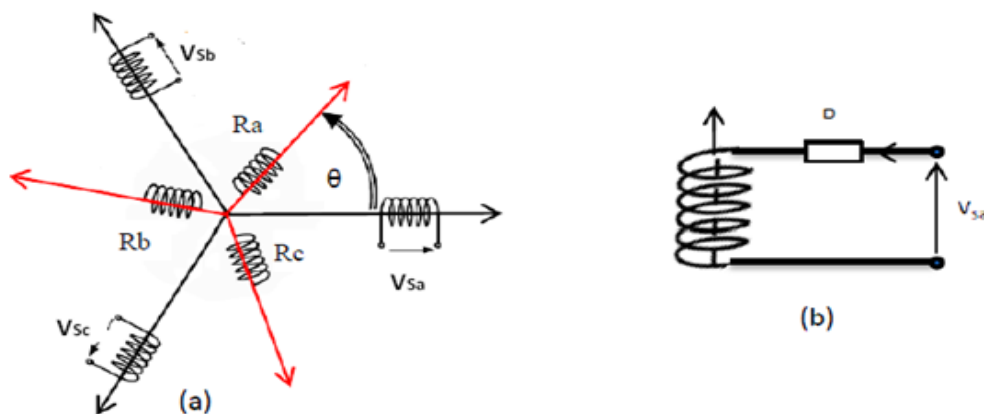


Figure (1.2) Schematic representation of a three-phase induction machine.

(a): diagram of the windings of the three-phase asynchronous machine in the electrical space,

(b): Stator phase with electromotive force

1.3.2. Modeling in the three-phase frame of reference

The general model of the asynchronous machine is obtained by considering the following simplifying hypotheses [4], [3], [5]:

- Constant and uniform air gap around the perimeter of the reinforcements and neglected notch effect,
- Sinusoidal spatial distribution of induction fields in the air gap,
- Linear magnetic characteristic (neglected saturation) and constant permeability,
- Effect of temperature, skin effect, hysteresis and eddy current are negligible
- Negligible ferromagnetic losses.

1.3.3. The consequences of these hypotheses are:

- The additivity of flows,
 - The constancy of the own inductances,
 - The law of sinusoidal variation of mutual inductances between the windings
- stator and rotor as a function of the angle of their magnetic axes.

The induction motor is characterized:

- by the presence of a single polyphase winding powered by a source external to the stator
- by the presence of a massive "winding" short-circuited to the rotor.

1.3.4. Electrical equation

By applying Faraday's law to one of the six stator or rotor windings (Fig 02), the mesh law for the first stator winding is expressed by the relation:

$$V_{sa} = R i_{sa} + \frac{d\varphi_{sa}}{dt} \quad (1)$$

deduced for all the stator phases (V_{sa} , V_{sb} , V_{sc}) and rotor phases (R_a , R_b , R_c) respectively expressed in the three-phase stator and rotor references:

$$\begin{aligned} V_{sa} &= R_s i_{sa} + \frac{d\varphi_{sa}}{dt} \\ V_{sb} &= R_s i_{sb} + \frac{d\varphi_{sb}}{dt}. \end{aligned} \quad (2)$$

$$V_{sc} = R_s i_{sc} + \frac{d\varphi_{sc}}{dt}.$$

These equations (I.2) can be rewritten by introducing the matrix notion (here quantities in square brackets), which results in:

$$[V_{sabc}] = [R_s][i_{sabc}] + \frac{d}{dt}[\varphi_{sabc}] \quad (3)$$

• for the rotor

$$\begin{aligned} V_{ra} &= R_r i_{ra} + \frac{d\varphi_{ra}}{dt} \\ V_{rb} &= R_r i_{rb} + \frac{d\varphi_{rb}}{dt}. \end{aligned} \quad (4)$$

$$V_{rc} = R_r i_{rc} + \frac{d\varphi_{rc}}{dt}.$$

Or in matrix form:

$$[V_{rabc}] = [R_r][i_{rabc}] + \frac{d}{dt}[\varphi_{rabc}] \quad (5)$$

1.3.5. Magnetic equation

The mutual inductances between phases of the stator and the rotor determined as follows:
The sinusoidal distribution of the flows makes it possible to write

$$M_{s1} = M \cos p\theta \quad (6)$$

With:

$$p\theta = w_r t + p\theta_0 \quad (7)$$

θ : geometric angle made by the axes of phases 1 of the stator and rotor at time t.

p: number of pole pairs

The phase shift between a stator and rotor phase is expressed by the relation:

$$\theta + (r - 1)\frac{2\pi}{3p} - (s - 1)\frac{2\pi}{3p} = \theta + (r - s)\frac{2\pi}{3p} \quad \text{with} \quad s = 1,2,3 \text{ et } r=1,2,3$$

hence the general expression of M_{sr}

$$M_{sr} = M \cos \left(p\theta + (r - s)\frac{2\pi}{3} \right) \quad (8)$$

So the instantaneous values of mutual inductance between the stator and rotor are:

$$M_{11} = M \cos(p\theta) \quad M_{12} = M \cos(p\theta + 2\pi/3) \quad M_{13} = M \cos(p\theta + 4\pi/3).$$

$$M_{21} = M \cos(p\theta - 2\pi/3) \quad M_{22} = M \cos(p\theta) \quad M_{23} = M \cos(p\theta + 2\pi/3).$$

$$M_{31} = M \cos(p\theta - 4\pi/3) \quad M_{32} = M \cos(p\theta - 2\pi/3) \quad M_{33} = M \cos(p\theta).$$

Where in matrix form:

$$[M_{sr}] = [M_{rs}]^t = \begin{bmatrix} M_{11} & M_{12} & M_{13} \\ M_{21} & M_{22} & M_{23} \\ M_{31} & M_{32} & M_{33} \end{bmatrix} \quad (9)$$

The equation of the total magnetic flux of a stator winding traversed by a current i_{sa} located in a field produced by a rotor (induced) current ($i_{ra}; i_{rb}; i_{rc}$) and a field by two stator phases (i_{sb}, i_{sc}) therefore the additivity of the flows gives the following writing:

$$\Phi_{sa} = l_{sa} i_{sa} + M_s i_{sb} + M_s i_{sc} + M_{11} i_{ra} + M_{12} i_{rb} + M_{13} i_{rc} \quad (10)$$

For all the stator fluxes (V_{sa}, V_{sb}, V_{sc}), expressed in the three-phase reference frame (figure I.1), are:

$$\varphi_{sa} = l_{sa} i_{sa} + M_s i_{sb} + M_s i_{sc} + M_{11} i_{ra} + M_{12} i_{rb} + M_{13} i_{rc}$$

$$\varphi_{sb} = l_{sa}i_{sb} + M_s i_{sc} + M_s i_{sa} + M_{21}i_{ra} + M_{22}i_{rb} + M_{23}i_{rc} \quad (11)$$

$$\varphi_{sc} = l_{sa}i_{sc} + M_s i_{sa} + M_s i_{sb} + M_{31}i_{ra} + M_{32}i_{rb} + M_{33}i_{rc}$$

and for the rotor:

$$\varphi_{ra} = l_{ra}i_{ra} + M_r i_{rb} + M_r i_{rc} + M_{11}i_{sa} + M_{21}i_{sb} + M_{31}i_{sc}$$

$$\varphi_{rb} = l_{ra}i_{rb} + M_r i_{rc} + M_r i_{ra} + M_{12}i_{sa} + M_{22}i_{sb} + M_{32}i_{sc} \quad (12)$$

$$\varphi_{rc} = l_{ra}i_{rc} + M_r i_{ra} + M_r i_{rb} + M_{13}i_{sa} + M_{23}i_{sb} + M_{33}i_{sc}$$

We obtain the inductance matrix $[L(\theta)]$ which establishes the relation between the fluxes and the currents of which half of the coefficients depend on time, via θ

Either:

$$\begin{bmatrix} \varphi_{sa} \\ \varphi_{sb} \\ \varphi_{sc} \\ \varphi_{ra} \\ \varphi_{rb} \\ \varphi_{rc} \end{bmatrix} = \begin{bmatrix} l_s & M_s & M_s & M_{11} & M_{12} & M_{13} \\ M_s & l_s & M_s & M_{21} & M_{22} & M_{23} \\ M_s & M_s & l_s & M_{31} & M_{32} & M_{33} \\ M_{11} & M_{21} & M_{31} & l_r & M_r & M_r \\ M_{12} & M_{22} & M_{32} & M_r & l_r & M_r \\ M_{13} & M_{23} & M_{33} & M_r & M_r & l_r \end{bmatrix} \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \\ i_{ra} \\ i_{rb} \\ i_{rc} \end{bmatrix} \quad (13)$$

The real flux matrix reveals four inductance sub-matrices

$$\begin{bmatrix} \varphi_{sabs} \\ \varphi_{rabs} \end{bmatrix} = \begin{bmatrix} [L_s] & [M_{sr}] \\ [M_{rs}] & [L_r] \end{bmatrix} \begin{bmatrix} i_{sabs} \\ i_{rabs} \end{bmatrix} \quad (14)$$

with :

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

$$[L_r] = \begin{bmatrix} l_r & M_r & M_r \\ M_r & l_r & M_r \\ M_r & M_r & l_r \end{bmatrix} \quad (16)$$

Finally, we obtain the system of general equations usable for the study of all regimes, balanced or unbalanced, permanent or transitory [3],[4]:

$$\begin{aligned}
 V_{sa} &= R_s i_{sa} + l_s \frac{di_{sa}}{dt} + M_s \frac{di_{sb}}{dt} + M_s \frac{di_{sc}}{dt} + \frac{d}{dt} (M_{11} i_{ra} + M_{12} i_{rb} + M_{13} i_{rc}) \\
 V_{sb} &= R_s i_{sb} + l_s \frac{di_{sb}}{dt} + M_s \frac{di_{sc}}{dt} + M_s \frac{di_{sa}}{dt} + \frac{d}{dt} (M_{21} i_{ra} + M_{22} i_{rb} + M_{23} i_{rc}) \\
 V_{sc} &= R_s i_{sc} + l_s \frac{di_{sc}}{dt} + M_s \frac{di_{sa}}{dt} + M_s \frac{di_{sb}}{dt} + \frac{d}{dt} (M_{31} i_{ra} + M_{32} i_{rb} + M_{33} i_{rc}) \quad (17). \\
 V_{ra} &= R_r i_{ra} + l_r \frac{di_{ra}}{dt} + M_r \frac{di_{rb}}{dt} + M_r \frac{di_{rc}}{dt} + \frac{d}{dt} (M_{11} i_{ra} + M_{12} i_{rb} + M_{13} i_{rc}) \\
 V_{rb} &= R_r i_{rb} + l_r \frac{di_{rb}}{dt} + M_r \frac{di_{rc}}{dt} + M_r \frac{di_{ra}}{dt} + \frac{d}{dt} (M_{21} i_{ra} + M_{22} i_{rb} + M_{23} i_{rc}) \\
 V_{rc} &= R_r i_{rc} + l_r \frac{di_{rc}}{dt} + M_r \frac{di_{ra}}{dt} + M_r \frac{di_{rb}}{dt} + \frac{d}{dt} (M_{31} i_{ra} + M_{32} i_{rb} + M_{33} i_{rc})
 \end{aligned}$$

because the rotor circuits are short-circuited. The dynamic model of the three-phase asynchronous machine can be described in matrix form by:

$$\begin{aligned}
 [V_{sabc}] &= [R_s][i_{sabc}] + \frac{d}{dt} ([L_s][i_{sabc}] + [M_{sr}][i_{rabc}]) \quad (18) \\
 [V_{rabc}] &= [R_r][i_{rabc}] + \frac{d}{dt} ([L_r][i_{rabc}] + [M_{rs}][i_{sabc}])
 \end{aligned}$$

The matrix of mutual inductances $[M_{sr}]$ varying as a function of the electrical angle of the rotor position θ , we simplify these equations by applying the de Park transformation [3].

1.4. Park Transformation

The Park transformation proposed by R. H. Park in 1920 [6], widely used for the modeling of alternating current machines, corresponds to a projection of three-phase quantities on a rotating two-phase reference frame, the aim being to eliminate the position in the mutual inductance matrices. To do this, it is enough to make a transition from the three-phase system ABC to the two-phase system d-q by the Park transformation.

This method is broken down into two steps:

Three-phase-two-phase transformation (in a fixed frame $\alpha\beta$) (Concordia)

Transformation Fixed mark-Rotating mark d-q (Park) The models presented below take into account the simplifying assumptions previous

1.4.1. Three-phase-two-phase transformation

We consider a balanced three-phase sinusoidal system where (x_a, x_b, x_c, x_0) represent physical quantities (current, voltages or magnetic flux).

$$x_a(t) = A \cos(\omega t + \phi)$$

$$x_b(t) = A \cos\left(\omega t + \phi - \frac{2\pi}{3}\right) \quad (19)$$

$$x_c(t) = A \cos\left(\omega t + \phi - \frac{4\pi}{3}\right)$$

Such a system (I.19) is said to be balanced. In this case, the homopolar component $(x_0 = x_a + x_b + x_c)$ is zero at every moment

This allows us to simplify the equations of an alternating current machine by going from a three-phase representation (x_a, x_b, x_c) to a two-phase representation (x_α, x_β, x_0) and vice versa using the C23 transformation matrix of CONCORDIA.

$$\begin{pmatrix} x_\alpha \\ x_\beta \\ x_0 \end{pmatrix} = C_3^T \begin{pmatrix} x_a \\ x_b \\ x_c \end{pmatrix} \quad (20)$$

$$C_{23} = \sqrt{\frac{2}{3}} \begin{bmatrix} \frac{1}{\sqrt{2}} & 1 & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \quad (21)$$

Since $x_0 = 0$, the three-phase to two-phase transformation become

$$\begin{pmatrix} x_\alpha \\ x_\beta \end{pmatrix} = C_{32}^T \begin{pmatrix} x_a \\ x_b \\ x_c \end{pmatrix} \quad (22)$$

$$C_{32} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 1 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \end{bmatrix} \quad (23)$$

The inverse transformation is written directly $x_{abc} = C_{32} x_{\alpha\beta}$, C_{32} is an orthogonal matrix

1.4.2. Transformation Fixed mark-Rotating mark:

To go from a fixed two-phase reference $\alpha\beta$ (sinusoidal quantities) to a reference rotating d,q which is offset by an angle ρ relative to the fixed reference frame, we use the rotation matrix $P(\rho)$ or Park matrix:

$$\begin{pmatrix} x_d \\ x_q \\ x_0 \end{pmatrix} = P(-\rho) \begin{pmatrix} x_\alpha \\ x_\beta \\ x_0 \end{pmatrix} \quad (24)$$

$$P(\rho) = \begin{pmatrix} \cos \rho & -\sin \rho & 0 \\ \sin \rho & \cos \rho & 0 \\ 0 & 0 & 1 \end{pmatrix} \quad (25)$$

For the induction motor of the stator quantities $\rho = \theta_s$ and for the rotor quantities. The inverse transformation is easily obtained

$$x_{\alpha\beta} = P(\rho) x_{dq0}; \text{ car } P(-\rho) = P(\rho)^{-1} P(\rho)^T.$$

So the product of the Park and inverse Park matrices gives the identity matrix $[I]$:

$$[P][P]^{-1} = [I] \quad (26)$$

Another important property used in the development of differential models in

Park landmark:

$$[P] \frac{d[P]^{-1}}{dt} = [P] \frac{d[P]^{-1}}{d\rho} \frac{d\rho}{dt} = \frac{d\rho}{dt} \begin{pmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} \quad (27)$$

In summary:

- Transition from the rotating mark (dq) to the fixed two-phase mark ($\alpha\beta$):

$$\begin{pmatrix} x_d \\ x_q \\ x_0 \end{pmatrix} = P(\rho)^T \begin{pmatrix} x_\alpha \\ x_\beta \\ x_0 \end{pmatrix}$$

Transition from the rotating mark (dq) to the three-phase fixed mark (a, b, c):

$$\begin{pmatrix} x_d \\ x_q \\ x_0 \end{pmatrix} = P(\rho)^T C_3^T \begin{pmatrix} x_\alpha \\ x_\beta \\ x_0 \end{pmatrix} \text{ where}$$

$$P(p)^T C_3^T = [P] = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\rho) & \cos(\rho - \frac{2\pi}{3}) & \cos(\rho + \frac{2\pi}{3}) \\ -\sin(\rho) & -\sin(\rho - \frac{2\pi}{3}) & -\sin(\rho + \frac{2\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (28)$$

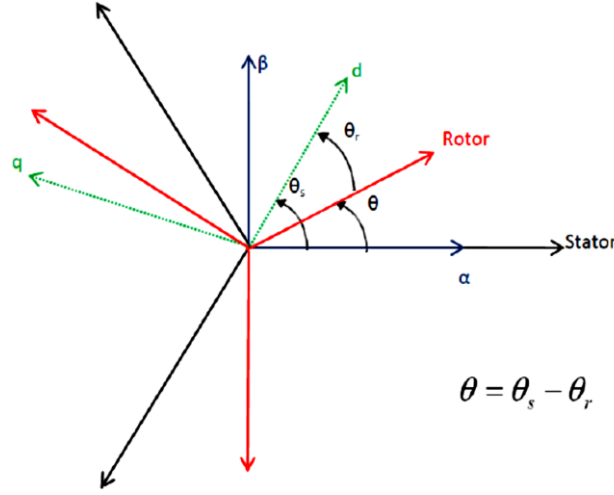


Figure (1.3) Position of the reference frames.

These transformations are all orthogonal; they therefore retain the power instantaneous in the equivalent windings [3].

1.4.3. Model of the induction motor in the d-q plane

1.4.3.1. Electrical equations of a three-phase winding in the d-q plane

The electrical equations of the induction motor in the two-phase system, obtained by applying the Park transformation to equations (15), (17), (18), and (19), are data as follows:

- **The stator:**

The linear transformation $[P]$ is applied to equation (3):

$$[P(\theta_s)]^{-1} [V_{sdqo}] = [R_s] [P(\theta_s)]^{-1} + \frac{d}{dt} [P(\theta_s)]^{-1} [\varphi_{sdqo}] \quad (29)$$

Multiplying the above equation on the left by $[P(\theta_s)]$:

$$[V_{sdqo}] = [R_s][i_{sdqo}] + [P(\theta_s)] \frac{d}{dt} [P(\theta_s)]^{-1} [\varphi_{sdqo}]. \quad (30)$$

On the other hand we have:

$$\frac{d}{dt} [P(\theta_s)]^{-1} [\varphi_{sdqo}] = \frac{d}{dt} \left(([P(\theta_s)]^{-1} [\varphi_{sdqo}]) + [P(\theta_s)]^{-1} \frac{d}{dt} ([\varphi_{sdqo}]) \right) \quad (31)$$

$$[V_{sdqo}] = [R_s][i_{sdqo}] + [P(\theta_s)] \frac{d}{dt} [P(\theta_s)]^{-1} [\varphi_{sdqo}] + \frac{d}{dt} ([\varphi_{sdqo}]) \quad (32)$$

We replace $P(\theta_s)$ by equation (I.25) in equation (I.32) we obtain the model

dynamic electrical for the equivalent two-phase stator winding:

$$\begin{bmatrix} V_{sd} \\ V_{rq} \\ V_{ro} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{so} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \varphi_{sd} \\ \varphi_{rq} \\ \varphi_{ro} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_s \\ \omega_s & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} \varphi_{sd} \\ \varphi_{rq} \end{bmatrix} \quad (33)$$

When the sums of the components (a, b, c) are zero, the equation v_{ro} is zero

therefore becomes useless. So the system (33) becomes:

$$\begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} = \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_s \\ \omega_s & 0 \end{bmatrix} \begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \end{bmatrix} \quad (34)$$

$$\text{With : } \omega_s = \frac{d\theta_s}{dt} \quad \text{et} \quad \rho = \theta_s.$$

For the rotor: Likewise, by applying the Park transformation on the rotor equations, we obtains the dynamic electrical model for the equivalent two-phase rotor winding:

$$\begin{bmatrix} V_{rd} \\ V_{rq} \end{bmatrix} = \begin{bmatrix} R_r & 0 \\ 0 & R_r \end{bmatrix} \begin{bmatrix} i_{rd} \\ i_{rq} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \varphi_{rd} \\ \varphi_{rq} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_r \\ \omega_r & 0 \end{bmatrix} \begin{bmatrix} \varphi_{rd} \\ \varphi_{rq} \end{bmatrix} \quad (35)$$

With :

$$\omega_r = \frac{d\theta_r}{dt} \quad \text{et} \quad \rho = \frac{\theta_r}{\theta_s} = \theta_s - \theta$$

1.4.3.2. Magnetic equations of a three-phase winding in the d-q plane

Following the same reasoning, the application of the Park transformation on the system (17) allows us to arrive at the matrix relationship between the flux vectors and the currents in the dq mark:

$$\begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \\ \varphi_{so} \\ \varphi_{rd} \\ \varphi_{rq} \\ \varphi_{ro} \end{bmatrix} = \begin{bmatrix} l_s - M_s & 0 & 0 & 3M_{sr}/2 & 0 & 0 \\ 0 & l_s - M_s & 3M_{sr}/2 & 0 & 3M_{sr}/2 & 0 \\ 0 & 0 & l_s - 2M_s & 0 & 0 & 0 \\ 3M_{sr}/2 & 0 & 0 & l_r - M_r & 0 & 0 \\ 0 & 3M_{sr}/2 & 0 & 0 & l_r - M_r & 0 \\ 0 & 0 & 0 & 0 & 0 & l_r - 2M_r \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{so} \\ i_{rd} \\ i_{rq} \\ i_{ro} \end{bmatrix} \quad (36)$$

We see that the number of electromagnetic parameters is reduced to five. These are :

$L_s = l_s - M_s$ Stator cyclic inductance.

$L_r = l_r - M_r$ Rotor cyclic inductance.

$L_{os} = l_s - 2M_s$ The stator homopolar inductance.

$L_{or} = l_r - 2M_r$ Rotor homopolar inductance

Since the system is balanced, we have :

$$\varphi_{so} = \varphi_{ro} = 0 \text{ et } i_{so} = i_{ro} = 0$$

Under these conditions the system (36) rewrite according to three parameters L_s, L_r, M constants comme suite :

$$\begin{bmatrix} \varphi_{sd} \\ \varphi_{sq} \\ \varphi_{rd} \\ \varphi_{rq} \end{bmatrix} = \begin{bmatrix} L_s & 0 & M & 0 \\ 0 & L_s & 0 & M \\ M & 0 & L_r & 0 \\ 0 & M & 0 & L_r \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad (37)$$

So we can represent the machine by the following figure:

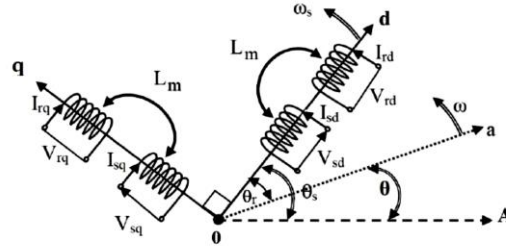


Figure (1.4) Schematic representation of a two-phase induction motor

The relations (34), (35) and (36) make it possible to write the electrical equations of the induction in the axis reference frame (d,q).

$$\begin{bmatrix} V_{sd} \\ V_{sq} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + L_s(d/dt) & -L_s(d\theta_s/dt) & M(d/dt) & -M(d\theta_s/dt) \\ -L_s(d\theta_s/dt) & R_s + L_s(d/dt) & M(d\theta_s/dt) & M(d/dt) \\ M(d/dt) & -M(d\theta_r/dt) & R_r + L_r(d/dt) & -L_r(d\theta_r/dt) \\ M(d\theta_r/dt) & M(d/dt) & L_r(d\theta_r/dt) & R_r + L_r(d/dt) \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad (38)$$

1.4.3.3. Mechanical equation

In the study of the dynamic characteristics of the asynchronous machine, it is necessary to introduce the mechanical parameters (torque, speed, flux) with the electrical parameters (voltage, current, flux, etc.)

The electromagnetic torque of the machine is calculated from a balance of power, hence the instantaneous electrical power p_e supplied to the stator and rotary windings as a function of the axis quantities (d, q) is given as follows:

$$p_e = (V_{sd}i_{sd} + V_{sq}i_{sq}) + (V_{rd}i_{rd} + V_{rq}i_{rq}) \quad (39)$$

It is broken down into three series of terms:

- Power dissipated in joule losses:

$$R_s(i_{sd}^2 + i_{sq}^2) + R_r(i_{rd}^2 + i_{rq}^2) \quad (40)$$

- Power representing the exchanges of electromagnetic energy with the source

$$i_{sd} \frac{d\varphi_{sd}}{dt} + i_{sq} \frac{d\varphi_{sq}}{dt} + i_{rd} \frac{d\varphi_{rd}}{dt} + i_{rq} \frac{d\varphi_{rq}}{dt} \quad (41)$$

Mechanical power P_m

$$P_m = (\varphi_{sd}i_{sq} - \varphi_{sq}i_{sd})\frac{d\theta_s}{dt} + (\varphi_{rd}i_{rq} - \varphi_{rq}i_{rd})\frac{d\theta_r}{dt} \quad (42)$$

Taking into account the flow equations we can therefore write that:

$$P_m = (\varphi_{sd}i_{sq} - \varphi_{sq}i_{sd})\frac{d(\theta_s - \theta_r)}{dt} = (\varphi_{sd}i_{sq} - \varphi_{sq}i_{sd})\omega \quad (43)$$

And on the other hand, the expression of the electromagnetic torque is given by:

$$C_e = \frac{P_m}{\Omega} = \rho \left(\frac{P_m}{\omega} \right) = C_e = \rho(\varphi_{sd}i_{sq} - \varphi_{sq}i_{sd}) \quad (44)$$

Manipulations of equation (I.44) we can deduce several expressions of the couple, at starting from the fundamental equation above, which will nevertheless be the starting points in the search for a strategy for its control. We thus find:

$$C_e = \rho(\varphi_{rd}i_{rq} - \varphi_{rd}i_{rq}) \quad (45)$$

$$C_e = \rho M(i_{sq}i_{rd} - i_{rd}i_{sq}) \quad (46)$$

$$M = \frac{3}{2}M_{sr}. \text{ The cyclic mutual inductance between stator and rotor}$$

1.4.3.4. Equation of motion

The dynamics of the moving part of the machine is expressed by the following equation of motion

$$J \frac{d\Omega}{dt} = C_e - C_r - f_r \Omega \quad (48)$$

J : moment of inertia of the rotating masses

C_r : resistant torque imposed on the machine shaft.

Ω: rotor speed.

C_e : electromagnetic torque.

f_r : coefficient of viscous friction.

f_rΩ : viscous friction torque term.

1.4.3.5. Choice of d-q frame of reference

There are three important choices concerning the orientation of the axis reference (d,q) which depend of the purpose of the application:

Mark (d,q) linked to the stator : $\omega_s = \frac{d\theta_s}{dt} = 0 \text{ et } \omega_r = -\omega$.

This frame of reference is immobile in relation to the stator, used for the study of starting and braking of alternating current machines with connection of resistors.

* Mark (d,q) linked to the rotor $\omega_s = \frac{d\theta_s}{dt} = 0 \text{ et } \omega_r = 0$

This frame of reference is immobile in relation to the rotor, used for the study of transient regimes in asynchronous and synchronous machines.

• Mark (d,q) linked to the rotating field $\omega_s = \omega_e \text{ et } \omega_r = \omega_e - \omega$.

The latter is used to carry out vector control because the adjustment quantities become continuous

La figure I.2 montre que $\theta = \theta_s - \theta_r$, the derivation of this equation gives:

This internal kinetic relation shows that the angular speeds of the axes (d,q) respectively in the stator and rotor frames and rigidly linked to the angular speed of the rotor Ω , which are not independent quantities. Equations (48) and (49) therefore constitute the basic model of the induction motor .

1.5. Modeling in MAS state representation and control

The state representation requires a choice of state variables, inputs and outputs of the system, this choice depends on the objective linked to the control or observation. For control applications, a suitable choice for the state vector is:

$$\mathbf{x} = [i_{sd}, i_{sq}, \varphi_{sd}, \varphi_{sq}, \omega]^T \quad (49)$$

The choice of stator currents is justified by the fact that they are accessible for measurement and the rotor fluxes for the orientation of the reference frame according to the rotor field. The vector of command includes the components of the supply voltage

(V_{sd} and V_{sq}), The model state of the asynchronous machine is that of a multi-variable system not linear of the following form:

$$\mathbf{x}(t) = \mathbf{f}(\mathbf{x}) + \mathbf{g}(\mathbf{x})\mathbf{u}(t) \quad (50)$$

$$\mathbf{y}(t) = \mathbf{h}(t)$$

After some mathematical manipulations of equations (I.34), (I.35) and (I.36), we obtain the four-dimensional state model of the asynchronous machine in a rotating reference frame:

$$\begin{bmatrix} \frac{di_{sd}}{dt} \\ \frac{di_{sq}}{dt} \\ \frac{d\varphi_{rd}}{dt} \\ \frac{d\varphi_{rd}}{dt} \end{bmatrix} = \begin{bmatrix} -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma T_r}\right) & \omega_s & \frac{1}{\sigma T_r} \frac{M}{L_s L_r} & \frac{1}{\sigma T_r} \frac{M}{L_s L_r} \\ -\omega_s & -\left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma T_r}\right) & \frac{-1}{\sigma} \frac{M}{L_s L_r} & \frac{-1}{\sigma} \frac{M}{L_s L_r} \\ 0 & 0 & -\frac{1}{T_r} & \omega_r \\ 0 & \frac{M}{T_r} & -\omega_r & -\frac{1}{T_r} \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} \quad (51)$$

we add the expression for the speed, so the state model becomes a system of five dimensions

$$\frac{d\omega}{dt} = \frac{p^2 M}{J L_r} (\varphi_{rd} i_{rq} - \varphi_{rd} i_{rq}) - \frac{p}{J} C_r - \frac{p_r}{J} \omega \quad (52)$$

With:

σ : dispersion coefficient

T_r : rotor time constant:

$$\sigma = 1 - \left(\frac{M^2}{L_s L_r} \right), \quad T_r = \frac{L_r}{R_r}$$

RECITATION N° 1

Exercise n° 1 :

Consider a system of balanced three-phase voltages of the form

$v_i(t) = v_m \sin(\omega t - \frac{2\pi}{3}(i-1))$, a balanced three-phase linear load connected in star with an isolated neutral.

1°) Calculate the two components of the voltage vector v_a and v_b using the transformation matrix defined by: $x_{ab} = T_{23} x_{23}$ with :

$$T_{23} = k \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}$$

Deduce the value of the constant k so that this transformation is capable of ensuring conservation amplitudes.

2) Calculate the instantaneous power consumed by the load in the three-phase benchmark. Recalculate this power in the reference $\alpha\beta$ taking into account the value of k calculated in the first question. Comment on the result obtained

Exercise n° 2

In order to transform a quantity x_{abc} of the natural three-phase system into another quantity x_{uvo} in a equivalent three-phase system rotating at an arbitrary speed, the following coordinate transformation is used:

$$x_{uvo} = [P(\psi)] x_{abc}$$

with $P(\Psi)$ is the Park matrix given as a function of the transformation angle as follows:

$$[P(\psi)] = k \begin{bmatrix} \cos(\psi) & \cos(\psi - \frac{2\pi}{3}) & \cos(\psi + \frac{2\pi}{3}) \\ -\sin(\psi) & -\sin(\psi - \frac{2\pi}{3}) & -\sin(\psi + \frac{2\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix}$$

1°) Show that to ensure the invariance of the instantaneous power in the two systems of representation, the Park matrix must be orthogonal verifying the following relation:

$$[P(\psi)]^t = [P(\psi)]^{-1}$$

2°) Let us assume that the previous condition is met, calculate the value of k.

Exercise 3

Knowing that the electrical and magnetic equations of the squirrel cage asynchronous machine (MAS) in the natural coordinate system ABC are given by the following equations in matrix form:

Electrical equations :

$$\begin{aligned} [v_{abcs}] &= R_s [i_{abcs}] + \frac{d[\varphi_{abcs}]}{dt} \\ 0 &= R_r [i_{abcr}] + \frac{d[\varphi_{abcr}]}{dt} \end{aligned}$$

Magnetic equations

$$\begin{aligned} [\varphi_{abcs}] &= [M_{ss}] [i_{abcs}] + [M_{sr}(\theta)] [i_{abcr}] \\ [\varphi_{abcr}] &= [M_{rs}(\theta)] [i_{abcs}] + [M_{rr}] [i_{abcr}] \end{aligned}$$

$$[M_{ss}] = \begin{bmatrix} l_s & m_s & m_s \\ m_s & l_s & m_s \\ m_s & m_s & l_s \end{bmatrix}, [M_{rr}] = \begin{bmatrix} l_r & m_r & m_r \\ m_r & l_r & m_r \\ m_r & m_r & l_r \end{bmatrix}, [M_{sr}(\theta)] = [M_{rs}(\theta)]^t = m_{sr} \begin{bmatrix} \cos(\theta) & \cos(\theta + \frac{2\pi}{3}) & \cos(\theta - \frac{2\pi}{3}) \\ \cos(\theta - \frac{2\pi}{3}) & \cos(\theta) & \cos(\theta + \frac{2\pi}{3}) \\ \cos(\theta + \frac{2\pi}{3}) & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta) \end{bmatrix}$$

Chapter 1 : Reminders on a modeling of alternating current machines (asynchronous and synchronous) in the Clark, Concordia and Park benchmarks

1°) Establish the model of the MAS in the reference frame linked to the rotating field (dq) by adopting the transformation of Park defined by the following passage matrix

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

2°) Put the MAS model in the benchmark (dq) in the following form:

$$\dot{x} = f(x) + g u$$

$$x = [i_{ds} \quad i_{qs} \quad \varphi_{dr} \quad \varphi_{qr} \quad \Omega]^T$$

Most modern AC variable speed drives (VSDs) are of modular construction. Some of the technical details of the main components, such as the input rectifier, DC link, output inverter.

Although the main function of the control system for modern PWM-type AC VF converters is to control the semiconductor switches of the PWM inverter, there are a number of other important functions, which need to be controlled. The overall control system can be divided into 4 main areas:

Inverter control system

Speed feedback and control system

Current feedback and control system

External interface, which includes the following:

- Parameter settings by the user
- Operator information and fault diagnostics
- Digital and analog inputs to receive control signals (start, stop, etc)
- Digital and analog outputs to pass on status information (running, faulted, etc)

With the rapid advances in digital electronics over the last decade, modern VSD controlsystems are based on one or more microprocessors. The control system must be designed to achieve the following main objectives:

- High level of reliability
- High inverter performance to ensure that the output current waveform provides sufficient motor torque, at selected speed, with minimum of motor losses
- Inverter losses should be minimized

- It must be possible to integrate the control system into the overall process

control system, with facilities for external control and communications

interface

High tolerance to power supply fluctuations and EMI

The voltage inverter ensures the conversion of direct to alternating energy (DC/AC) This application is very widespread in the world of electrical energy conversion today. The inverter can be used at fixed frequency, for example powering an alternating system from a battery, or at variable frequency (PMI) for dimming. speed of electrical machines The PWM voltage inverter allows voltage waves of variable amplitudes and frequencies to be imposed on the machine from the standard 230/400V, 50Hz network. structure of the static converter which powers the machine is essentially made up of a rectifier bridge (AC/DC) connected to the network, controlled or not [7]. After rectification, the voltage (continuous stage) is filtered by passive components C or LC ,to be finally applied to the inverter.

The inverter which is connected to the machine is made up of three arms made up of switches electronics chosen essentially according to the power and the working frequency, each arm has two complementary power components equipped with a diode mounted in anti-parallel. The freewheeling diodes ensure continuity of current in the MAS once the switches are opened.

Note that a delay time must exist practically between the high and low switches of the same arm in order to avoid short-circuiting the DC source.

The inverter is controlled by the pulse width modulation (PWM) technique, called in English (Pulse Width Modulation PWM). There are several PWM techniques, two of which will be mentioned, the so-called sine-triangle PWM (STPWM), and the vector PWM or (abbreviated space vector PWM) (SVPWM), which has become very popular with manufacturers and researchers in the control of electrical machines.

Before starting the modeling of the inverter, we considered it interesting to provide a description of the static semiconductor switches used in power electronics that currently exist, because the key element of energy conversion is the static switch which will allow, by interrupting or not the transfer of energy between the various elements of the circuit, and to manage the average values of currents and voltages.

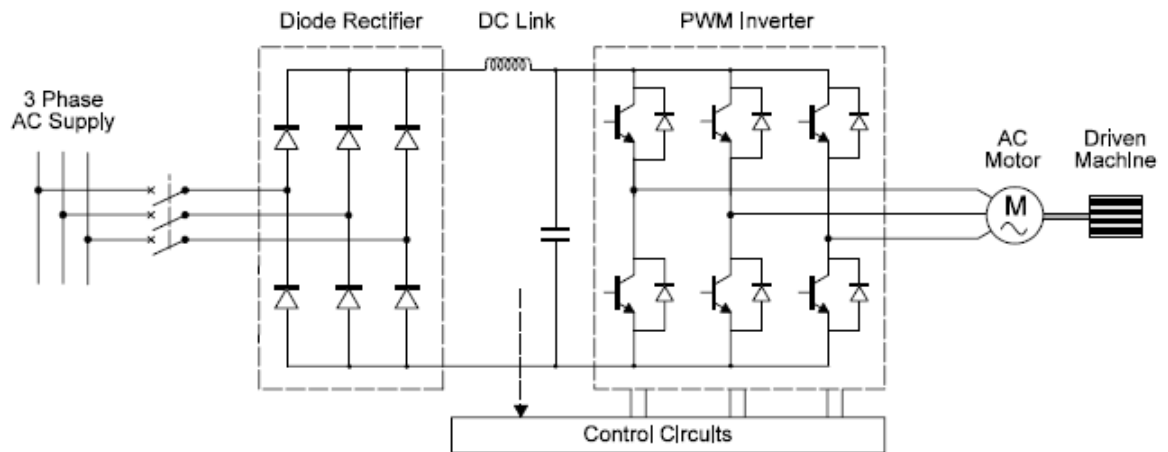


Figure (2.1) Main components of an AC variable speed drive

2.1 The components of power electronics

The power electronics components must respond to one command which exiles the on-state outblouse, this change in conductivity. possible thanks to the special structures of the semiconductors which control the mobile charge carriers, three different types sent defined:

uncontrolled semiconductors: power diodes: PiN, SCHOTTKY, JBS.

semiconductor controls with ignition control: Thyristors, Triacs

semiconductors controlled with firing/blocking control:

Thyristors **GTO** (Gate Turn Off)

Bipolar power transistors BJT (Bipolar Junction Transistor)

MOSFET (Metal Oxide Semiconductor Field Effect Transistor) power transistors

IGBT (Insulated Gate Bipolar Transistor)

GCT et **IGCT** (Integrated Gate Commutated Turn-off Thyristor)

MCT (Mos Controlled Thyristor)

SIT (Static Induction Transistor)

SITH (Static Induction thyristor)

2.1.1 Choice of switches

The following parameters are important for a semiconductor switch designed for power conversion applications

- ☐ Maximum current supported.
- ☐ Maximum blocked voltage.
- ☐ On-state voltage drop and its dependence on temperature.
- ☐ Leakage current in off state.
- ☐ Thermal capacity.
- ☐ Switching transition time during opening and closing.
- ☐ End/dv holding in the blocked state or during opening.
- ☐ Holding in dt/di in the on state or during closing.
- ☐ Ability to withstand excessive current and voltage.
- ☐ Switching losses.
- ☐ Controlled power

2.3. Mathematical model of voltage inverter

The asynchronous machine is often powered by a static converter (voltage inverter). The diagram of the inverter is shown in Figure II.1. It consists of three transistor switching arms. Each arm is made up of two cells, each comprising a diode and a transistor. for the modeling of the voltage inverter we assume for simplicity:

- Instant switching of switches.
- The voltage drop across the switches is negligible.
- Balanced and star-coupled load with isolated neutral.

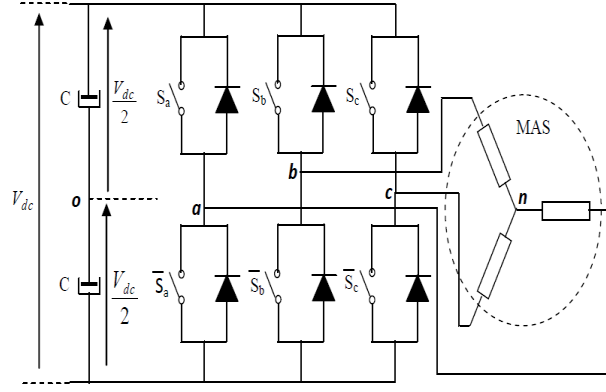


Figure (2.2) the voltage inverter associated with the MAS

The voltages of phases a, b, c relative to the midpoint “o” of the source are given as follows:

- For phase a:

$$\begin{cases} V_{ao} = +\frac{V_{dc}}{2} & ; \text{Si } K_1 \text{ est fermé ; } S_a = 1 \\ V_{ao} = -\frac{V_{dc}}{2} & ; \text{Si } K_4 \text{ est fermé ; } S_a = 0 \end{cases} \quad (1)$$

- For phase b:

$$\begin{cases} V_{bo} = \frac{V_{dc}}{2} & ; \text{Si } K_2 \text{ est fermé ; } S_b = 1 \\ V_{bo} = -\frac{V_{dc}}{2} & ; \text{Si } K_5 \text{ est fermé ; } S_b = 0 \end{cases} \quad (2)$$

- For phase c:

$$\begin{cases} V_{co} = \frac{V_{dc}}{2} & ; \text{Si } K_3 \text{ est fermé ; } S_c = 1 \\ V_{co} = -\frac{V_{dc}}{2} & ; \text{Si } K_6 \text{ est fermé ; } S_c = 0 \end{cases} \quad (3)$$

The compound voltages are:

$$\begin{cases} V_{ab} = V_{ao} - V_{bo} \\ V_{bc} = V_{bo} - V_{co} \\ V_{ca} = V_{co} - V_{ao} \end{cases} \quad (4)$$

For simple voltages, we can write the following relations:

$$\begin{cases} V_{ao} = V_{an} - V_{no} \\ V_{bo} = V_{bn} - V_{no} \\ V_{co} = V_{cn} - V_{no} \end{cases} \quad (5)$$

With :

V_{an}, V_{bn}, V_{cn} : the simple voltages, and V_{no} is the fictitious voltage between the neutral of the MAS and the fictitious point “o”. The system V_{an}, V_{bn} et V_{cn} being balanced, therefore:

$$V_{an} + V_{bn} + V_{cn} = 0 \quad (6)$$

so :

$$V_{no} = \frac{1}{3}(V_{an} + V_{bn} + V_{cn}) \quad (7)$$

By replacing (II-07) in (II-04) we obtain (II-08):

$$\begin{cases} V_{an} = \frac{1}{3}(2V_{ao} - V_{bo} - V_{co}) \\ V_{bn} = \frac{1}{3}(-V_{ao} + 2V_{bo} - V_{co}) \\ V_{cn} = \frac{1}{3}(-V_{ao} - V_{bo} + 2V_{co}) \end{cases} \quad (8)$$

We can also write these tensions in the following matrix form:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} V_{ao} \\ V_{bo} \\ V_{co} \end{bmatrix} \quad (9)$$

$$\text{Soit } [V_{jn}] = [T][V_{jo}] \quad (10)$$

Therefore, our voltage inverter can be modeled by a matrix [T] ensuring the DC-AC transition

$$[T] = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \quad (11)$$

By substituting the values of V_{jo} of each phase a, b, and c in the system (9) we obtain the voltages across the load as a function of the Boolean values of the states of the switches:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (12)$$

2.4. Inverter control strategy

The pulse width modulation (PWM) technique makes it possible to control the output voltage of the inverter in amplitude and frequency from the switch control signals [7]. To obtain more efficient control, the PWM must be precise and fast. there are different Pulse Width Modulation structures making it possible to manage the three supply voltages of the asynchronous machine from the precalculated PWM reference voltage, sine-triangle PWM, vector PWM

The PWM control techniques will be developed in this memory, namely the sine – triangle PWM control.

$$\begin{cases} V_{bo} = \frac{V_{dc}}{2} ; Si K_2 \text{ est fermé} ; S_b = 1 \\ V_{bo} = -\frac{V_{dc}}{2} ; Si K_5 \text{ est fermé} ; S_b = 0 \end{cases} \quad (13)$$

Pour la phase c :

$$\begin{cases} V_{co} = \frac{V_{dc}}{2} ; Si K_3 \text{ est fermé} ; S_c = 1 \\ V_{co} = -\frac{V_{dc}}{2} ; Si K_6 \text{ est fermé} ; S_c = 0 \end{cases} \quad (14)$$

The compound voltages are:

$$\begin{cases} V_{ab} = V_{ao} - V_{bo} \\ V_{bc} = V_{bo} - V_{co} \\ V_{ca} = V_{co} - V_{ao} \end{cases} \quad (15)$$

For simple voltages, we can write the following relations:

$$\begin{cases} V_{ao} = V_{an} - V_{no} \\ V_{bo} = V_{bn} - V_{no} \\ V_{co} = V_{cn} - V_{no} \end{cases} \quad (16)$$

With :

V_{an}, V_{bn}, V_{cn} : the simple voltages, and V_{no} is the fictitious voltage between the neutral of the MAS and the fictitious point “o”. The system V_{an}, V_{bn} et V_{cn} being balanced, therefore:

$$V_{an} + V_{bn} + V_{cn} = 0 \quad (17)$$

so :

$$V_{no} = \frac{1}{3} (V_{an} + V_{bn} + V_{cn}) \quad (18)$$

By replacing (7) in (8) we obtain (19):

$$\begin{cases} V_{an} = \frac{1}{3} (2V_{ao} - V_{bo} - V_{co}) \\ V_{bn} = \frac{1}{3} (-V_{ao} + 2V_{bo} - V_{co}) \\ V_{cn} = \frac{1}{3} (-V_{ao} - V_{bo} + 2V_{co}) \end{cases} \quad (19)$$

We can also write these voltages in the following matrix form:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} V_{ao} \\ V_{bo} \\ V_{co} \end{bmatrix} \quad (20)$$

$$\text{Soit } [V_{jn}] = [T][V_{jo}] \quad (21)$$

Therefore, our voltage inverter can be modeled by a matrix [T] ensuring the DC-AC transition.

$$[T] = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \quad (22)$$

By substituting the values of V_{jo} of each phase a, b, and c in the system (2) we obtain the voltages across the load as a function of the Boolean values of the states of the switches:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (23)$$

2.4.1 Inverter control strategy

The pulse width modulation (PWM) technique makes it possible to control the output voltage of the inverter in amplitude and frequency from the switch control signals [8]. To obtain more efficient control, the PWM must be precise and fast. there are different Pulse Width Modulation structures making it possible to manage the three supply voltages of the asynchronous machine from the precalculated PWM reference voltage, sine-triangle PWM, vector PWM

The PWM control techniques will be developed in this memory, namely the sine – triangle PWM control.

2.4.2 Principle of Sine-Triangle Pulse Width modulation:

Sine-triangle M.L.I is carried out by comparing a low-frequency modulating wave (reference voltage) to a high-frequency triangular-shaped carrier wave. The switching times are determined by the points of intersection between the carrier and the modulating [3], The switching frequency of the switches is fixed by the carrier. This technique is characterized by the following two parameters [6]:

The modulation index “m” equal to the ratio of the carrier frequency (f_p) to the reference frequency (f_{ref}) $m = \frac{f_p}{f_{ref}}$

The voltage adjustment coefficient (the modulation rate) “r”, equal to the ratio of the amplitude of the reference voltage (V_{m_ref}) to the peak value of the modulation wave (V_{pm}) $r = \frac{V_{m_ref}}{V_{pm}}$ the fundamental amplitude of the signal obtained are adjustable, The sine-triangle PWM is carried out by comparison of a generally sinusoidal reference wave of bass figure (1.2) shows the principle of M.L.I control with sine triangle modulation.

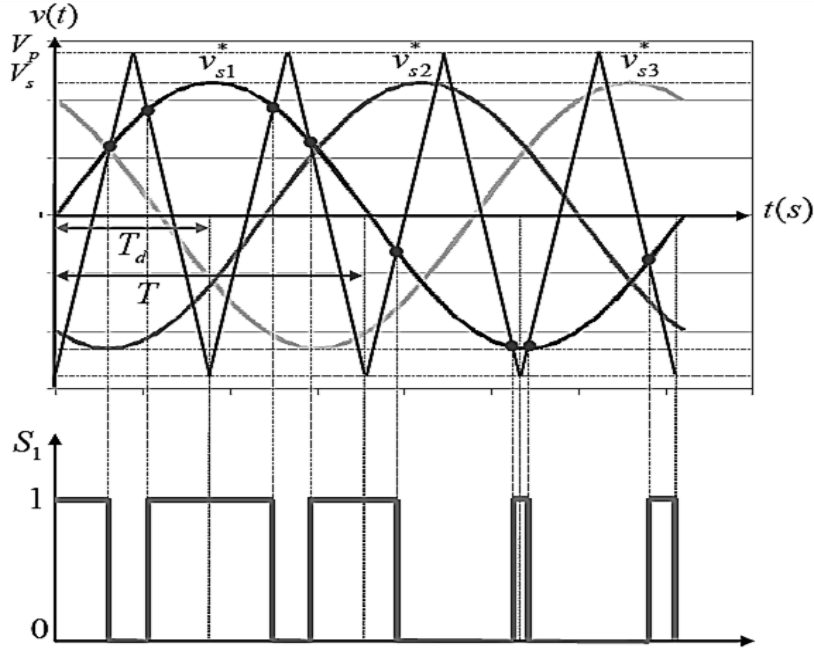


Figure (2.3) the principle of the PWM command.

The carrier is a triangular signal characterized by its frequency f_p and its peak value V_p . We define the equation of the carrier in its period $[0, T_d]$ by:

$$\begin{cases} y_1(t) = V_p \left(-1 + \frac{4t}{T_d} \right) & \text{si } t \in \left[0, \frac{T_d}{4} \right] \\ y_2(t) = V_p \left(+3 - \frac{4t}{T_d} \right) & \text{si } t \in \left[\frac{T_d}{2}, T_d \right] \end{cases} \quad (24)$$

The reference is a sinusoidal signal of amplitude V_r of frequency f_r . In three-phase, the three sinusoidal reference voltages are given by:

$$\begin{cases} V_{a_ref} = \frac{V_{dc}}{2} \sin(\omega t) \\ V_{b_ref} = \frac{V_{dc}}{2} \sin\left(\omega t - \frac{2\pi}{3}\right) \\ V_{c_ref} = \frac{V_{dc}}{2} \sin\left(\omega t + \frac{2\pi}{3}\right) \end{cases} \quad (25)$$

The principle of this strategy can be summarized by the following algorithm:

- If $V_{ref} > V_p$ the upper bridge arm transistor conducts.
- If $V_{ref} < V_p$ the lower transistor of the bridge arm conducts.

RECITATION N° : 2

Exercise 1 :

Consider a three-phase asynchronous machine with a wound rotor connected in a star connection 4150 V, 50 Hz, 6 poles. This machine is controlled by a chopper connected to the rotor via a diode rectifier . There machine drives a load whose torque is proportional to the square of the rotation speed. The report transformation between the stator and the rotor is 0.9. Note that for the nominal speed 979 rpm, the torque is 750 Nm.

1°) If we admit that the machine must operate at a minimum speed of 667 rpm:

a°) Calculate the slip maximum allowable.

b°) Calculate the average value of the voltage rectified.

c°) Calculate the value of the load resistance R.

2°) Using the value of R calculated previously and for a speed of 875 rpm calculate:

a°) Direct current I_d .

b°) The average value of the rectified voltage.

c°) The duty cycle of the chopper.

d°) The efficiency assuming that the iron losses are negligible and the stator Joule losses represent 30% of rotor Joule losses.

e°) The input power factor.

Exercise 2 :

In the assembly of Figure (2), the stator of the machine asynchronous (MAS) to wound rotor is connected to a network three-phase voltage effective compound U_s while his rotor, tension effective compound U_r is connected to a hyposynchronous cascade formed by a three-phase diode bridge, a filter and a three-phase thyristor bridge operating as an assisted inverter. the direct current is assumed to be perfectly smoothed using the smoothing inductance

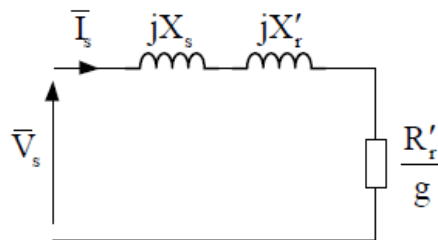
- 1°) What is the main advantage of the variator in figure (1).
- 2°) Trace the shapes of the tensions $u_d(t), u_{d0}(t)$ and calculate their average values.
- 3°) Show that the slip can be expressed as a function of the start delay angle by: $-g k \cos \alpha$, where k is a constant to calculate. Deduce the expression for the rotation speed of the motor depending on α .
- 4°) Give the expression of the electromagnetic torque as a function of α
- 5°) Neglecting mechanical losses, calculate the direct current for a given load torque.

Exercise 3 :

Or a two-pole three-phase motor intended to work with an effective compound voltage of 415V and a frequency of 50Hz.

1°) By neglecting the resistance and the stator iron losses and assuming that the magnetization branch is infinite, the equivalent diagram per phase of the machine simplifies to that of the figure opposite where the parameters of the machine at nominal frequency

$$X_s = 0.75 \, \Omega, \quad R'_r = 0.07 \, \Omega \text{ et } X'_r = 0.7 \, \Omega$$



- a°) Give the expression for the amplitude of the stator current $s I$ absorbed by the machine depending on V_s, X_s, R'_r, X'_r .
 - b°) Give the expression for the electromagnetic torque C_{em} developed by the machine depending on $V_s, X_s, R'_r, X'_r, g, \Omega_r$.
- 2°) To operate the motor at variable speed, it is powered by a three-phase inverter controlled by a scalar control law with a ratio

- a°) Tracer le schéma de la commande scalaire en boucle ouverte en considérant la fréquence f_s comme entrée.

- b°) Calculate the current and the electromagnetic torque of the motor when it operates at a slip of 0.2 with a frequency equal to half the nominal frequency.

- c°) If the supply frequency is increased quickly to its nominal value, calculate the current and torque immediately after this rapid change in frequency. Comment on the results found and propose a solution.

3.1 Introduction to Vector Control of Permanent Magnet Synchronous Machines

Permanent Magnet Synchronous Machines (PMSMs) are increasingly prominent in various applications, ranging from electric vehicles to industrial automation, due to their high efficiency, compact design, and superior performance characteristics. The vector control technique, also known as field-oriented control (FOC), plays a pivotal role in maximizing the performance of these machines by decoupling the torque and flux control. This method allows for precise regulation of the motor's torque and speed, facilitating rapid dynamic response and enhancing overall system efficiency. Vector control operates by transforming the three-phase currents into a two-dimensional coordinate system, aligning with the rotor's magnetic field. This transformation enables independent control of the magnetic flux and the torque-producing current, akin to how one would control a DC motor. The advantages of this approach include improved torque response, reduced current ripple, and enhanced efficiency across a wide range of operating conditions.

As the demand for high-performance electric drives grows, understanding the intricacies of vector control for PMSMs becomes crucial for engineers and researchers. This introduction sets the stage for exploring the principles, methodologies, and applications of vector control, highlighting its significance in advancing modern motor control technologies.

3.2 Principle of rotor flux orientation

The principle of rotor flux orientation is central to the vector control (or field-oriented control) of Permanent Magnet Synchronous Machines (PMSMs). This method focuses on aligning the control of the motor's magnetic fields with the rotor's magnetic field, allowing for precise and efficient motor operation. Here's an overview of how it works:

3.2.1. Flux Decoupling

- The primary goal of rotor flux orientation is to decouple the torque and flux control in the motor. By controlling these two aspects independently, the system can respond more effectively to changes in load and speed.

3.2.2. Coordinate Transformation

- The three-phase currents flowing through the motor are transformed into a two-axis system, typically using the Clarke and Park transformations. The Park transformation, in particular, aligns the d-axis with the rotor's magnetic field, effectively rotating the reference frame to match the rotor's position.

3.2.4 Control Axes

- In the d-q coordinate system:
 - The **d-axis** (direct axis) is aligned with the rotor's magnetic field and is responsible for controlling the flux.
 - The **q-axis** (quadrature axis) is perpendicular to the d-axis and controls the torque.
- By adjusting the currents in these axes independently, one can regulate the motor's torque and magnetic flux optimally.

3.2.5 Feedback Mechanism

- A feedback loop using rotor position sensors (like encoders) or sensorless methods ensures that the system can accurately determine the rotor's position and speed. This information is essential for maintaining proper alignment of the d-q axes with the rotor flux.

3.2.6 Dynamic Response

- With rotor flux orientation, the motor can achieve a fast dynamic response to changes in load or speed. This capability is particularly important in applications that require rapid acceleration or deceleration.

3.2.7. Applications

- Rotor flux orientation is widely used in electric vehicles, robotics, and industrial drives, where precise control and high efficiency are critical.

In summary, the principle of rotor flux orientation enables the effective and efficient control of PMSMs by aligning the control strategy with the rotor's magnetic field, thus facilitating independent control of torque and flux for optimal performance.

Il existe trois types de commande vectorielle :

- Commande vectorielle directe.
- Commande vectorielle indirecte.
- Commande vectorielle simplifiée.

3.4 Vector Control Strategy of Permanent Magnet Synchronous Machines

The vector control strategy, also known as field-oriented control (FOC), is a sophisticated technique used to manage the performance of Permanent Magnet Synchronous Machines (PMSMs). This approach allows for precise control of the motor's torque and speed, making it particularly advantageous for applications requiring high performance and responsiveness. Here's an overview of the key elements of this strategy:

3.4.1 Principle of Operation

- The fundamental concept of vector control is to decouple the control of torque and magnetic flux. This is achieved by aligning the motor's control axes (d-axis and q-axis) with the rotor's magnetic field, allowing independent manipulation of each component.

3.4.2 . Coordinate Transformation

- **Clarke Transformation**: Converts the three-phase current (abc) into a two-phase stationary coordinate system ($\alpha\beta$).

- **Park Transformation**: Rotates the stationary reference frame to align with the rotor's magnetic field, producing the d-q coordinate system. In this system, the d-axis is aligned with the rotor's flux, and the q-axis is responsible for torque generation.

3.4.3 Control Strategy

- **Current Control**: The strategy employs a two-loop control structure:

- **Inner Loop**: Regulates the d-axis and q-axis currents using proportional-integral (PI) controllers. The d-axis current controls the magnetic flux, while the q-axis current controls the torque.

- **Outer Loop**: Controls speed or position by adjusting the reference values for the d-q currents based on the desired performance.

3.4.4 . Feedback Mechanism

- The system typically uses rotor position sensors (like encoders or resolvers) to provide real-time feedback on the rotor's position and speed. This information is crucial for maintaining proper alignment of the d-q axes and ensuring accurate control.

3.4.5. Advantages

- **High Performance**: Vector control allows for rapid response to changes in load and speed, making it suitable for dynamic applications.

- **Efficiency**: By optimizing the use of currents, vector control enhances the overall efficiency of the PMSM.

- **Smooth Operation**: The strategy minimizes torque ripple and improves the smoothness of operation, which is vital in precision applications.

3.4.6. Applications

- The vector control strategy is widely used in various fields, including:

- Electric and hybrid vehicles
- Robotics
- Industrial drives
- HVAC systems

3.4.7. Challenges

- Implementing vector control requires precise estimation of rotor position and knowledge of machine parameters, which can be challenging in variable conditions. Sensorless control techniques and adaptive algorithms are often employed to address these challenges.

In summary, the vector control strategy for Permanent Magnet Synchronous Machines provides an effective framework for achieving high-performance control, enhancing the machine's efficiency, and ensuring smooth operation across diverse applications.

The MSAP model supplied with voltage is given by the following equations:

$$\begin{cases} V_d = R_s I_d + L_d \frac{dI_d}{dt} - L_q \omega I_q \\ V_q = R_s I_q + L_q \frac{dI_q}{dt} + L_d \omega I_d + \varphi_{sf} \omega \end{cases} \quad (1)$$

$$J \frac{d\omega_r}{dt} = C_{em} - C_r - f \omega_r \quad (2)$$

$$C_{em} = (P(L_d - L_q)I_d I_q + \varphi_{sf} \omega \quad (3)$$

The model described by the previous equations shows that the MSAP is a multivariable system, non-linear and strongly coupled.

The main purpose of vector control of alternating current machines is to improve their static and dynamic behaviors thanks to a control structure similar to that of a direct current machine.

The machine studied has smooth poles, its electromagnetic torque is maximum when the stator current in the Park frame of reference is kept fixed on the q axis ($I_s = I_q$). This comes back to impose the zero direct component of the current (figure III.1) which has the effect of reducing the joule losses. The system of equations (1) becomes:

$$\begin{cases} V_d = -L_q \omega I_q \\ V_q = R_s I_q + L_q \frac{dI_q}{dt} + \varphi_{sf} \omega \end{cases} \quad (4)$$

$$C_{em} = P \cdot \varphi_{sf} \cdot I_q \quad (5)$$

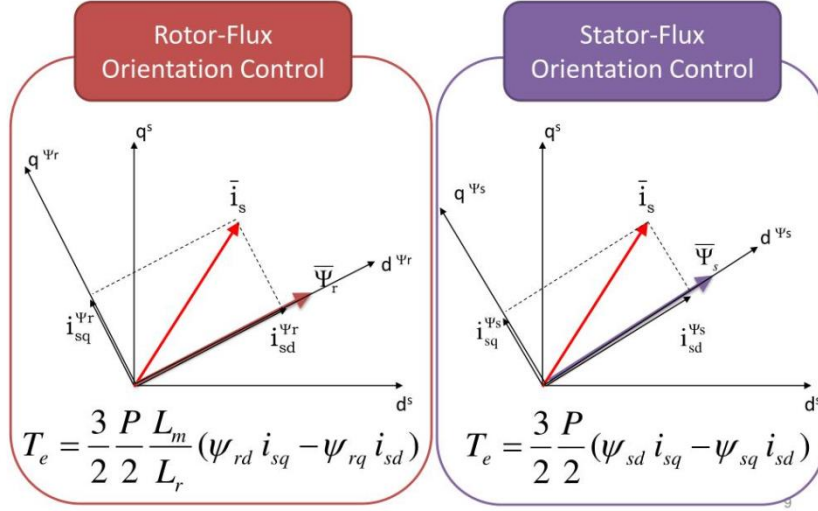


Figure (3.1) Principles of field orientation control

We note that this strategy makes it possible to eliminate the coupling problem between the “d” and “q”. The model (4), (5) and the mechanical equation can be represented by the diagram, functional figure (3) which resembles that of the direct current machine.

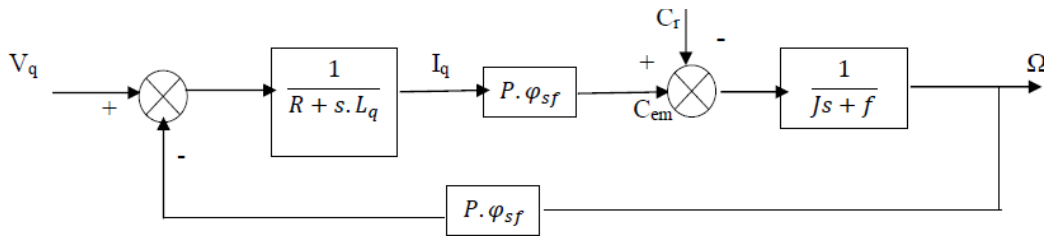


Figure (3.2) MSAP model when I_d is zero

3.4.8 Cutting by compensation:

The voltage supply is obtained by imposing reference voltages at the input of the control of the inverter. These tensions make it possible to define the cyclical ratios on the arms of the inverter so that the voltages delivered by this inverter to the terminals of the stator of the machine are as close as possible to the reference voltages. But, it is necessary to define compensation terms, because, in the stator equations, there are coupling terms between the d and q axes.

The voltages along the axes (d, q) can be written in the following forms

$$\begin{cases} V_d = \left(R_s I_d + L_d \frac{dI_d}{dt} \right) - \omega L_q I_q \\ V_q = \left(R_s I_q + L_q \frac{dI_q}{dt} \right) + \omega (L_d I_d + \varphi_{sf}) \\ \omega = P \cdot \omega_r \end{cases} \quad (6)$$

Figure 3.3 represents the coupling between the “d” and “q” axis:

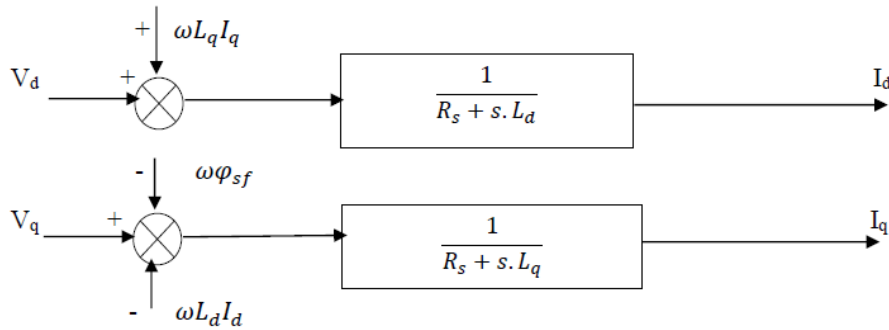


Figure (3.3) Coupling description

The voltages \$V_d\$ and \$V_q\$ depend both on the currents on the “d” and “q” axes, we are therefore led to implement decoupling. This decoupling is based on the introduction of compensatory terms and eq

$$\begin{cases} e_d = \omega L_q I_q \\ e_q = \omega (L_d I_d + \varphi_{sf}) \end{cases} \quad (7)$$

From equation 6 and 7 we have :

$$\begin{cases} V_d = V_{d1} - e_d \\ V_q = V_{q1} + e_q \end{cases} \quad (8)$$

$$\begin{cases} V_{d1} = (R_s + S \cdot L_d) I_d \\ V_{q1} = (R_s + S \cdot L_q) I_q \end{cases} \quad (9)$$

$$\begin{cases} e_d = \omega L_q I_q \\ e_q = \omega (L_d I_d + \varphi_{sf}) \end{cases} \quad (10)$$

$$\begin{cases} I_d = \frac{V_{d1}}{R_s + S \cdot L_d} \\ I_q = \frac{V_{q1}}{R_s + S \cdot L_q} \end{cases} \quad (11)$$

s: Laplace operator;

The regulation principle consists of regulating the stator currents from the magnitudes of reference (desired) by conventional regulators. The basic diagram for regulating stator currents is represented by the figure below[26]:

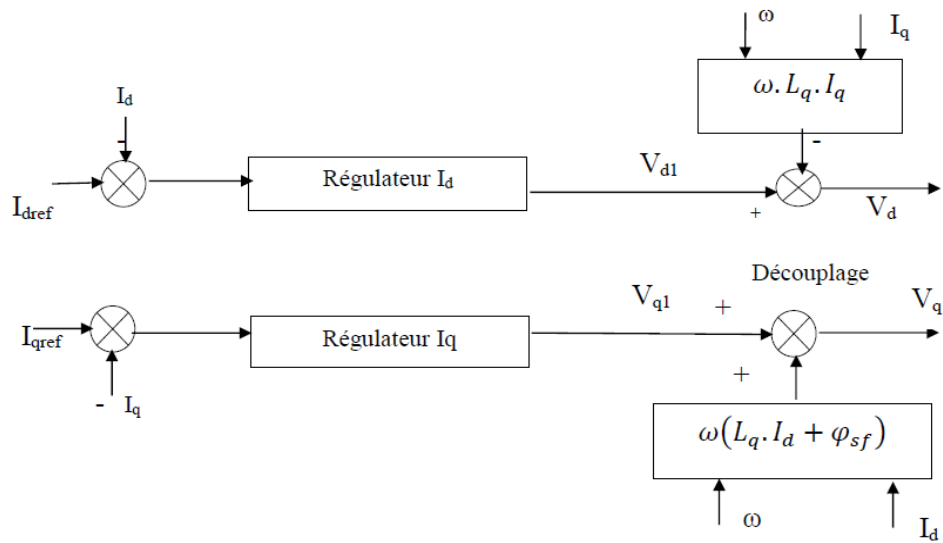


Figure (3.4) Principle of decoupling by compensation

avec :

vd1: La tension à la sortie de régulateur de courant “Id”

vq1: La tension à la sortie de régulateur de courant “Iq”

Si on associe la machine avec le bloc de compensation on obtient la figure (3.5)

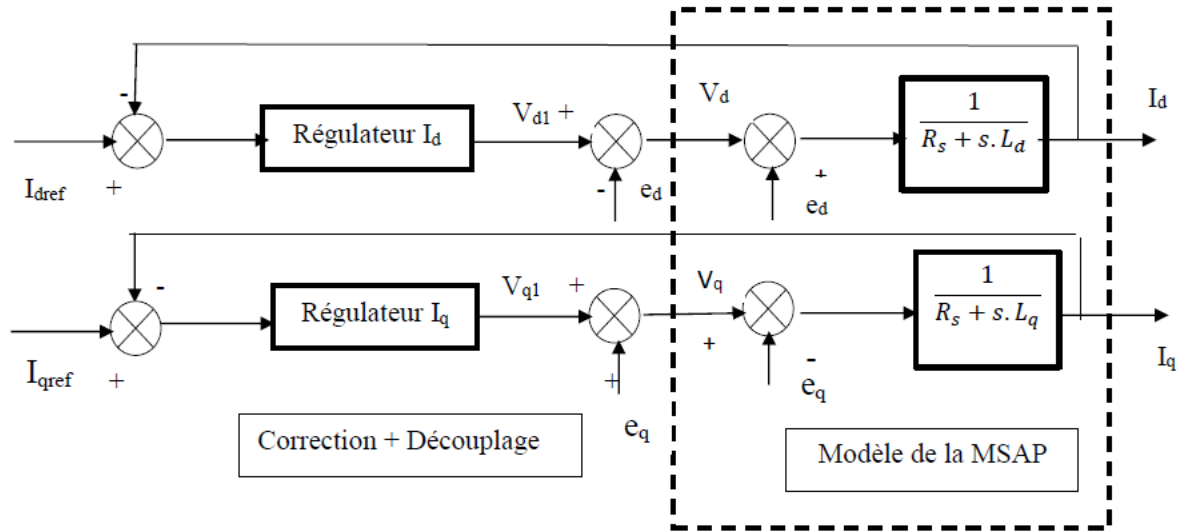


Figure (3.5) General structure: (machine-decoupling by compensation)

The actions on the “d” and “q” axes are therefore decoupled and represented by figure (3.6):

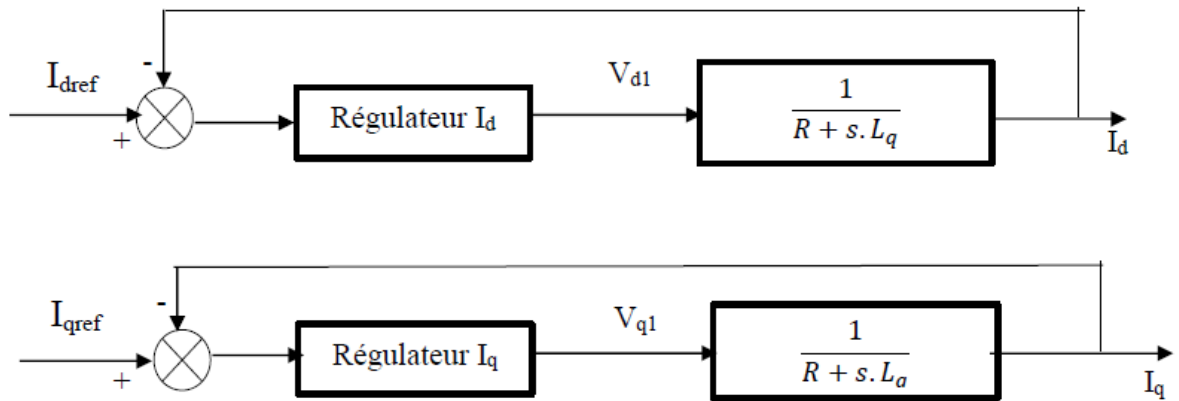


Figure (3.6) Decoupled control

3.5. Direct vector control

This method requires a good knowledge of the flow module and its phase and this must be verified regardless of the transitional regime carried out. A first possibility is to put flow sensors in the air gap and directly measure the components

$\phi_r\beta$ and $\phi_r\alpha$ so as to deduce the amplitude and phase.

The sensors, mechanically fragile, are subjected to severe conditions due to vibrations and warm-ups. The accuracy of the flux definition depends on the inductive parameters affected by the saturation of the magnetic circuit. On the other hand, the signals captured are tainted by noises generated by the notches and require adjustable filters. Direct measurement allows 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 motor equipped with flux sensors, which increases considerably the cost of its manufacture and makes its use more fragile

3.5.1 Principle of indirect vector control

In this method, we do not regulate the rotor flux, we therefore need neither a sensor nor estimation or flux observer. If the amplitude of the real rotor flux is not used, its position must be known to make coordinate changes. This requires the presence a rotor position/speed sensor. Poor information on velocity can interfere with determining the position of the flux in indirect control. In addition, this command is very sensitive to parametric variations and in particular the rotor time constant, i.e. R_r , which is involved in the definition of ω_s .

3.5.2 Advantages and Disadvantages of vector control

3.5.2.1 Advantages of vector control

- ☐ It is based on the transitional model (treating transitional regimes which was not possible with the classic variator)
- ☐ It is precise and fast.
- ☐ There is torque control when stopping.
- ☐ The control of quantities is done in amplitude and phase.

3.5.2.2 Disadvantages of Vector Control

Vector control by orientation of the rotor flux presents a certain number disadvantages:

Very expensive (incremental encoder or speed estimator, DSP.). Low robustness to parametric variations and in particular to those of the time constant of the motor.

Need for a modulator for close control of the inverter which causes delays, especially at low modulation frequency (high power). These delays are responsible for an increase in torque response time, which penalizes the variator used in traction. Presence of coordinate transformations depending on an estimated angle θ .

the rotation speed intervenes explicitly in the control algorithm. When we

does not measure this speed (variator without speed sensor), the errors in the estimate of this speed degrade the performance of the drive.

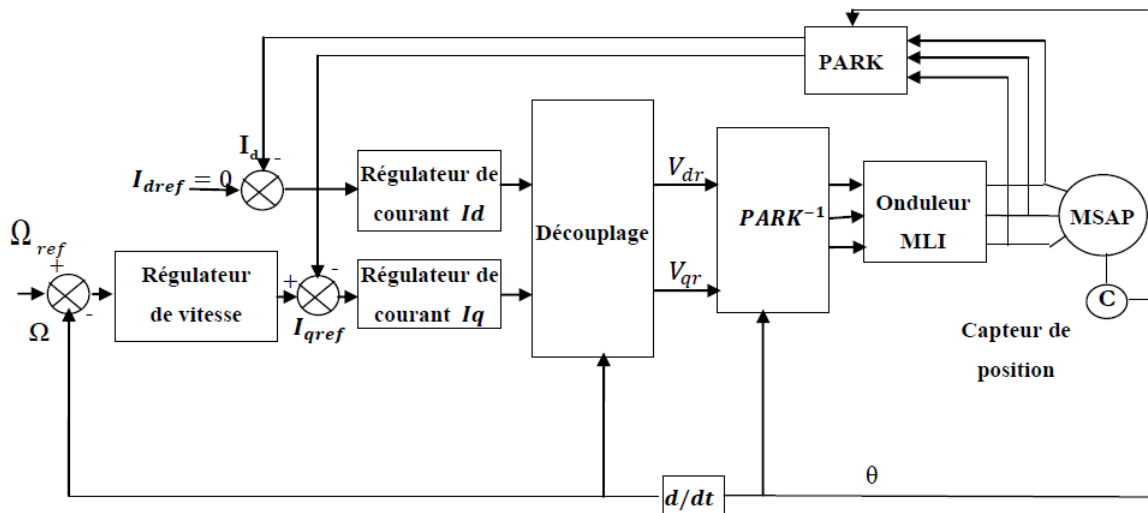


Figure (3.7) Block diagram of MSAP speed regulation supplied with voltage and controlled by flow orientation

Recitation n° :3

Exercise n° 1 :

The figure below represents the block diagram of the vector control by orientation of the rotor flux of a three-phase asynchronous cage machine supplied with current. This is a 12kW machine, 230 (between lines), 60Hz, 4 poles whose parameters are : $R_s=0.095 \Omega$; $R_r=0.2 \Omega$, $L_s=51.41\text{mH}$, $L_r=51.39 \text{ mH}$, $M=49.6 \text{ mH}$, $J=0.01667 \text{ kgm}^2$

En négligeant la chute tension aux bornes de la résistance statorique, calculer en régime sinusoïdal le flux statorique nominal maximal.

2°) En admettant que la machine tourne avec une vitesse de 1680 tr/min et développe une puissance utile de 9.7 kW avec des pertes mécanique négligeables; et que la valeur du flux rotorique de référence est égale au flux statorique nominal maximal :

a°) Express the currents i_{sd}^* , i_{sq}^* and the sliding speed w_r^* , outputs of the FOC (Field Oriented) block Control), depending on torque C_{em}^* and rotor flux ϕ_r . Calculate the values of these quantities of reference and deduce that of the effective value of the stator current.

Exercise n° 2 :

The figure below represents the structure of the direct vector control of a current-fed squirrel cage asynchronous machine (MAS)

1) Give the model of the rotor side of a three-phase asynchronous machine with a cage in a frame linked to the field turning (dq) by specifying:

a) The equations of the rotor tensions as a function of the rotor fluxes ϕ_{rd} , ϕ_{rq} , the rotor currents i_{rd} , i_{rq} , the electrical rotation speed of the machine mark (dq)

b) The equations of the rotor fluxes ϕ_{rd} , ϕ_{rq} as a function of the rotor currents i_{rd} and i_{rq} and the currents stator i_{sd} and i_{sq} .

2) Express the derivatives of the rotor fluxes in the following form:

$$\begin{aligned}\frac{d\phi_{rd}}{dt} &= f_1 \\ \frac{d\phi_{rq}}{dt} &= f_2\end{aligned}$$

Where f_1 and f_2 are functions to be determined

3) Taking into account the expressions of f_1 and f_2 , transform equation (1) into a reference frame linked to the rotor (xy) then deduce the expressions of the rotor fluxes estimated in the reference frame (xy) in the form following:

$$\begin{aligned}\hat{\phi}_{rx} &= \int g_1 dt \\ \hat{\phi}_{ry} &= \int g_2 dt\end{aligned}$$

Where g_1 and g_2 are functions to be determined

Exercise n°: 3

In a frame of reference linked to the rotating field, the model of the asynchronous machine can be put in the form following matrix :

$$\frac{d}{dt} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \phi_{dr} \\ \phi_{qr} \end{bmatrix} = \begin{bmatrix} -(\frac{1}{\sigma T_r} + \frac{1-\sigma}{\sigma T_r}) & \omega_s & \frac{1-\sigma}{\sigma M T_r} & \frac{1-\sigma}{\sigma M} \omega \\ -\omega_s & -(\frac{1}{\sigma T_r} + \frac{1-\sigma}{\sigma T_r}) & -\frac{1-\sigma}{\sigma M} \omega & \frac{1-\sigma}{\sigma M T_r} \\ \frac{M}{T_r} & 0 & -\frac{1}{T_r} & \omega_s - \omega \\ 0 & \frac{M}{T_r} & -(\omega_s - \omega) & -\frac{1}{T_r} \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \phi_{dr} \\ \phi_{qr} \end{bmatrix} + \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix}$$

with T_s the stator time constant, L_s the stator cyclic inductance, M the mutual inductance, the dispersion coefficient, and T_r the rotor time constant

1°) Calculate the voltages v_{ds} v_{qs} in the case where a vector control with orientation of the rotor flux is adopted .

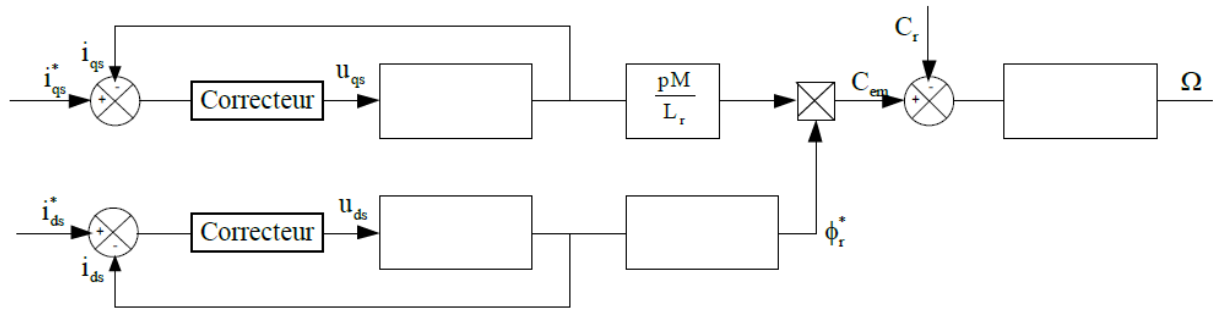
2°) In order to decouple the system, new control variables are introduced as follows

$$u_{ds} = v_{ds} + e_d = \sigma L_s \frac{di_{ds}}{dt} + R_{eq} i_{ds}$$

$$u_{qs} = v_{qs} + e_q = \sigma L_s \frac{di_{qs}}{dt} + R_{eq} i_{qs}$$

Identify the terms e , q , e , eq R .

3°) In the case of a completely decoupled system, complete the following block diagram by adding the functions adequate transfer



Direct Torque Control (DTC) is an advanced method for managing the performance of induction motors and Permanent Magnet Synchronous Motors (PMSMs). Developed in the late 1980s, DTC offers a robust solution for achieving high dynamic response and precise torque control, making it ideal for applications requiring rapid acceleration and deceleration.

Unlike traditional control methods, such as scalar control or field-oriented control, DTC directly regulates the motor's torque and flux in real-time without the need for complex coordinate transformations. This is achieved through a torque and flux estimator, which continuously monitors the motor's state and adjusts the voltage vectors applied to the motor accordingly.

The primary advantages of DTC include improved responsiveness, reduced torque ripple, and enhanced efficiency. By enabling direct manipulation of torque and flux, DTC provides excellent performance across a range of operating conditions. Its simplicity in implementation and effectiveness in control make it a popular choice in applications like electric vehicles, industrial drives, and robotics.

Overall, Direct Torque Control represents a significant advancement in motor control technology, providing engineers with the tools to enhance performance, efficiency, and reliability in electric drive systems.

Direct Torque Control (DTC) is a sophisticated technique used to regulate the torque and flux of induction motors and Permanent Magnet Synchronous Motors (PMSMs). The fundamental principles of DTC can be summarized as follows:

4.1. Direct Control of Torque and Flux:

DTC directly controls the motor's electromagnetic torque and magnetic flux without requiring complex transformations or modulation schemes. This is achieved by manipulating the inverter's voltage vectors in real-time based on the desired torque and flux levels.

4.1.1 Real-Time Estimation

DTC employs a torque and flux estimator that calculates the current state of the motor based on measurable parameters, such as current and voltage. These estimations provide the necessary feedback for controlling torque and flux effectively.

4.1.2. Switching Table

The core of DTC involves a predefined switching table that relates the estimated torque and flux values to specific voltage vector commands. Based on the current estimates, the controller selects the optimal voltage vector to apply, facilitating rapid adjustments to both torque and flux.

4.1.3. Hysteresis Control:

DTC uses hysteresis bands for both torque and flux control. This means that if the estimated torque or flux moves outside of predefined limits, the controller activates the necessary voltage vector to bring the values back within range. This results in quick responses to changes in load or speed.

4.1.4. Voltage Vector Selection:

The controller selects from a finite set of voltage vectors generated by the inverter. Each vector corresponds to a specific torque and flux change. By rapidly switching between these vectors, DTC achieves high-performance control.

4.1.5. Minimal Torque Ripple:

One of the key advantages of DTC is its ability to minimize torque ripple, resulting in smoother operation. The direct control approach ensures that the motor operates efficiently across various loads and speeds.

4.1.6. Implementation:

DTC is implemented in digital controllers, such as microcontrollers or digital signal processors (DSPs), which perform the necessary calculations and voltage vector selections at high speeds.

In essence, Direct Torque Control revolutionizes motor management by allowing for immediate and precise adjustments to torque and flux through real-time estimations and rapid

voltage vector selection. This leads to enhanced performance, efficiency, and responsiveness in electric drive systems.

4.2. Basic direct torque control and direct self-control

The direct torque control (DTC) has been proposed by Takahashi for induction motor drives in the middle of 1980s. It bases on the direct selection of voltage vector (switching state) for the voltage inverter which fed the motor according to the instantaneous errors of the stator flux and the electromagnetic torque. DTC uses separated hysteresis controllers to ensure a decoupled control of flux and torque without requiring a complex field orientation or current regulation loop. The outputs of the hysteresis comparators choose the appropriate voltage vector through a look-up switching table along the estimated position of the flux vector.

Usually, the mathematical model of IM is used to estimate the stator flux and the electromagnetic torque. Fig.1.4 shows a simple structure of the switching table based DTC which proposed by Takahashi [Tak86].

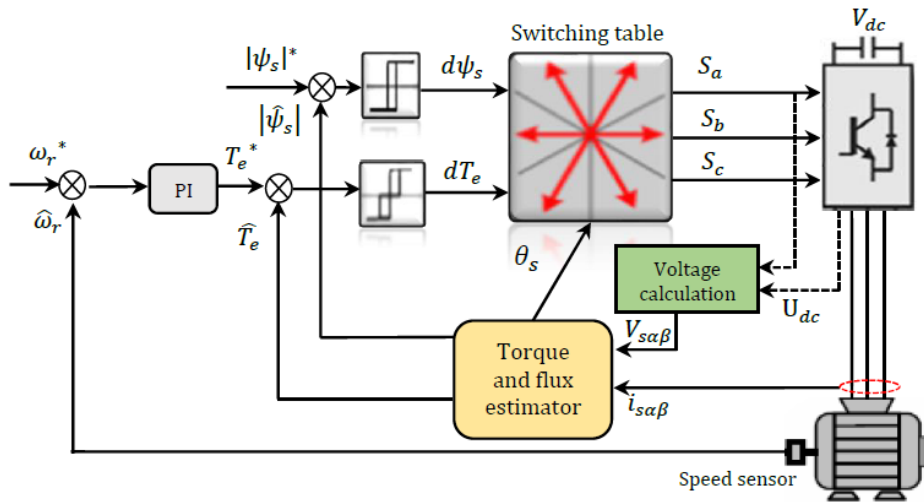


Figure (4.1) Direct torque control strategy based on lookup switching table.

This control scheme is very often used in the industry in low and medium power applications. The main advantages of DTC are summarized in its fast dynamic of the drive, the absence of coordinate transformations and current control loops and its universal structure, where the switching table DTC can be used for all AC machines. In the other hand, the main disadvantages of DTC are the variable switching frequency, high torque ripples and high switching losses.

Another similar method which is dedicated for high power applications, it was presented in the middle of 1980 also by Depenbrock, it is so-called direct self-control (DSC). It can be characterized as a simplified version of DTC.

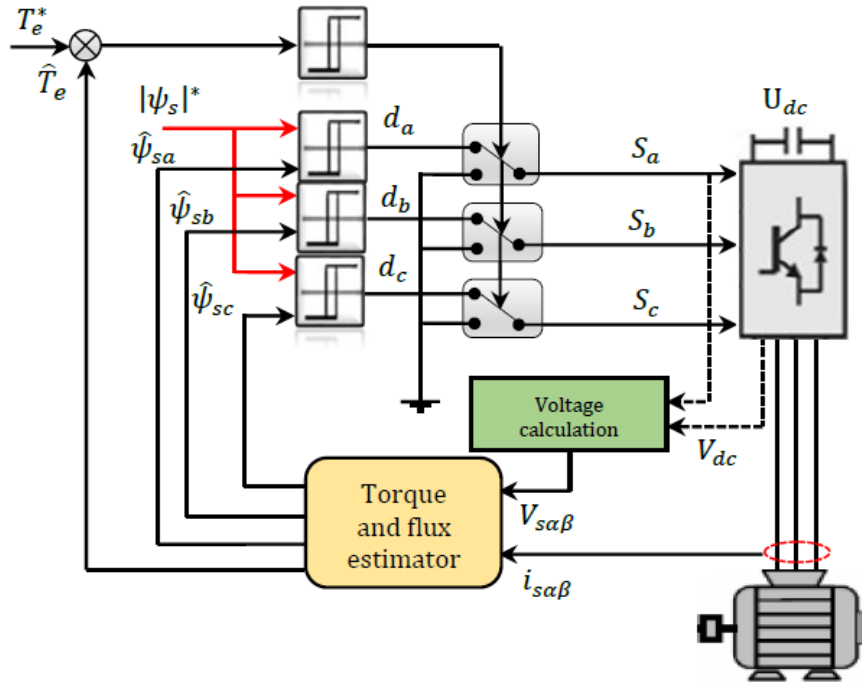


Figure (4.2) Block diagram of direct self-control (DSC)

This technique use the reference and the three phase components of flux to generate the switching states which correspond to active voltage vectors through three hysteresis flux comparators. The hysteresis torque comparator generates the signal which correspond to the zero states.

The common applications of DSC scheme are the high-power drives and traction systems, DSC is mainly featured by low inverter switching frequency (hexagon stator flux trajectory), the absence of switching table or PWM modulator, good torque and flux dynamics. In the other side, it has non-sinusoidal current and high ripples.

4.3. Direct torque control with constant switching frequency

In this DTC strategy, the lookup switching table and the hysteresis comparators are replaced by the space voltage modulation (SVM) unit and PI controllers for the purpose of voltage vector selection. This control schemes can preserve a constant switching frequency, consequently it reduces the high torque and flux ripples which is the main drawback

of the conventional switching table DTC. SVM was first presented in the second half of the 1980s as an alternative strategy to the basic Pulse Width Modulator (PWM). Since then, a lot of work has been done on its theory and implementation.

SVM bases on the space vector representation of the voltage inverter output. It has several advantages such as minimizing ripple and Total Harmonic Distortion (THD) and switching losses. The incorporation of SVM with DTC schemes has been presented at the first time by Habetler in 1992, then several SVM-DTC schemes have been proposed.

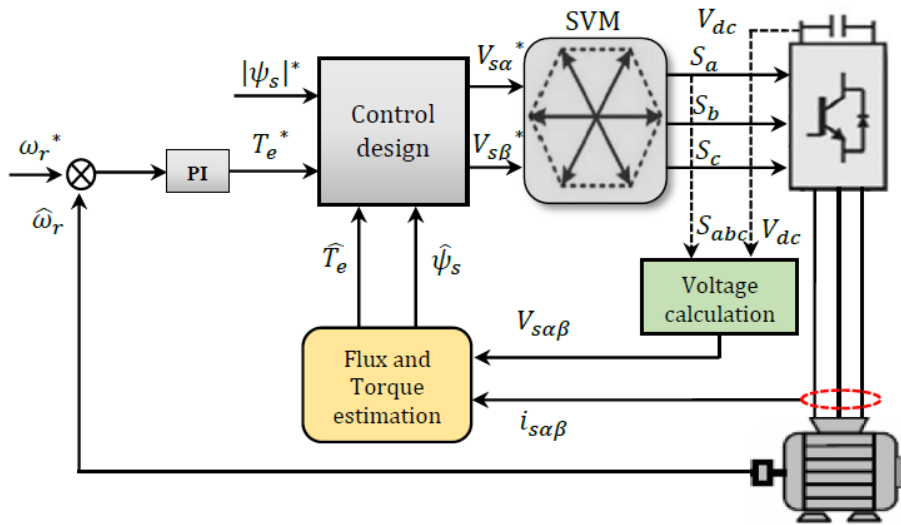


Figure (4.3) Block diagram of direct self-control (DSC)

The common target of these various structures is providing the reference voltages to modulate them by the SVM unit in order to generate the inverter switching state. Figure 4.3 above illustrates a general diagram of SVM-DTC control algorithm. The SVM-DTC keeps a number of classical DTC features and offer more advantages, such as the constant switching frequency and low flux and torque ripples.

4.3. Control of stator flux and electromagnetic torque

4.3.1 Control of stator flux

Basing on the IM model in stationary frame, the stator flux equation can be expressed as follows:

$$\frac{d\psi_s}{dt} = V_s - R_s i_s \quad (1)$$

$$\psi_s(t) = \int_0^{T_z} (V_s - R_s i_s) dt + \psi_s(0) \quad (2)$$

ψ_{s0} : is the flux vector at the instant $t=0$ s.

By applying a non-zero vector in T_z sampling period, we can neglect the stator resistance voltage drop $R_s i_s$ compared to V_s for high speed regions. Then Eq (5) can be written as:

$$\psi_s(t) \approx V_s T_z + \psi_s(0) \quad (3)$$

The relation between the stator voltage and the stator flux change can be established as:

$$\Delta\psi_s = \psi_s(t) - \psi_s(0) = V_s T_z \quad (4)$$

The Eq (3) means that the stator flux can be changed by the application of stator voltage during a time T_z . The stator flux vector's extremity moves in direction given by the voltage vector and making a circular trajectory

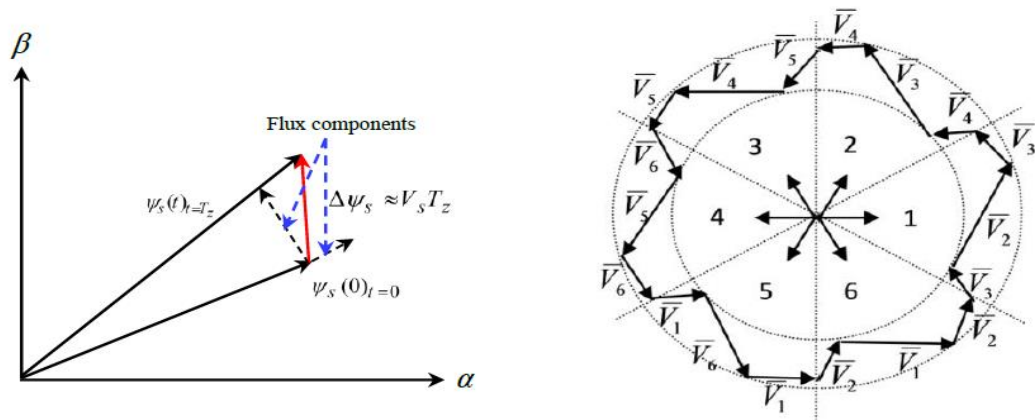


Figure (4.4) Evolution of stator flux vector in the complex plan.

A two-level hysteresis comparator is used for flux regulation. It allows to drop easily the flux vector extremity within the limits of the two concentric circles with close radius, as shown in Figure.4.4. The choice of the hysteresis bandwidth $h\psi_s$ depends on the switching frequency of the inverter

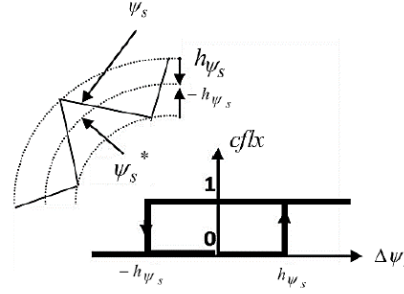


Figure (4.5) Two-level hysteresis comparator for stator flux control.

The logical outputs of the flux controller are defined as:

$$\begin{cases} cflx = 1 & \text{if } \Delta\psi_s > h\psi_s \\ cflx = 0 & \text{if } \Delta\psi_s \leq -h\psi_s \end{cases} \quad (5)$$

$h\psi_s$ is hysteresis band of stator flux

The stator flux error is defined by the difference between the references value of flux and the actual estimated value:

$$\Delta\psi_s = |\psi_s^*| - |\psi_s|$$

4.4 Control of electromagnetic torque

During one sampling period, the rotor flux vector is supposed invariant. The torque of induction motor can be expressed in terms of stator and rotor flux vectors as follows:

$$T_e = p \frac{M_{sr}}{\sigma L_s L_r} \psi_s \times \psi_r \quad (6)$$

$$|T_e| = p \frac{M_{sr}}{\sigma L_s L_r} |\psi_s| |\psi_r| \sin(\delta) \quad (7)$$

where:

p is the number of poles pairs

Ψ_s, Ψ_r are stator and rotor flux vectors.

δ angle between the stator and rotor flux vectors

From expression , it is clear that the electromagnetic torque is controlled by the stator and rotor flux amplitudes. If those quantities are maintaining constant, the torque can be controlled by adjusting the load angle δ .

The torque regulation can be realized using three-level hysteresis comparator (**Fig.4.6**). It allows to control the motor in both rotation senses. The two-level comparator can be used for one rotation sense.

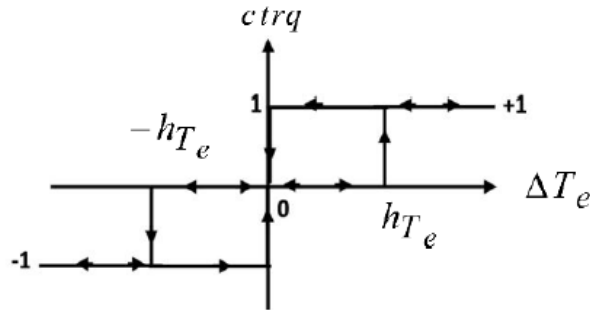


Figure (4.6) Hysteresis comparator two level

$$\begin{cases} ctrq = 1 & \text{if } \Delta T_e > h_{T_e} \\ ctrq = 0 & \text{if } -h_{T_e} \leq \Delta T_e \leq h_{T_e} \\ ctrq = -1 & \text{if } \Delta T_e < -h_{T_e} \end{cases}$$

h is hysteresis band of torque.

The torque error is defined by the difference between the references values of the torque and the actual estimated values:

$$\Delta T_e = T_e^* - T_e$$

4.5 Switching Table Construction and Control Algorithm Design

4.5.1 Six sectors Switching table

To maintain a decoupled control, pair of hysteresis comparators receives the stator flux and torque errors as inputs. Then, the comparators outputs determine the appropriate voltage vector selection. However, the choice of voltage vector is not only depending on the output of hysteresis controllers, but on the position of stator flux vector also. Thus, the circular stator flux vector trajectory will be divided into six symmetrical sectors.

where:

Sector 1:

$$11\pi/6 \leq \theta_s < \pi/6, \text{ sector 2: } \pi/6 \leq \theta_s < \pi/2, \dots, \text{ sector 6: } 3\pi/2 \leq \theta_s < 11\pi/6$$

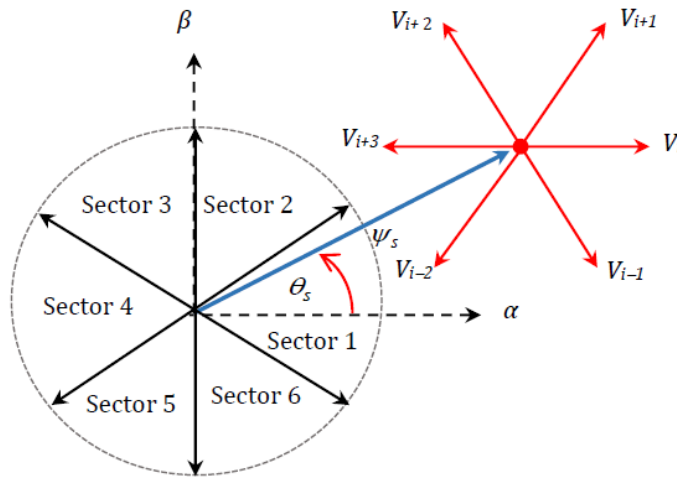


Figure (4.7) Voltage vector selection when the stator flux vector is located in sector i .

While the stator flux vector is located in the sector i we have

If V_{i+1} is selected, Ψ_s increases and T_e increases.

– If V_{i-1} is selected, Ψ_s increases and T_e decreases.

– If V_{i+2} is selected, Ψ_s decreases and T_e increases.

– If V_{i-2} is selected, Ψ_s decreases and T_e decreases.

For each sector, the vectors ($i V$ and $i+3 V$) are not considered because both of them can increase or decrease the torque in the same sector according to the position of flux vector on the first or the second sector. If the zero vectors V_0 and V_7 are selected, the stator flux will stop moving and its magnitude will not change, the electromagnetic torque will decrease, but not as much as when the active voltage vectors are selected. The resulting look-up table for DTC which was proposed by Takahashi is presented in **Table 4.1**:

Error	Sectors	I	II	III	IV	V	VI
$cflx = 1$	$ctrq = 1$	V_2	V_3	V_4	V_5	V_6	V_1
	$ctrq = 0$	V_7	V_0	V_7	V_0	V_7	V_0
	$ctrq = -1$	V_6	V_1	V_2	V_3	V_4	V_5
$cflx = 0$	$ctrq = 1$	V_3	V_4	V_5	V_6	V_1	V_2
	$ctrq = 0$	V_0	V_7	V_0	V_7	V_0	V_7
	$ctrq = -1$	V_5	V_6	V_1	V_2	V_3	V_4

Table 4.1 Look-up table for basic direct torque control.

4.5.2 DTC improvement using twelve sectors switching table

In the conventional six sectors DTC, two switching states per sector are not considered. This presents an ambiguity in torque control. A sector shifting takes the first sector from $(0$ to $\pi/3)$ instead of $(11\pi/6$ to $\pi/6)$. This provides a new switching table. However, in a similar way, it has two unused vectors per sector also.

which present an ambiguity in flux instead of torque. Another strategy divides the circular flux locus into 12 sectors instead of 6 as shown in **Figure 4.4**. This makes all the six states used per sector.

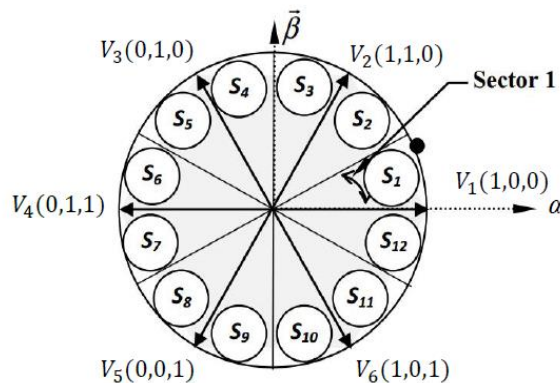


Figure (4.8) Voltage space vector in 12 sectors case.

In twelve sectors DTC, the vector V_1 produces a large increase in flux and a small increase in torque for the sector 12. On the contrary, V_2 produces a large increase in torque and small in flux. We can deduce that it is necessary to define now small and large torque variations. This requires to divide the hysteresis band of the torque into four parts. Then, a twelve-sectors look-up table is provided as shown in Table 4.2

Error	Sector	I	II	III	IV	V	VI	VII	VIII	IX	X	XI	XII
$cflx=1$	$ctrq = 2$	V_2	V_3	V_3	V_4	V_4	V_5	V_5	V_6	V_6	V_1	V_1	V_2
	$ctrq = 1$	V_2	V_2	V_3	V_3	V_4	V_4	V_5	V_5	V_6	V_6	V_1	V_1
	$ctrq = -1$	V_1	V_1	V_2	V_2	V_3	V_3	V_4	V_4	V_5	V_5	V_6	V_6
	$ctrq = -2$	V_6	V_1	V_1	V_2	V_2	V_3	V_3	V_4	V_4	V_5	V_5	V_6
$cflx=0$	$ctrq = 2$	V_3	V_4	V_4	V_5	V_5	V_6	V_6	V_1	V_1	V_2	V_2	V_3
	$ctrq = 1$	V_4	V_4	V_5	V_5	V_6	V_6	V_1	V_1	V_2	V_2	V_3	V_3
	$ctrq = -1$	V_5	V_5	V_6	V_6	V_1	V_1	V_2	V_2	V_3	V_3	V_4	V_4
	$ctrq = -2$	V_5	V_6	V_6	V_1	V_1	V_2	V_2	V_3	V_3	V_4	V_4	V_5

Table 4.2 Look up switching table with twelve sectors.

Considerable studies in literature mentioned that the increasing of the number of the sectors have a slight effect on reducing the high ripples and current harmonics. In addition, the twelve sectors DTC provides good dynamic in high and low speed regions

The integration of static converters with AC generators has brought significant advancements to variable speed wind systems. This combination enables seamless control of SCIG systems in wind conversion, allowing precise adjustments of speed and torque to meet specific requirements. The result is improved energy efficiency, reduced maintenance costs, and enhanced overall performance. This chapter delves into the control strategy employed for the Wind Energy Conversion System (WECS) utilizing a squirrel cage induction generator and employing PI controllers. The control approach entails the utilization of two controllers. The first controller is dedicated to governing the machine side converter, with the primary goal of extracting maximum power from the wind turbines. This controller regulates the electromagnetic torque and the field of the squirrel cage induction generator, optimizing power generation. On the other hand, the second controller is responsible for regulating the DC-bus voltage and managing the grid powers, encompassing both active and reactive components. Through the integration of these controllers, the control system ensures efficient power extraction from the wind turbines and effective management of the grid powers within the wind energy conversion system

5.1 Simplified description of the system to be controlled

The aero-turbine efficiently harnesses a specific portion (P turbine) of the available power (P incident) from the wind. This captured power is subsequently converted into electrical energy by the SCIG. The generated electrical power is primarily transmitted directly to the power grid from the stator windings of the generator via a DC bus, PWM inverter and filter [8]. Fig. (3.1) shows the adopted wind energy conversion system and its control.

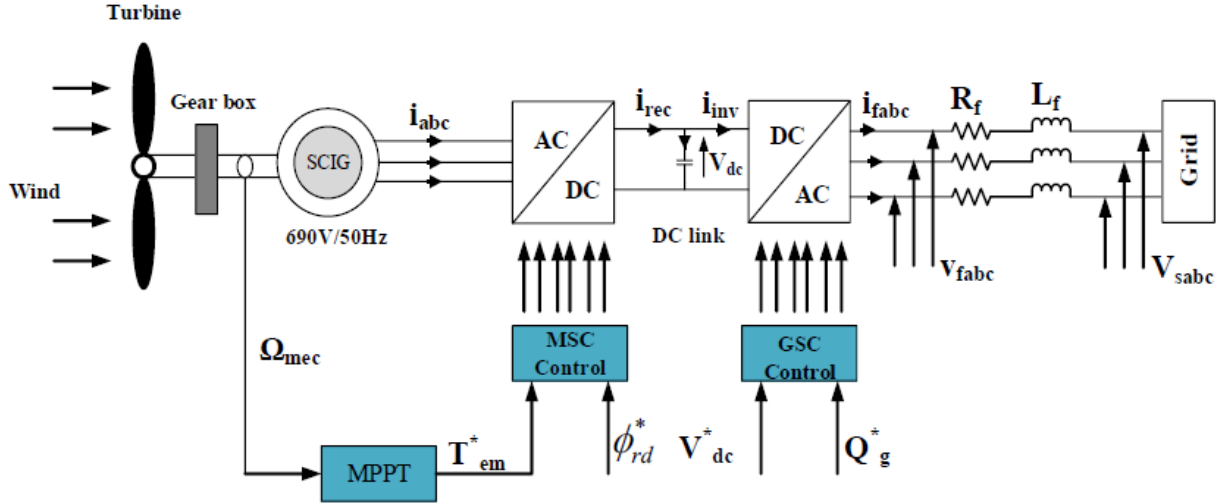


Figure (5.1) Control of wind energy conversion system based on SCIG.

5.2.1 MPPT control

Torque control, specifically the Maximum Power Point Tracking (MPPT) technique, plays a crucial role in optimizing the wind turbine's power extraction. To achieve maximum available power, it is essential to adjust the turbine speed relative to the wind speed. It is indispensable to maintain the tip speed ratio at an optimal value ($\lambda = \lambda_{opt}$) and the power coefficient at its maximum value ($C_p = C_{pmax}$). The speed of the squirrel cage induction generator (SCIG) is utilized as a reference for a proportional integral (PI) controller. The PI controller determines the control set point, which represents the desired electromagnetic torque to be applied to the machine, allowing the generator to operate at its optimal speed. By employing this torque control strategy and maintaining the appropriate control set point, the wind turbine can maximize power generation efficiency. The electromagnetic torque is given by

$$T_{em}^* = H_w(s) \cdot (\Omega_{mec}^* - \Omega_{mec}) \quad (1)$$

Where $H(s)$ is the speed regulator transfer function and Ω_{mec} is the reference mechanical speed. Reference mechanical speed depends on the turbine's speed, which corresponds to the optimal value of the speed ratio λ_{opt} (the pitch angle is fixed to $\beta=0$)

The reference mechanical speed is given by:

$$\Omega_{mec}^* = \frac{v_{wind} \lambda_{opt}}{R} G \quad (2)$$

With turbine's speed is given by :

$$\Omega_t^* = \frac{v_{wind} \lambda_{opt}}{R} \quad (3)$$

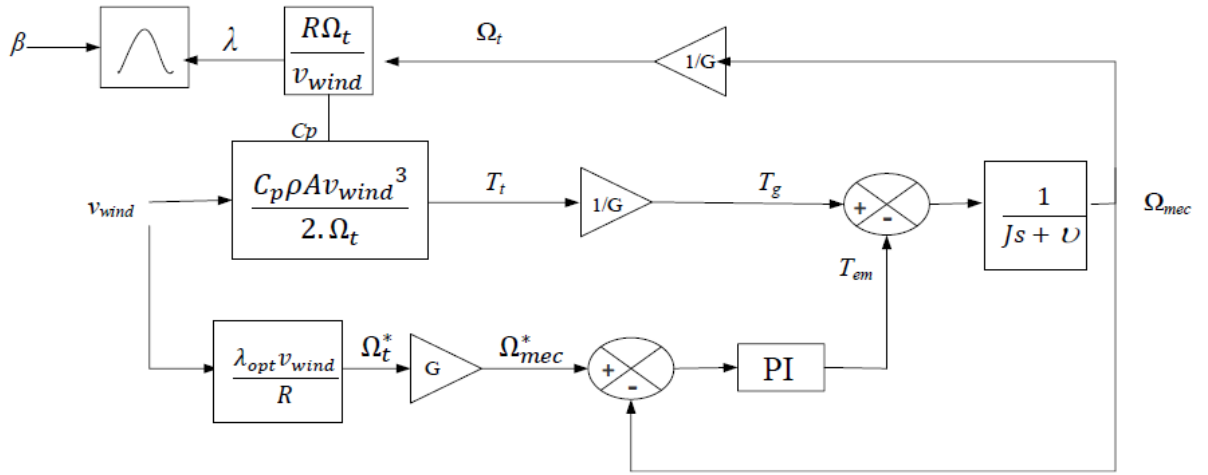


Figure. (5.2): MPPT control scheme.

The speed regulation loop will be defined by the parameters (k_{pw} , k_{iw}). Using the mechanical equation, the transfer function linking the speed to the electromagnetic and mechanical torques is as follows:

$$F_w(s) = \frac{\Omega_{mec}}{T_g - T_{em}} = \frac{1}{Js + \nu} \quad (4)$$

The PI transfer function is given by:

$$H_w(s) = \frac{k_{iw}}{s} + k_{pw} \quad (5)$$

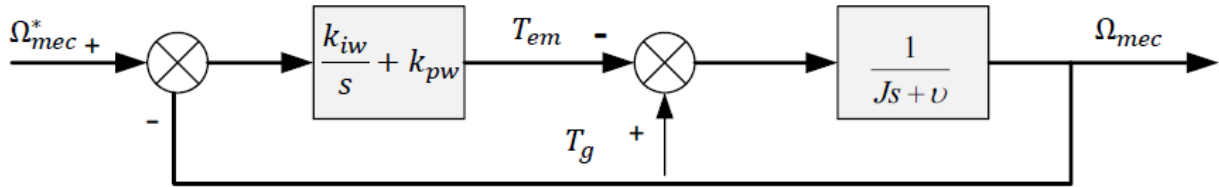


Figure. (5.3) Speed closed loop control.

The system closed-loop transfer function as presented in figure (5.3) is:

$$G_{dcs}(s) = \frac{\frac{k_{iw}}{J} + \frac{k_{pw}}{J}s}{s^2 + 2\left(\frac{k_{pw} + \nu}{J}\right)s + \frac{k_{iw}}{L_f}} \quad (6)$$

Comparing this characteristic equation of **(6)** with that of the following general form of a second-order transfer function:

$$T(s) = \frac{2\xi\omega_n s + \omega_n^2}{s^2 + 2\xi\omega_n s + \omega_n^2} \quad (7)$$

which yields:

$$\begin{aligned} k_{iw} &= J\omega_{nw}^2 \\ k_{pw} &= 2\xi\omega_{nw}J - \nu \end{aligned} \quad (8)$$

5.2.2 Machine side control

The control of the Machine Side Converter (MSC) in the WEC System, employing PI controllers, is widely adopted due to its straight forward implementation. The primary objective of this control strategy is to regulate both the electromagnetic torque and the field of the SCIG, aiming to extract the maximum mechanical power (P_{turbine}) available from the turbine across various wind speeds. To achieve the decoupled control of the electromagnetic torque and the generator's field, the approach of Field Oriented Control (FOC) is utilized. FOC enables precise regulation and optimization of both torque and field, ensuring efficient power extraction from the wind turbine in the WEC system the control scheme shown in figure (3.10)

5.2.2.1 Field Oriented Control (FOC)

Field-oriented control (FOC), by [7], is widely acknowledged as a high-performance control method. FOC offers exceptional control capabilities by decoupling the torque and magnetic flux of the machine, leading to rapid torque response. Unlike scalar control, FOC utilizes dynamic model of the induction machine, enabling precise control over the amplitude, frequency, and position of the space vectors of voltage, current, and magnetic flux.

By employing FOC, an AC machine can be effectively controlled as a separately excited DC machine. However, it is important to note that FOC techniques heavily rely on machine parameters and speed, which can impact the robustness of the control. Despite the ability of FOC to individually control the torque and magnetic flux of the induction machine, variations in these parameters and the machine's speed can potentially compromise control robustness. FOC is divided to the following two main types as described by [5]:

□ Direct Field-Oriented Control (DFOC) is a control method where the flux vector's amplitude and position are directly determined from measured quantities of the machine. In DFOC, the necessary information for controlling the flux vector is obtained directly from sensor measurements, enabling accurate and real-time calculations.

Indirect Field-Oriented Control (IFOC) calculates the flux vector's amplitude and position indirectly using speed and slip estimations.

Both DFOC and IFOC are techniques used in field-oriented control systems, aiming to achieve precise control over the flux vector in induction machines. While DFOC directly

measures the required quantities for flux control, IFOC utilizes estimations and mathematical models to derive the flux information indirectly. The choice between DFOC and IFOC depends on factors such as the availability of accurate sensor measurements and the desired level of control accuracy and complexity.

There are three types of flux orientation: Orientation of the rotor flux with the conditions , , orientation of the stator flux with the conditions

$$\phi_{rd} = \phi_r, \phi_{rq} = 0 \quad (9)$$

$$\phi_{sd} = \phi_s, \phi_{sq} = 0 \quad (10)$$

orientation of the air gap flux with the conditions

$$\phi_{rd} = \phi_r, \phi_{rq} = 0 \quad (11)$$

In this study, the indirect rotor flux orientation control is adopted (IR-FOC). This approach is achieved by aligning the d-axis of the synchronous reference frame with the rotor flux vector while the q-axis rotor flux linkage is zero. By Applying the orientation of the rotor flux the equation (11) of the electromagnetic torque becomes:

$$T_{em} = \frac{pM}{L_r} \phi_r i_{sq} \quad (12)$$

The electromagnetic torque expression is a proportional with the quadrature stator current as long as that the direct rotor field is constant. In steady-state, the direct stator current is given by:

$$i_{sd} = \frac{\phi_r}{M} \quad (13)$$

The rotor field angle is:

$$\theta_f = \int (\omega_{slip} + p\Omega_{mec}) dt \quad (14)$$

With the slip frequency being obtained by:

$$\omega_{slip} = \frac{R_r M i_{sq}}{L_s \phi_r} \quad (15)$$

The equation (12) obtained by replacing the rotor quadratic flux by zero in the model of SCIG in (d,q) frame:

$$\begin{cases} v_{sd} = R_s i_{sd} + \sigma L_s \frac{di_{sd}}{dt} + \frac{M}{L_r} \frac{d\phi_{rd}}{dt} - \sigma L_s \omega_s i_{sq} \\ v_{sq} = R_s i_{sq} + \sigma L_s \frac{di_{sq}}{dt} + \frac{M}{L_r} \phi_{rd} + \sigma L_s \omega_s i_{sd} \end{cases} \quad (16)$$

If we admit that the flux varies very slowly with respect to the stator currents, the previous equation is simplified to:

$$\begin{cases} v_{sd} = R_s i_{sd} + \sigma L_s \frac{di_{sd}}{dt} + \frac{M}{L_r} \frac{d\phi_{rd}}{dt} - \sigma L_s \omega_s i_{sq} \\ v_{sq} = R_s i_{sq} + \sigma L_s \frac{di_{sq}}{dt} + \frac{M}{L_r} \phi_{rd} + \sigma L_s \omega_s i_{sd} \end{cases} \quad (17)$$

Applying the Laplace transform results in:

$$\begin{cases} v_{sd} = (R_s + \sigma L_s s) i_{sd} - \sigma L_s \omega_s i_{sq} \\ v_{sq} = (R_s + \sigma L_s s) i_{sq} + \frac{M}{L_r} \omega_s \phi_{rd} + \sigma L_s \omega_s i_{sd} \end{cases} \quad (18)$$

Based on this (18) is defined by (19) The linear part of the system of equations system a regulation the stator currents can be established using PI controllers

$$\begin{cases} u_{sd} = (R_s + \sigma L_s s) i_{sd} \\ u_{sq} = (R_s + \sigma L_s s) i_{sq} \end{cases} \quad (19)$$

The system (19) shows that the stator voltages (u_{sd} , u_{sq}) are directly related to the stator currents (i_{sd} , i_{sq}). To compensate the system during decoupling, the reference stator voltages at constant flux are given by:

$$\begin{cases} v_{sd} = u_{sd} + e_{sd} \\ v_{sq} = u_{sq} + e_{sq} \end{cases} \quad (20)$$

Where

$$\begin{cases} e_{sd} = -\sigma L_s \omega_s i_{sq} \\ e_{sq} = \frac{M}{L_r} \omega_s \phi_{rd} + \sigma L_s \omega_s i_{sd} \end{cases} \quad (21)$$

a) Rotor flux control steps

-Step 1: Calculation of the reference current i_{sd} by direct inversion of the equation giving the flux according to the following equation: ϕ_{rd}

$$i_{rd}^* = \frac{\phi_{rd}^*}{M} \quad (22)$$

Step 2: control of the current i_{sd} while neglecting the coupling term e_{sd} . A PI regulator is used to calculate the reference voltage u_{sd}

* as follows:

$$u_{sd}^* = H(s)(i_{sd}^* - i_{sd}) \quad (23)$$

-Step 3: Application of a decoupling law making it possible to cancel the effect of the coupling term e_{sd} and to calculate the reference voltage according to the following equation:

$$v_{sd}^* = u_{sd}^* + e_{sd}^m \quad (24)$$

Where $e_{sd}^m = -\sigma L_s \omega_s i_{sq}$ value of e_{sd} .

b) Electromagnetic torque control steps

If the rotor flux is controlled so as to be maintained at a given value, the torque becomes proportional to the component in quadrature of the stator current i_{sq} .

- Step 1: Calculation of the reference current i_{sq} by direct inversion of the equation giving the torque according to the following equation: T_{em}

$$i_{sq}^* = \frac{L_r T_{em}^*}{pM} \quad (25)$$

- Step 2: control of the current i_{sq} while neglecting the coupling term e_{sq} . A PI regulator is used to calculate the reference voltage as follows:

$$u_{sq}^* = H(s)(i_{sq}^* - i_{sq}) \quad (26)$$

- Step 3: Application of a decoupling law which cancels the effect of the coupling term e_{sq} and to calculate the reference voltage according to the following equation:

$$v_{sq}^* = u_{sq}^* + e_{sq}^m \quad (27)$$

Where : e_{sq}^m is the measured value of e_{sq}

After compensation, an identical control diagram is obtained for the two axes (d) and (q). The scheme for adjusting direct and quadrature currents using PI regulators is shown in the figure (5.4)

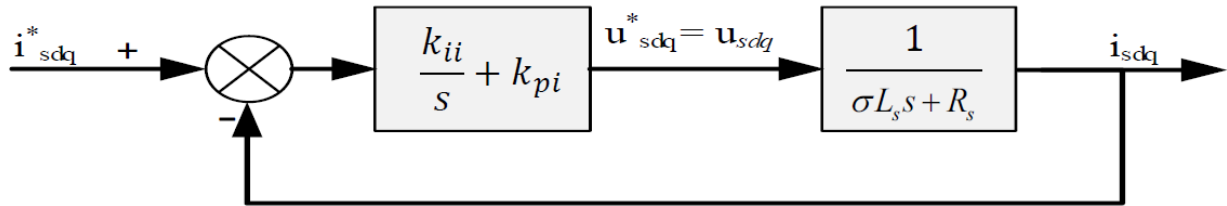


Figure (5.4) Stator current closed loop control.

The transfer function of the system is given by:

$$F(s) = \frac{i_{sdq}}{u_{sdq}} = \frac{1}{\sigma L_s s + R_s} \quad (28)$$

The PI transfer function is given by:

$$H(s) = \frac{k_{ii}}{s} + k_{pi} \quad (29)$$

The system closed-loop transfer function as presented in Figure (5.4) is given by:

$$G_s(s) = \frac{\frac{k_{pi}}{\sigma L_s} s + \frac{k_{ii}}{\sigma L_s}}{s^2 + \frac{k_{pi} + R_s}{\sigma L_s} s + \frac{k_{ii}}{\sigma L_s}} \quad (30)$$

PI controller gains can be obtained by identifying the characteristic equation of (26) with that of a transfer function of a second-order system defined by (27), which gives:

$$\begin{aligned} k_{ii} &= \sigma L_s \omega_{ni} \\ k_{pi} &= 2\sigma L_s \xi \omega_{ni} - R_s \end{aligned} \quad (31)$$

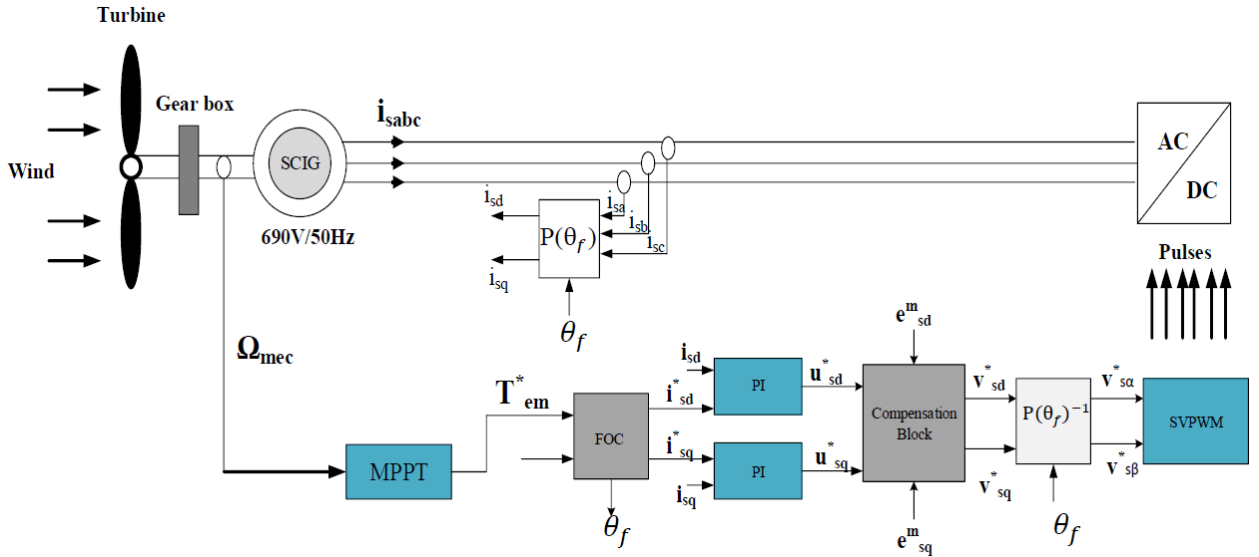


Figure (5.5): Generator side converter control by PI controllers

a) Fixed electromagnetic torque

Throughout the simulation period of 30 seconds, various aspects of the system were observed and evaluated. The obtained results are then visually represented in a series of figures, namely **Figure. (5.1)** to **Figure. (5.4)**.

5.2.2.1 DC-bus voltage regulation

The DC voltage regulator is essential for maintaining control over the voltage on the DC-bus, ensuring a balanced power exchange between the generation source and the grid. So, it is crucial to keep the voltage (V_{dc}) across the storage capacitor terminals constant. Variations in this voltage are primarily caused by the exchange of active power with the grid. It is important to minimize voltage fluctuations to avoid surpassing the voltage limit of the semiconductor switches and to preserve the active filter's performance. To fulfill the role of a DC voltage source, regulation of this voltage becomes necessary. A PI regulator is employed for the regulation of the DC voltage, assuming neglect of losses in the inverter and output filter

The capacitor current i_{dc} can be expressed by:

$$C_{dc} \frac{dV_{dc}}{dt} = i_{dc} \quad (32)$$

Laplace transform leads to the following transfer function:

$$F_{dc}(s) = \frac{V_{dc}}{i_{dc}} = \frac{1}{C_{dc}s} \quad (33)$$

The PI transfer function is given by:

$$H_{dc}(s) = \frac{k_{i_{dc}}}{s} + k_{p_{dc}} \quad (34)$$

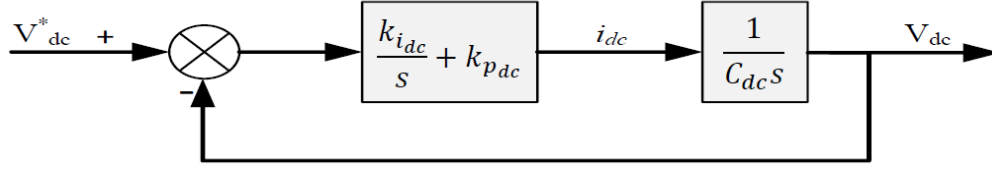


Figure (5.6) DC voltage PI regulation closed-loop

The system closed-loop transfer function as presented in figure (5.5) is given by:

$$G_{dcs}(s) = \frac{\frac{k_{i_{dc}}}{C_{dc}} + \frac{k_{p_{dc}}}{C_{dc}} s}{s^2 + \left(\frac{k_{p_{dc}}}{C_{dc}}\right)s + \frac{k_{i_{dc}}}{C_{dc}}} \quad (35)$$

Comparing the characteristic equation of (35) with that of general form of a second-order transfer function leads to the following gains:

$$\begin{aligned} k_{i_{dc}} &= C_{dc} \omega_{n_{dc}}^2 \\ k_{p_{dc}} &= 2\xi \omega_{n_{dc}} C_{dc} \end{aligned} \quad (36)$$

5.2.2.2 Current regulation

- Calculation of reference currents

The active and reactive powers provided to the grid have the following expressions:

$$\begin{cases} P_g = v_{sq} \cdot i_{fq} + v_{sd} \cdot i_{fd} \\ Q_g = v_{sq} \cdot i_{fd} - v_{sd} \cdot i_{fq} \end{cases} \quad (37)$$

By applying the VOC method by aligning the d-axis with the grid voltage ($v_{sq} = 0$, $v_{sd} = v_d$), the equation (33) can be rewritten:

$$\begin{cases} P_g = v_d \cdot i_{fd} \\ Q_g = -v_d \cdot i_{fq} \end{cases} \quad (38)$$

From this result, it can be easy to impose the powers references (P_g^* and Q_g^*) provided to the grid by imposing the following currents:

$$\begin{cases} i_{fd}^* = \frac{P_g^*}{v_d} \\ i_{fq}^* = -\frac{Q_g^*}{v_d} \end{cases} \quad (39)$$

The reference reactive power is set to zero ($Q_g=0$); the power factor on the grid side is therefore unity. The reference currents becomes:

$$\begin{cases} i_{fd}^* = \frac{P_g^*}{v_d} \\ i_{fq}^* = 0 \end{cases} \quad (40)$$

The regulation of the currents can be achieved by considering the following equation:

$$\begin{cases} v_{fd}^* = V_{sd}^* + V_{sd} - L_f \omega i_{fq} \\ v_{fq}^* = V_{sq}^* + V_{sq} + L_f \omega i_{fd} \end{cases} \quad (41)$$

Where:

$$\begin{cases} v_{sd}^* = R_f i_{fd} + L_f \frac{di_{fd}}{dt} \\ v_{sq}^* = R_f i_{fq} + L_f \frac{di_{fq}}{dt} \end{cases} \quad (42)$$

Applying the Laplace transformation, the transfer function between the controller voltages and converter currents can be derived as:

$$\frac{i_{fd}(s)}{v_{sd}^*(s)} = \frac{1}{sL_f + R_f} \quad (43)$$

$$\frac{i_{fq}(s)}{v_{sq}^*(s)} = \frac{1}{sL_f + R_f} \quad (44)$$

The PI transfer function is given by:

$$H_c(s) = \frac{k_{ic}}{s} + k_{pc} \quad (45)$$

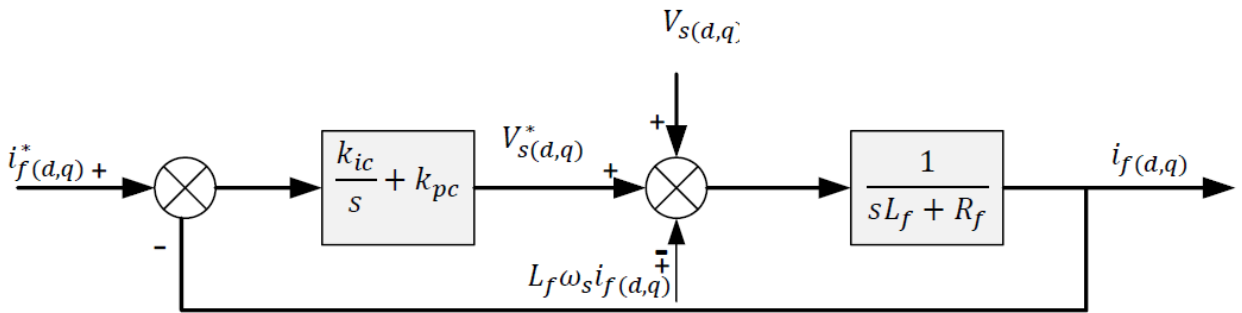


Figure (5.7) Inner current close loop control

The system closed-loop transfer function as presented in figure (5.7) is given by:

$$G_{cs}(s) = \frac{\frac{k_{ic}}{L_f} + \frac{k_{pc}}{L_f}s}{s^2 + 2\left(\frac{k_{pc} + R_f}{L_f}\right)s + \frac{k_{ic}}{L_f}} \quad (46)$$

Comparing the characteristic equation of (5.7) with that of general form of a second-order transfer function leads to the following gains:

$$\begin{aligned} k_{ic} &= L_f \omega_{nc}^2 \\ k_{pc} &= 2\xi \omega_{nc} L_f - R_f \end{aligned} \quad (47)$$

5.2.2.3 Phase Locked Loop

The Phase-Locked Loop (PLL) plays a crucial role in modern power electronics control techniques, serving as a fundamental component for retrieving phase and frequency information. It enables the determination of the angle and angular velocity of the electrical network. In a three-phase PLL, the q-axis voltage component is fed back and filtered by a Proportional-Integral (PI) controller. The output of the controller represents the angular velocity of the electrical grid, while the integration of this signal yields the grid angle [6].

The basic structure of a PLL illustrated in figure (5.6) comprises three main components: a phase detector (PD), a loop filter (LF), and a voltage-controlled oscillator (VCO). The PD compares the phase of the feedback signal with a reference signal, generating an error signal that is subsequently filtered by the LF. The filtered error signal is then used to control the frequency of the VCO, which generates an output signal with a frequency corresponding to the desired phase and angular velocity of the electrical network.

By employing a PLL, accurate phase and frequency information can be extracted, enabling precise synchronization and control of the grid-connected system. The combination of the phase detector, loop filter, and voltage-controlled oscillator facilitates the estimation of the electrical grid's angle and angular velocity, contributing to the overall stability and performance of the control system.

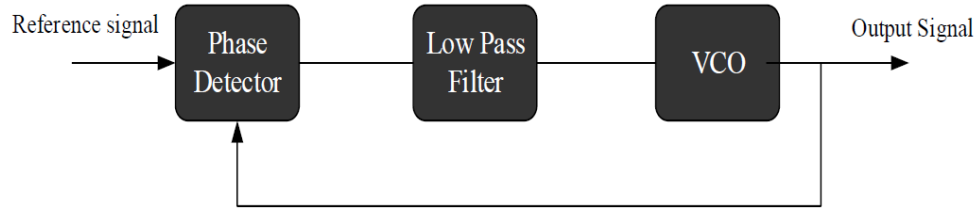


Figure (5.8) Basic form of the P.L.L

The Phase-Locked Loop (PLL) serves the purpose of generating a rotating reference angle, denoted as θ that is synchronized with the voltage signals. This synchronization is particularly valuable during grid voltage restarts, allowing the system to be seamlessly reconnected to the grid.

Various techniques for PLL implementation have been developed over time. Figure (5.6) depicts the block diagram of the classic PLL used in many applications. In this configuration, the loop filter takes the form of a Proportional-Integral (PI) controller. The choice of a PI controller is motivated by the need to strike a balance between stability and dynamic performance, ensuring robust operation and rapid response to changes.

Through the utilization of a PLL, a reliable and accurate rotating reference angle is established, facilitating the synchronization and reconnection of the system to the grid. The PI controller, as a crucial element of the loop filter, ensures a well-balanced performance with respect to stability and dynamic response, guaranteeing the effectiveness of the PLL in various power electronics applications. Its transfer function is given

$$H_{pll}(s) = k_{p_{pll}} + \frac{k_{i_{pll}}}{s} \quad (49)$$

The Voltage Controlled Oscillator VCO is an integrator.

In order to determine the parameters of the PI regulator, the block diagram of figure. (5.9) can be simplified .as it appears on the figure. (5.9).

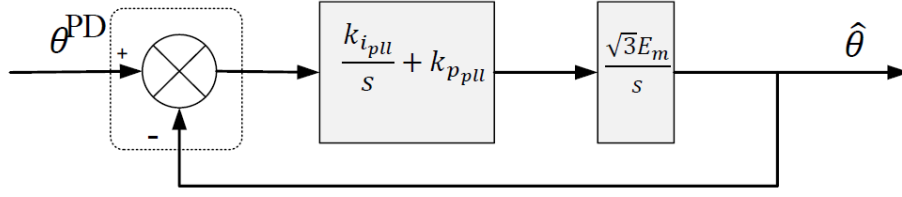


Figure (5.9) PLL closed loop control

The closed-loop transfer function of this system is given by:

$$G_{pll}(s) = \frac{\sqrt{3}E_m(k_{p_{pll}} + \frac{k_{i_{pll}}}{s})}{s + \sqrt{3}E_m(k_{p_{pll}} + \frac{k_{i_{pll}}}{s})} \quad (50)$$

By comparing the characteristic equation of (5.8) with that of a general form transfer function of a second-order leads to the following PI gains:

$$k_{i_{pll}} = \frac{\omega_{n_{pll}}^2}{\sqrt{3}E_m} \quad (51)$$

$$k_{p_{pll}} = \frac{2\xi\omega_{n_{pll}}}{\sqrt{3}E_m}$$

The following figure depicts the control of the grid side converter

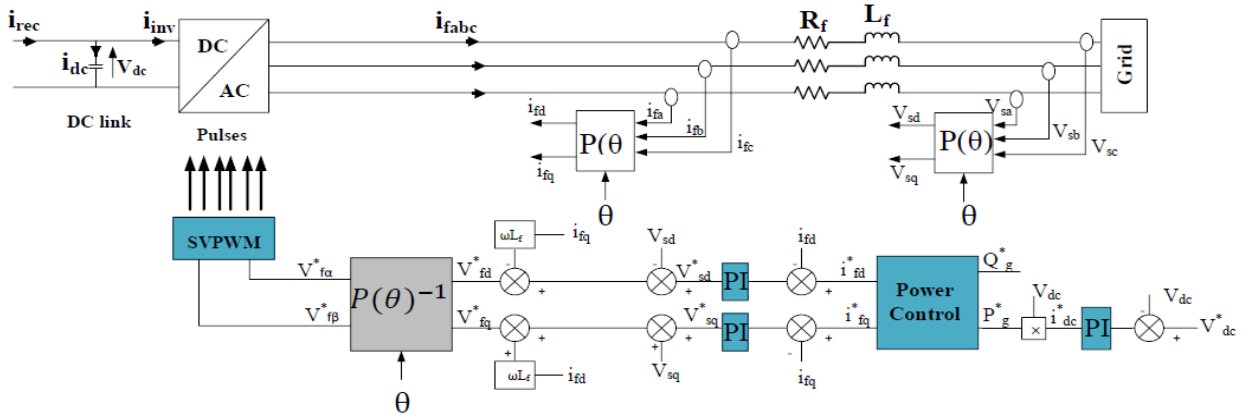


Figure (5.10) Grid side converter control by PI

5.2.3.1 Grid side Design

a) DC-Bus Voltage Choice

To determine the value of the DC-bus required to transmit a given power, the following expression is adopted:

$$V_{dc} = 1.11 \times 2\sqrt{2}E$$

Where E is the RMS voltage of the grid voltage.

b) The storage capacitor capacitance (C_{dc}) design

This approach is based on the role of the DC-bus capacitor which is to absorb or produce the power required during transient periods

$$C_{dc} = \frac{2P_{\max} 20 \cdot 10^{-5}}{V_{dc}^2 (1 - k^2)} \quad \text{With } k = \frac{V_{dc\min}}{V_{dc}} = 10\% \quad (52)$$

c) Filter Design

for a three-phase inverter the value L_f of the filter (L) is calculates as follows:

$$L_f < \frac{3V_{dc}^2}{10\omega P_n}$$

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