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*Commande par mode glissant d'une machine asynchrone
multiphasée alimentée avec des onduleurs multiniveaux*

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Dedication

First I pray for mercy upon my father's soul and ask for his eternal forgiveness. This work is dedicated to him.

I dedicate this work to the precious woman my mother, who sacrificed her youth and all her strength during these years to see me here.

Her unfailing support, consistent encouragement, and strong faith in my abilities and behaviors became a guiding light on my journey.

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List of Abbreviations

DSIM	Dual star induction machine.
NPC	Neutral Point Clamped.
THD	total harmonic distortion.
EMI	electromagnetic interference.
PWM-ST	Sinusoidal-Triangle Pulse Width Modulation.
FOC	Field-Oriented Control.
PI	Proportional-Integral Controller.
SMC	Sliding Mode Control.
VSS	Variable Structure System.
CM	Convergence Mode.
SSM	Steady-State Mode.
As1, Bs1, Cs1	The phases of the first star of the DSIM.
As2, Bs2, Cs2	the phases of the second star of the DSIM.
α	The angular displacement between phase windings.
Ω_s	Synchronous mechanical speed (in rad/s).
ω_s	Angular frequency of stator supply (in rad/s).
P	Number of pole pairs.
g	The slip of the machine.
ω_s	Angular frequency of the stator magnetic field (rad/s).
ω_r	Angular speed of the rotor (rad/s).
θ_0	Initial rotor angle with respect to stator 1
Ω_m	Mechanical rotor speed.
B_{xi}	Transistor basic control S_{xi} of an arm x , $x = a, b$.
$V_{sa1}, V_{sb1}, V_{sc1}$	Phase voltages of the stator winding 1 (phases a, b, and c).
$V_{sa2}, V_{sb2}, V_{sc2}$	Phase voltages of the stator winding 2.
$I_{sa1}, I_{sb1}, I_{sc1}$	Phase currents of stator winding 1.
$I_{sa2}, I_{sb2}, I_{sc2}$	Phase currents of stator winding 2.
I_{ra}, I_{rb}, I_{rc}	Rotor phase currents.
$\phi_{sa1}, \phi_{sb1}, \phi_{sc1}$	Magnetic flux linkages in stator winding 1.
$\phi_{sa2}, \phi_{sb2}, \phi_{sc2}$	Magnetic flux linkages in stator winding 2.
$\phi_{ra}, \phi_{rb}, \phi_{rc}$	Rotor flux linkages.

$R_{sa1}, R_{sb1}, R_{sc1}$	Phase resistances of stator winding 1.
$R_{sa2}, R_{sb2}, R_{sc2}$	Phase resistances of stator winding 2.
R_{ra}, R_{rb}, R_{rc}	Rotor resistances.
ϕ_{ds}, ϕ_{qs}	Stator Flux Components along the d and q axes
ϕ_{dr}, ϕ_{qr}	Rotor Flux Components along the d and q axes
I_{ds}, I_{qs}	Stator current along the d and q axes.
V_{ds}, V_{qs}	Stator voltage along the d and q axes.
$L_{s1s1}, L_{s2s2}, L_{rr}$	Inductance matrix of star 1, star 2 and rotor.
L_{s1s2}, L_{s1r}	Mutual inductance matrix between star 1 and star 2, star 1 and rotor.
L_{s2s1}, L_{s2r}	Mutual inductance matrix between star 2 and star 1, star 2 and rotor.
L_{rs1}, L_{rs2}	Mutual inductance matrix between rotor and star 1, rotor and star 2.
L_{s1}, L_{s2}, L_r	Cyclic self-inductance of 1st star and 2nd star and the rotor.
M_{sr}	The maximum value of mutual inductance coefficients between a star and the rotor.
ω_{mag}	Magnetic energy.
θ_m	Mechanical rotor angle.
θ_e	Electrical rotor angle.
θ	Electrical angle between stator and rotor axes.
θ_a	Rotor angle in vectorial (rotating) reference frame.
θ_r	Rotor electrical position.
θ_{coord}	The angle formed by the axes A_{s1} - d.
θ_{rcoord}	The angle formed by the axes A_r , d.
ω_{coord}	The speed of rotation of (d, q) relative to star 1.
ω_{rcoord}	The speed of rotation of frame (d, q) relative to the rotor.
T_{em}	Electromagnetic torque produced by the motor.
T_L	Load (resistant) torque.
Ω	Mechanical angular speed.
J	Moment of inertia.
ω_a	The electrical angular of the UV axis system.
ω_m	Electrical angular of the rotor.
$[P(\theta_a)]$	Park transformation matrix.
$L_{s1} + L_m$	cyclic self-inductance of star 1.
$L_{s2} + L_m$	The cyclic self-inductance of star 2.
$L_r + L_m$	The rotor's cyclic self-inductance.
P_{em}	Electromagnetic power and torque .

ω_{gl}	Slip angular frequency.
ω_s	Synchronous electrical angular speed of the stator field.
ω_r	Rotor electrical angular speed.
ϕ_f	Magnetic flux generated by the excitation current I_f .
k', k	Fixed parameters.
ϕ_{rref}	The rotor flux reference.
ϕ_n	Nominal (rated) rotor flux (maximum flux allowed).
Ω_n	Nominal (rated) speed.
k_{pv}	Proportional gain.
k_{iv}	Integral gain.
C	Desired closed-loop time constant.
C_{emref}	Reference electromagnetic torque.
C_{emmax}	Maximum allowed electromagnetic torque.
C_{ij}	Sliding surface parameters.
C	Matrix of $n \times m$ surface parameters (slope matrix of the surface line).
λ	A positive constant.
r	Relative degree, equal to the number of times you need to derive the output to make the command appear.

General introduction

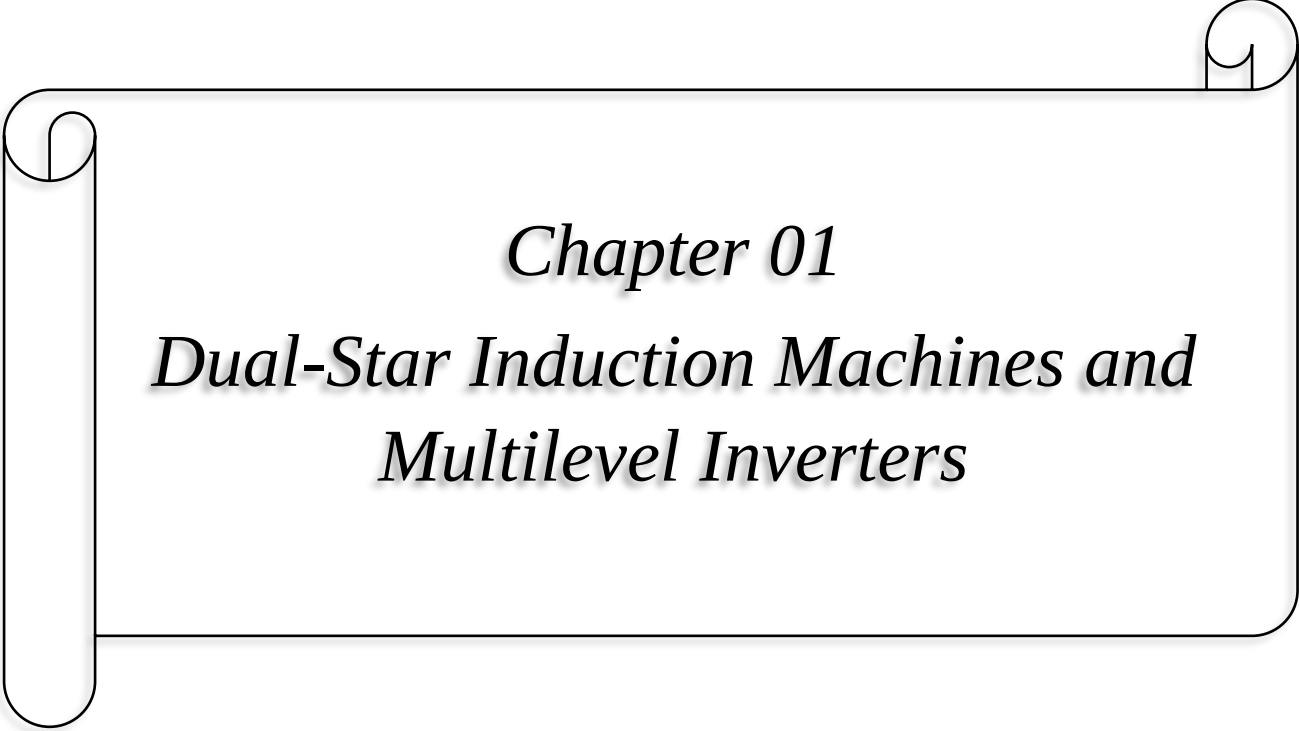
The increasing demand for high-performance and reliable electric drive systems in industrial, automotive, and renewable energy applications has motivated the innovation of new machine architecture and control methods. Among these machines, the Dual Star Induction Machine (DSIM) has received considerable interest thanks to its improved torque operation, higher fault tolerance, and enhanced thermal behavior when compared to traditional induction machines. Having two space-shifted stator windings on a common rotor, the DSIM is suitably designed for safety-critical and high-power applications.

Control of a Distributed Static Induction Machine (DSIM) presents new challenges because of its higher complexity and inter-winding effects. To meet such challenges, Sliding Mode Control (SMC) provides a robust nonlinear control strategy which guarantees the system performance despite uncertainties and disturbances. SMC is best suited to preserve both system stability and dynamic performance; however, it requires careful design in order to reduce the chattering effect normally exhibited by this control technique.

In parallel, the application of Three-Level Neutral Point Clamped (NPC) inverters offers enhanced voltage quality and less harmonic distortion over its conventional two-level counterpart. Therefore, such inverters are best suited to drive Double-Sided Induction Motors (DSIMs) especially where precision and efficiency are of utmost importance.

To simulate and evaluate the proposed control method, a comprehensive MATLAB/Simulink model was used, which included the DSIM, three-level NPC inverter, sliding mode control, and modulation. This simulation is required for assessing system behavior and verifying control performance under diverse scenarios.

The purpose of this work is to establish robust, high-performance control of the DSIM utilizing SMC and a three-level inverter, as demonstrated by simulation results that show enhanced dynamic behavior, efficiency, and resilience to disturbances. This contributes to the creation of intelligent, dependable driving systems for next-generation applications.

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Chapter 01

Dual-Star Induction Machines and Multilevel Inverters

Part one I: Dual-Star Induction Machines

1.1.1. Introduction

Due to its straightforward design, small size, great efficiency, and remarkable dependability, the induction motor continues to be the most popular electrical device in industrial settings. The magnetic circuit, which constantly needs reactive power to magnetize the air gap, is one of its intrinsic drawbacks.

It is commonly acknowledged that three-phase induction motors account for between 40% and 50% of industrial energy consumption and about 70% of all installed motor power. The importance of induction machines in contemporary industry is amply demonstrated by these estimates, even though they may differ based on the source.

1.1.2. Characteristics of multiphase machines

Multiphase machines can be generally divided into two groups according to the number of stator phases: those whose phase number is a multiple of three and those that lack this property. for any number of phases, there exists a variety of winding configurations, depending on the angular displacement α which can be between adjacent phase windings, frequently also characterized by the expression phase shift between the stars. For example, a dual-star (six-phase) machine having an angle α of 0 degrees has different electromagnetic and operating characteristics than one for which α is set at 30 degrees. In order to rectify these discrepancies and to differentiate among the various possible configurations, the notion of the equivalent number of phases is presented. This is defined as [1]:

$$n_{ph} = n_{ph\alpha} = \frac{180}{\alpha} \quad (1.1)$$

α : the angular displacement between phase windings.

1.1.3. Advantages of multiphase machines

Multiphase machines are more advantageous than conventional machines. These advantages include: power segmentation, reliability, minimization of torque ripples and rotor losses.

1. Power partitioning

By increasing the number of phases, power is automatically increased. One way of reducing phase currents without reducing supply voltages is to increase the number of stator phases. The total power required by a machine is then reduced in each phase. With this power, the machine can be supplied by an inverter whose lower-grade semiconductor components can operate at higher switching frequencies. This minimizes current and torque ripples. Power segmentation is the main advantage of multiphase machines, and the one that is being promoted the most these days.

2. Reliability

Degraded operation occurs when one phase fails in the inverter, causing loss of machine control and high torque ripples. A solution is to connect the machine's neutral to the midpoint of the DC source. In multiphase machines, keeping at least three phases active allows for control without this connection, providing more freedom with additional phases.

3. Minimizing torque ripple and rotor losses

Electromagnetic torque ripple, occurring at six times the fundamental frequency, is caused by five- and seven-time harmonics. These harmonics are present in three-phase machines but are removed in double-star machines. Multiphase machines also eliminate low-rank torque harmonics, leading to lower rotor losses compared to three-phase machines. [2]

1.1.4. Disadvantages of multiphase machines

The number of semiconductors increases with the number of phases, which may eventually increase the cost of the converter-machine assembly. However, as power increases, this problem becomes less significant.

Multiplying the number of semiconductors obviously complicates the control system. It is therefore necessary to develop specific and adapted close control techniques (static converter control).

The dual-star machine is the most common multiphase machine, no doubt because it represents a good compromise between sufficient power segmentation and a converter-machine assembly that's not too complicated and reasonably priced.

1.1.5. Dual star induction machine DSIM

1.1.5.1. Main components of the DSIM

- Fixed part (stator or inductor)

It features two offset stators, each composed of three identical windings, at an angle $\alpha = 30^\circ$. An electrical angle of $2\pi/3$ separates their axes in space. They are logged into the notches of the electrical circuit.

A three-phase current equilibrium system powers each of the two stator mechanical components, forming a rotating chamber that glides into the center. The pulsation of the statorical currents and the number of pairs of poles on the machine are inversely correlated he synchronous speed of the rotating magnetic field in the stator as:

$$\Omega_s = \frac{\omega_s}{P} \quad (1.2)$$

- Ω_s : Synchronous mechanical speed (in rad/s)
 - ω_s : Angular frequency of stator supply (in rad/s)
 - P : Number of pole pairs
- Moving part (rotor or armature)

The rotor is constructed in such a way as to obtain three windings having a number of pairs of poles identical to that of the stator.

The electrical structure of the rotor is assumed to be a squirrel cage rotor consisting of conductor bars short-circuited by a conducting ring at each end (ferromagnetic aluminum conductor bar)

This choice results in machines that are inexpensive, robust, easy to use and require limited maintenance. The rotor rotates at a speed ω_r different from ω_s for this reason, the rotor cage becomes the seat of a system of three-phase electromotive forces generating themselves three rotational currents. Thus the effects of stator induction on rotor induced currents are manifested by the development of a couple of electromagnetic forces on the rotor such that the difference in speeds is reduced by the slip of the machine: [3]

$$g = \frac{\omega_s - \omega_r}{\omega_s} \quad (1.3)$$

- ω_s : Angular frequency of the stator magnetic field (rad/s).
- ω_r : Angular speed of the rotor (rad/s).

1.1.5.2. Description of the DSIM

The dual-star induction machine has two three-phase winding systems in the stator, offset from each other by an electrical angle α (in this model we take $\alpha = 30^\circ$), and either a wound rotor or a squirrel-cage rotor. To simplify the study, we consider the rotor's electrical circuits to be equivalent to a short-circuited three-phase winding. Figure 1.2 shows the position of the winding axes of the nine phases making up the machine. Six phases for the stator and three phases for the rotor, values for the first star (stator 1) are given the subscript s1, while those for the second star (stator 2) are given the subscript s2. The phases of the first star A_{s1}, B_{s1}, C_{s1} and the phases of the second star take A_{s2}, B_{s2}, C_{s2} , the rotor phases are noted by A_r, B_r, C_r . The offset angle between the two stars is α . θ_m expresses the position of the rotor (A_r phase) in relation to star1 (A_{s1} phase) and θ_2 the position of the rotor in relation to star2, these angles are defined by the following equations:[4]

$$\theta_m = \Omega_m * t + \theta_0 \quad (1.4)$$

$$\theta_2 = \theta_m + \alpha \quad (1.5)$$

- θ_0 : Initial rotor angle with respect to stator 1 at time $t = 0$.
- Ω_m [rad/s]: mechanical rotor speed.

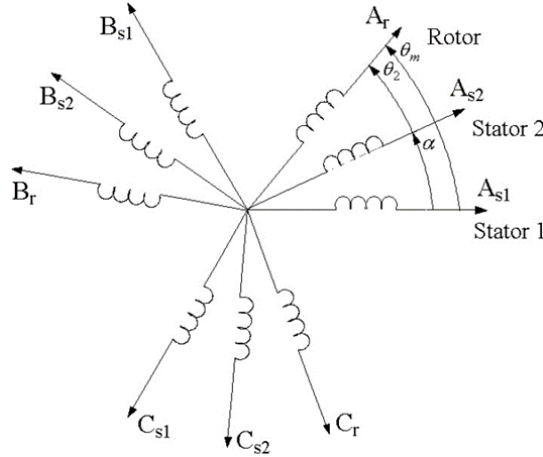


Figure 1.1: Schematic Representation of the Windings of the Dual-Star Induction Machine (DSIM)

1.1.5.3. Operating principle of the DSIM

The three-phase currents of frequency f_s feeding the machine's stator winding 1 give rise to a rotating field at synchronous speed N_s , such that:

$$N_s = \frac{f_s}{P} * 60 \quad (1.6)$$

Where:

- P : The number of pole pairs.
- f_s : Stator supply frequency

The same three-phase currents, offset by an angle α , feeding winding 2 of the same stator also give rise to another field rotating at the same speed of synchronism N_s . These two rotating fields produced by the two stator windings will induce currents in the rotor conductors, generating electromotive forces that will cause the rotor to rotate at a speed N [rpm] lower than that of synchronism ($N < N_s$). The effects of stator induction on rotor induced currents are thus manifested by the development of an electromagnetic force torque on the rotor such that the speed difference is reduced. [6][5]

1.1.5.4. Application of the DSIM

Multi-phase machines are used much more in high power applications. These applications include marine propulsion, fans, compressors, compressor mills, cement mills, etc.

Electric machines with a large number of phases are increasingly used because they operate in degraded mode. Already after the First World War, in order to increase the power of alternators, this machine with a large number of phases replaced conventional alternators.

Today, the multi-phase machine is used in railway traction and the marine, avionics and aerospace industries thus offering additional safety in terms of operation, torque, vibration and noise. In industry, these machines are present as fans, compressors, compressor mills, cement mills and pumps.

1.1.6. Conclusion

This part presented an overview of multiphase machines, focusing their essential features, benefits, and limits. The Dual Star Induction Machine (DSIM) received special attention, including its structure, essential components, and operating principles. The DSIM design demonstrates how multiphase systems can improve reliability, power density, and performance in advanced electric drive applications.

Part two: multi-level Inverters

1.2.1. Introduction

Inverters play a core role in the area of power electronics, appearing in a broad spectrum of application areas. Their continual development has been achievable, on the one hand, due to progress in the technology of the manufacture of fully controllable high-power, long-lived, high-speed semiconductor devices; and on the other hand, due to extensive use of "pulse-width modulation" methods. The latter are dependent on the potentialities of the switching frequency achievable by the former.

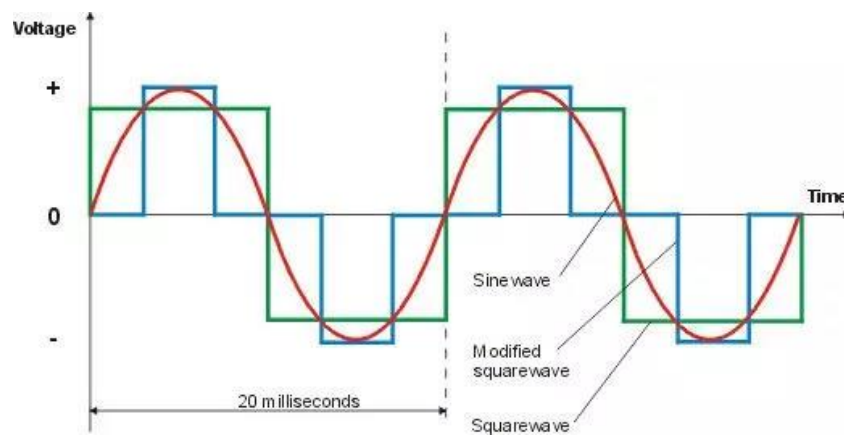


Figure 1.2: Inverter Output Waveform Comparison

1.2.2. Multilevel Invertes

In the early 1990s, the improvement of the characteristics of power semiconductors drove the expansion of the converter market, utilized for medium-voltage and medium- and high-power applications, with power levels from a few kilowatts to several megawatts in the kilovolt range.

Noting an increase in the levels of power, voltage, and current of the converters brought through an increase in the number of conversions and modification of new power components (MOSFETs; GTOs; IGBTs; IGCTs): further evolution of new multilevel conversion structure, basic principle is a common parallel/series combination of power components. [6]

1.2.3. Interest in Multilevel Inverters

Multilevel inverters are developed for two main reasons: higher power generation and greater definition of harmonic content in output waveforms. They have a switched voltage limited to the converter's step value, reducing switching losses. The switching frequency of components is lower than the load voltage's, allowing for higher overall switching frequency and reducing filter

size or signal quality. Modular converters can be created using replica modules, with the number of modules determined by the converter's operating voltages. [7]

1.2.4. Diode-Clamped Multilevel Inverter NPC

The diode-clamped multilevel inverter, also known as the Neutral Point Clamped (NPC) inverter, is a common topology for multilevel power conversion. It uses diodes to clamp DC bus voltage, creating additional voltage steps of output. In a three-level inverter configuration, each switch sees only half of the DC bus voltage, improving efficiency and protection. However, practical implementations are limited to three-level inverters due to the difficulty of balancing capacitor voltages and the large number of parts. PWM control also brings complications like diode reverse recovery, complicating the design process. Despite these challenges, the three-level diode clamped inverter is common in medium-voltage industrial applications.

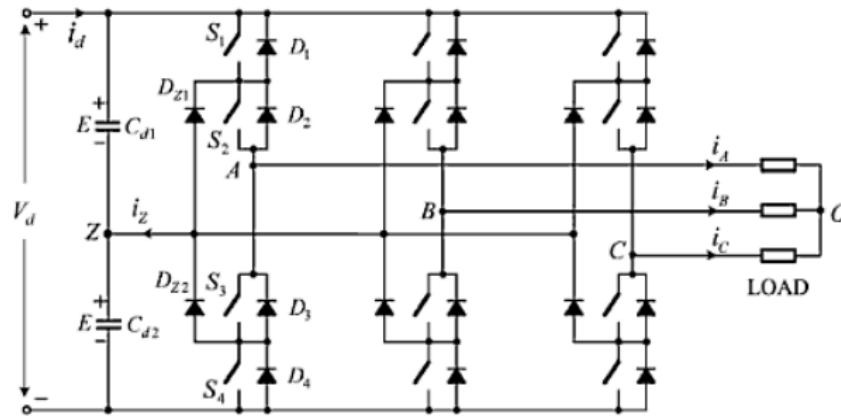


Figure 1.3: Three-level diode clamped inverter. [8]

The diode-clamped multilevel inverter is a complex structure that operates on known principles. Each phase can connect with various capacitor nodes when switches are activated in different combinations. This is known as complementary switching, with a small dead-time to avoid short circuiting. A four-level inverter adds more diodes per phase but applies the same principle. After defining switching states, gate signals for transistors are translated, allowing for expressions for phase voltages and DC currents. [10] [9]

1.2.5. Advantages and disadvantages of multilevel inverters

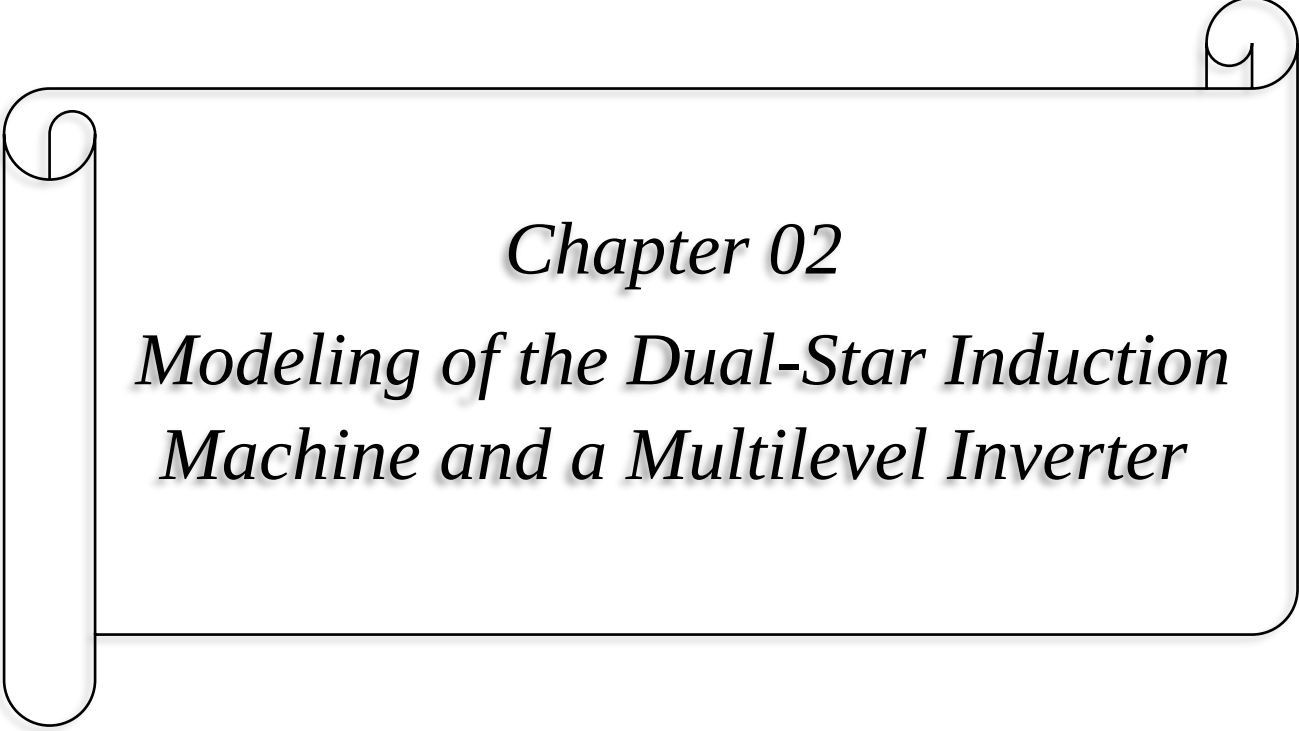
Multilevel inverters exhibit several advantages that make them highly suitable for medium and high voltage applications. They produce output waveforms with significantly reduced total harmonic distortion (THD), which is very close to a sinusoidal shape, thus improving power

quality. Furthermore, the use of multiple voltage levels helps to reduce switching losses and electromagnetic interference (EMI), leading to better overall efficiency and reduced stress on power switches. In addition, voltage sharing between several semiconductor devices allows the use of lower-rated devices in high-voltage systems for enhancing system reliability. These inverters are also highly scalable and flexible to be used in applications such as industrial drives, renewable energy systems. [11]

Multilevel inverters also possess several drawbacks. The higher quantity of power switches and control circuits rises the complexity and price of the system. The utilization of sophisticated modulation techniques, i.e., carrier-based or space vector pulse-width modulation, is more complex with the rise in the number of levels. In capacitor-based topologies, voltage balance among components can be challenging to uphold and might necessitate supplementary control methods. Furthermore, the size of multilevel inverters in general would be larger because of the greater number of components, and this can be a disadvantage in applications where limited space is a concern. [12]

1.2.6. Conclusion

The evolution of multilevel inverters is a great progress in the field of power electronics with greater functionality compared to the conventional two-level inverters. With their capability to process high power with improved output quality, lower harmonic distortion, and higher efficiency, they are essential for applications including renewable energy systems, electric vehicles, and industrial motor drives. Different topologies, such as NPC, floating capacitor, and H-bridge, offer solutions to issues such as voltage balancing and circuit complexity. With advancing technology, these inverters will persist in pushing the limits of energy conversion and system reliability.

A decorative frame resembling a scroll, with a vertical bar on the left and a horizontal bar at the top, both ending in scroll-like curls.

Chapter 02

Modeling of the Dual-Star Induction Machine and a Multilevel Inverter

Part one: multi-level inverter modulation

2.1.1. Introduction

The need for high-efficiency power conversion systems has led to the development of new inverter topologies and control techniques, with multilevel inverters being a potential solution. The modulation technique of these inverters directly impacts their performance, efficiency, reliability, and electromagnetic compatibility. This chapter discusses various pulse width modulation methods, including vector modulation, their principles, features, implementation methods, and practical benefits.

2.1.2. Structure and operating principle of the three-level NPC inverter

The three-level NPC inverter in figure (2.1) , has a DC bus that comprises two capacitors (C_1 and C_2) in series forming a midpoint noted "O" afforded the inverter one extra level of voltage to reference compared to the regular two-level inverter. The total voltage of the DC bus is E ; under normal operation, this condition distributes the voltage over the two capacitors evenly which operate at $E/2$ across their terminals. Each of the three arms of the inverter consists of four controlled switches (K_1, K_2, K_3 and K_4 , in arm a) and two holding diodes connected to the midpoint of the DC bus. The controlled switches are unidirectional in voltage and bidirectional in current; a transistor to a diode connected in antiparallel. [13]

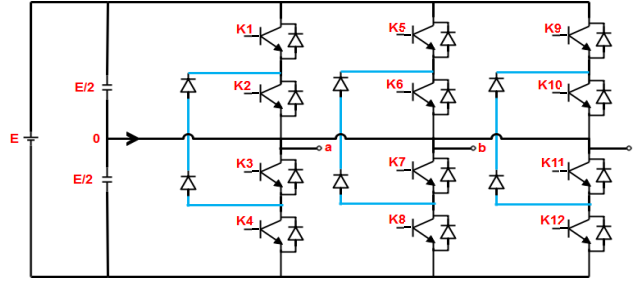


Figure 2.1: 3-level NPC inverter schematic.

2.1.3. Three-level inverter model

The connection function of the switch is defined first. The connection function of the switch equals 1 if the switch is closed or equals 0 otherwise. In controllable mode, the connection functions of the inverter are related by: [14] [15]

$$\begin{cases} F_{K1} = 1 - F_{K4} \\ F_{K2} = 1 - F_{K3} \end{cases} \quad (2.1)$$

And $k=1, 2$ or 3 , represents the arm number.

The connection function F_m of the half-bottom is defined as follows:

$$\begin{cases} F_{K1} = F_{K1} * F_{K2} \\ F_{K0} = F_{K3} * F_{K4} \end{cases} \quad (2.2)$$

We define $m=1$ as the top half-bottom and $m=0$ as the bottom half-bottom. Nodes A, B, and C of the three-phase, three-level inverter have the following potentials in relation to point M:

$$\begin{cases} V_{AM} = F_{11} * U_{c1} - F_{10} * U_{c2} \\ V_{BM} = F_{21} * U_{c1} - F_{20} * U_{c2} \\ V_{CM} = F_{31} * U_{c1} - F_{30} * U_{c2} \end{cases} \quad (2.3)$$

The simple output voltages are:

$$\begin{bmatrix} V_A \\ V_{BM} \\ V_{CM} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} * \left(\begin{bmatrix} F_{11} \\ F_{21} \\ F_{31} \end{bmatrix} U_{c1} - \begin{bmatrix} F_{10} \\ F_{20} \\ F_{30} \end{bmatrix} U_{c2} \right) \quad (2.4)$$

2.1.3 .1. Advantages and disadvantages

▪ Advantages of NPC 3 level Inverters

- **More Voltage Levels:** NPC inverters have one more voltage level compared to standard two-level inverters. The output waveforms tend to be richer in sine wave characteristics and have smaller harmonics.
- **Improved Static and Dynamic voltage:** NPC inverters have a better static and dynamic voltage because working with series connected devices enhances voltage performance.
- **Improved Spectral Performance:** The output waveform quality is better than that of normal two-level inverters.
- **Better Overvoltage Constraints with Limited Overvoltage Problems:** NPC inverters have natural overvoltage constraints.

▪ Disadvantages of NPC Inverters

- **Complexity:** NPC inverters are intrinsically more complex than two-level inverters, which can lead to higher system cost and control complexity.
- **Diodes:** The use of clamping diodes can cause power losses and reduce efficiency.
- **Applications:** NPC inverters are more suitable for higher power applications, and the use of NPC will be less efficient for low power systems. [16]

2.1.4. Sine-wave PWM inverter control technology

2.1.4.1. High-Frequency Modulation

The use of control techniques at higher switching frequencies causes the switching harmonics to shift to higher, more easily filterable frequencies, thus enhancing voltage spectrum quality. These methods include hysteresis control, pulse width modulation (PWM), harmonic injection control, and space vector PWM.

These methods not only regulate the fundamental component of the AC-side voltage, but they also handle the full low-frequency range. In the next section of this chapter, we will look at the following control method:

➤ Pulse Width Modulation Control.[17]

2.1.4.2. Pulse Width Modulation (PWM)

Pulse Width Modulation, or PWM, is a technique used to control static converters that connect a load, like an electric motor, to its power source, such as a three-phase inverter. This method is commonly utilized for converting energy, PWM is essential for regulating speed, which significantly affects the performance of the entire system, including drive efficiency, losses in the inverter and machine, and the amount of noise produced.

This method enables the creation of needed voltages or currents at the load's terminals from a power source that has a constant frequency and voltage, usually a DC voltage, by utilizing a direct converter that links the source to the load, control is achieved by modifying the durations when the semiconductor switches are turned on or off, along with their operational sequences.

2.1.4.3. Pulse Width Modulation principle

Sinusoidal Pulse Width Modulation (SPWM) is a method for controlling inverter switching by comparing a low-frequency sinusoidal reference signal to a high-frequency triangular carrier wave. In a three-phase system, three phase-shifted reference signals are employed. The switching instants are determined by the positions at which the reference and carrier waves intersect. This approach regulates the output voltage frequency and reduces harmonic distortion, resulting in improved overall performance and efficiency. [18][19]

$$\begin{cases} V_{p1}(t) = \frac{2}{\pi} V_{mp} \cdot \arcsin(\cos(2\pi f_p \cdot t)) \\ V_{p2}(t) = \frac{2}{\pi} V_{mp} \cdot \arcsin(\cos(2\pi f_p \cdot t) - \frac{2\pi}{3}) \end{cases} \quad (2.5)$$

Sinusoidal-triangular modulation is also known as intersective pulse width modulation because it works by intersecting a low-frequency modulating signal, known as the reference voltage, which is typically sinusoidal, with a high-frequency carrier wave that is triangular in shape.

The comparison of these two signals determines how the power circuit switches are turned on and off and to generate the SPWM control pulses for the three-level voltage converter, two triangle carrier signals with the same amplitude and frequency are needed. These carriers are then compared to a sinusoidal reference signal with a specific amplitude and frequency. Each comparison returns a logic '1' if the carrier is greater than or equal to the reference and a '0' otherwise.[20]

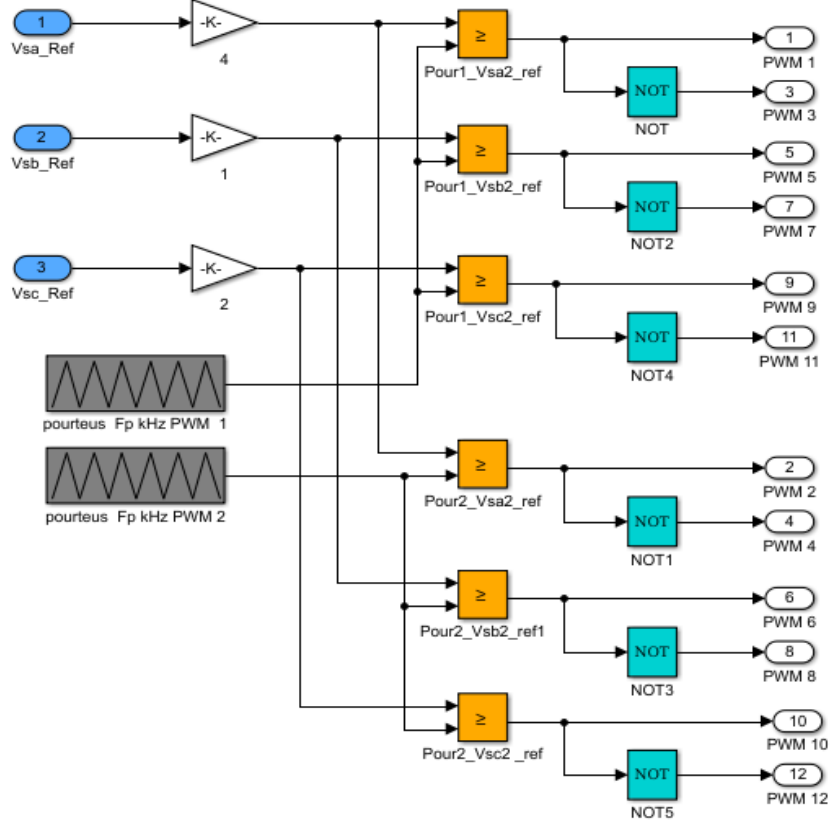


Figure 2.2: PWM Control of Three-Level Inverter – MATLAB/Simulink Model

2.1.5. Voltage inverter modeling [21] [19]

Inverters are static power converters that convert direct current (DC) into alternating current (AC) and are widely utilized in industrial applications such as variable-speed drives and uninterruptible power supplies. While two-level inverters are simple, they struggle to generate a pure sine wave, resulting in harmonic distortion. The control approach adopted has a significant impact on the output waveform's quality.

To boost performance, this study uses a three-level Neutral Point Clamped (NPC) inverter. This design uses two series switches per leg, each consisting of a transistor and an anti-parallel diode. It can generate three voltages resulting in improved waveform quality and decreased harmonics. Advances in power electronics and control systems have increased the efficiency and dependability of these inverters.

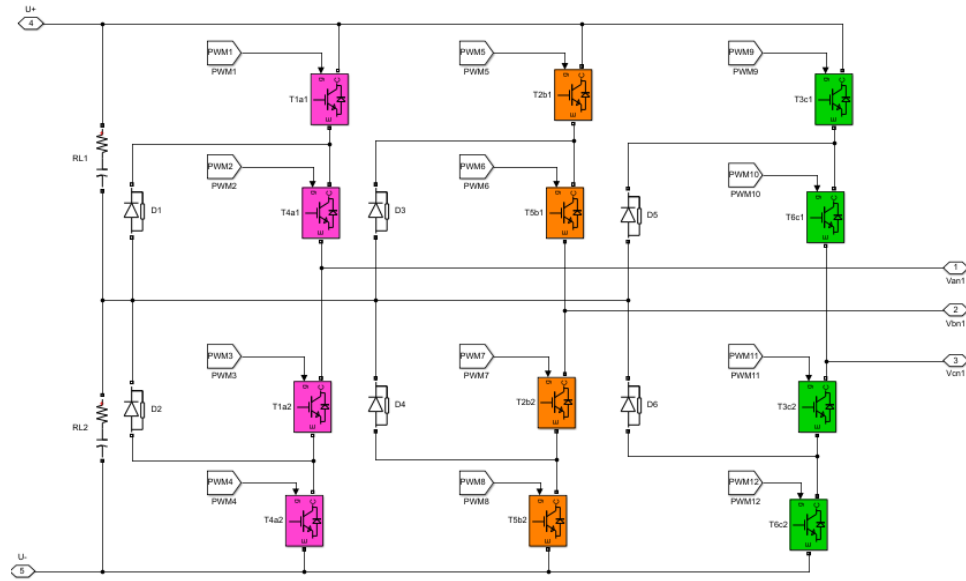


Figure 2.3: Simulink model of a three level inverter

The inverter's information effort is intended to be high. As a result, regardless of the current carried by this supply, the voltage at its terminals remains constant $U_{c1} = U_{c2} = U_{c3}$. The voltage loss at the semiconductor terminals is insignificant before the reserve voltage, which is typically two or three hundred volts. An additional order is taken to avoid short-circuiting the voltage sources due to the conduction of a few switches, as well as to make the converter fully controlled.

$$\begin{cases} B_{x1} = \overline{B_{x4}} \\ B_{x2} = \overline{B_{x3}} \end{cases} \quad (2.6)$$

And :

- **B_{xi}** : Transistor basic control S_{xi} of an arm x , $x = a, b$ and c ; $i = 1, 2, 3$ and 4 .

The connection functions of the three-level inverter arms are given by:

$$\begin{cases} F_{x1}^b = F_{x1} \cdot F_{x2} \\ F_{x0}^b = F_{x3} \cdot F_{x4} \end{cases} \quad (2.7)$$

- **X** : Arm number and $x=a, b$ et c ;

The following system determines the potentials of nodes a, b, and c of the three-phase, three-level inverter relative to point N of the input voltage source:

$$\begin{cases} V_{an} = F_{11} \cdot F_{12} U_{C1} - F_{13} F_{14} \cdot U_{C2} = (F_{11} \cdot F_{12} - F_{13} F_{14}) U_c \\ V_{bn} = F_{21} \cdot F_{22} U_{C1} - F_{23} F_{24} \cdot U_{C2} = (F_{21} \cdot F_{22} - F_{23} F_{24}) U_c \\ V_{cn} = F_{31} \cdot F_{32} U_{C1} - F_{33} F_{34} \cdot U_{C2} = (F_{31} \cdot F_{32} - F_{33} F_{34}) U_c \end{cases} \quad (2.8)$$

The three-level voltage inverter is equivalent to two two-level inverters in series.

$$\begin{cases} F_{11}^b = F_{11} \cdot F_{12} \\ F_{10}^b = F_{13} \cdot F_{14} \\ F_{21}^b = F_{21} \cdot F_{22} \\ F_{20}^b = F_{23} \cdot F_{24} \\ F_{31}^b = F_{31} \cdot F_{32} \\ F_{30}^b = F_{33} \cdot F_{34} \end{cases} \quad (2.9)$$

Composite voltages are given by:

$$\begin{cases} V_{ab} = V_{an} - V_{bn} = (F_{11} \cdot F_{12} - F_{21} \cdot F_{22}) U_{C1} - (F_{13} F_{14} - F_{23} F_{24}) \cdot U_{C2} \\ V_{bc} = V_{bn} - V_{cn} = (F_{21} \cdot F_{22} - F_{31} \cdot F_{32}) U_{C1} - (F_{23} F_{24} - F_{33} F_{34}) \cdot U_{C2} \\ V_{ca} = V_{cn} - V_{an} = (F_{31} \cdot F_{32} - F_{11} \cdot F_{12}) U_{C1} - (F_{33} F_{34} - F_{13} F_{14}) \cdot U_{C2} \end{cases} \quad (2.11)$$

If we assume that the single voltages of the receiver form a balanced three-phase system, we have:

$$\begin{cases} V_{an} = \frac{V_{ab} - V_{ca}}{3} \\ V_{bn} = \frac{V_{bc} - V_{ab}}{3} \\ V_{cn} = \frac{V_{ca} - V_{bc}}{3} \end{cases} \quad (2.12)$$

By introducing the connection functions of the half-lows, we obtain:

$$\begin{bmatrix} U_{ab} \\ U_{bc} \\ U_{ca} \end{bmatrix} = \begin{bmatrix} 1 & -1 & 0 \\ 0 & 1 & -1 \\ -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} F_{11}^b \\ F_{21}^b \\ F_{31}^b \end{bmatrix} U_{c1} - \begin{bmatrix} F_{10}^b \\ F_{210}^b \\ F_{30}^b \end{bmatrix} U_{c2} \quad (2.13)$$

The simple voltages are given as follows:

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = \frac{1}{3} U_c \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} F_{11}^b - F_{10}^b \\ F_{21}^b - F_{20}^b \\ F_{31}^b - F_{30}^b \end{bmatrix} \quad (2.14)$$

2.1.6. Numerical Simulation and Analysis of Results

After modeling the PWM-ST control of the three-level voltage inverter, we get the waveforms below:

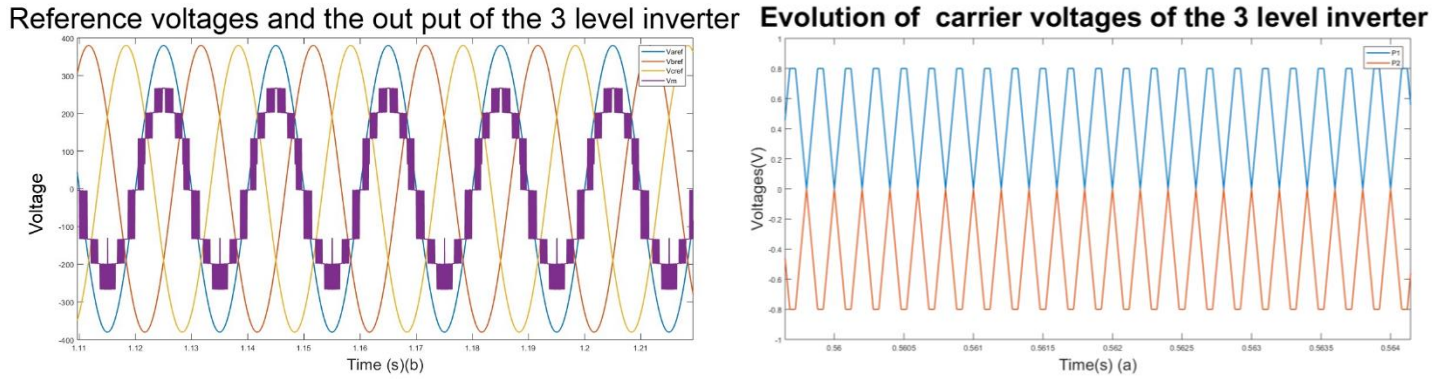


Figure 2.4: Voltage Signals of the Three-Level Inverter: Reference, Carrier, and Output

The simulation results confirm the success of the PWM modulation technique in synthesizing improved output voltages with the three-level inverter topology. The phase output voltage successfully tracks the sinusoidal reference waveforms (± 380 V peak) through the use of discrete levels of +200 V, 0 V, and -200 V, confirming proper implementation of the multilevel approach.

The technique decreases voltage stress while improving harmonic performance simultaneously. The 2 kHz carrier waves, symmetrically centered at ± 0.75 V, ensure secure and synchronized switching. In general, PWM enables clean voltage synthesis with improved spectral quality and is hence suitable for medium-voltage and high-performance applications.

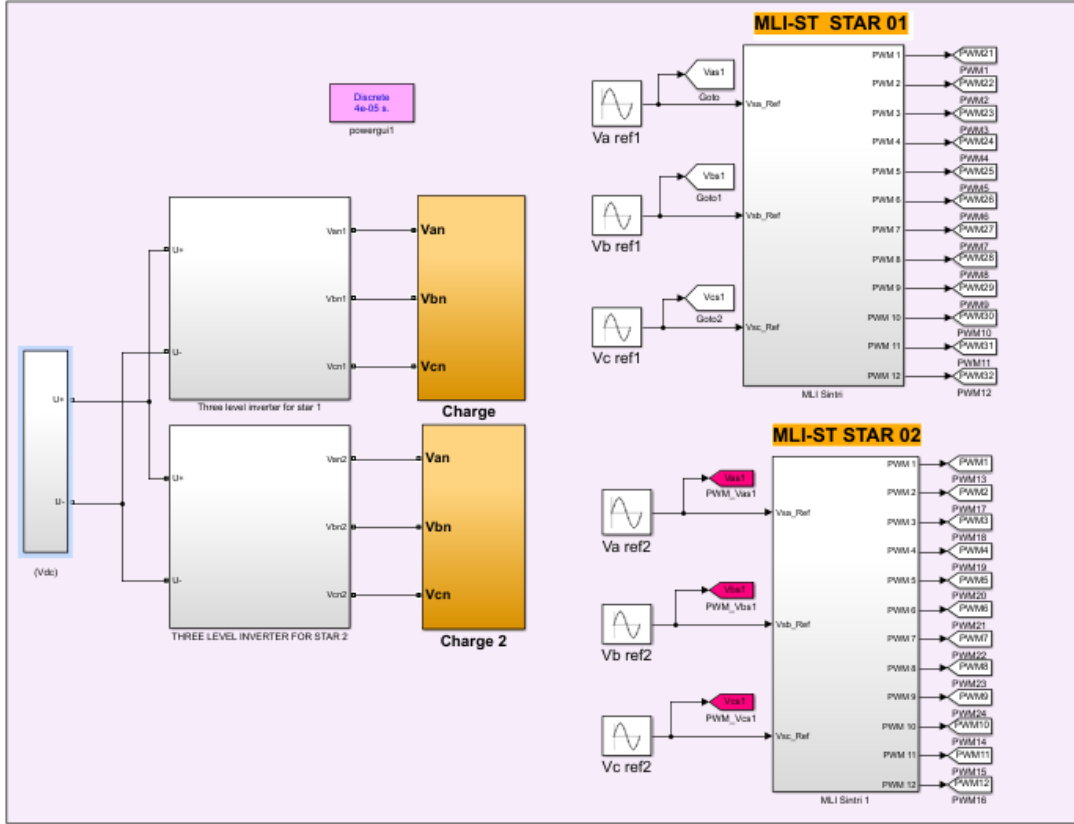


Figure 2.5: Matlab /simulink block of three-level inverter with PWM-ST control

2.1.7. Conclusion

The control strategy in a multilevel inverter plays a key role in determining its operating performance. Through the overview of different PWM strategies ranging from natural and normal modulation to advanced vector modulation, this chapter has highlighted the direct impact of modulation on voltage waveform quality, harmonic reduction, and system efficiency. Furthermore, the presentation of analog, digital, and hybrid implementation strategies offers system design flexibility and optimization. Such observations form a strong foundation for the modeling and simulation of modulation strategies, which will be discussed at greater length in the following sections with the use of MATLAB/Simulink tools to validate theoretical concepts through practical application.

Part two: Modeling of the dual star induction machine

2.2.1. Introduction

The Dual-Star Induction Machine (DSIM) is increasingly used in high-power industrial applications due to its robustness, reliability, and versatility. With rising demand for efficient and controllable drives, precise DSIM modeling is essential for implementing advanced control strategies and optimizing performance. This section presents a detailed mathematical model encompassing the machine's electrical, magnetic, and mechanical behavior. It includes the derivation of voltage and flux linkage equations in both natural and synchronous reference frames, applies the Park transformation for vector control, and formulates the electromagnetic torque and mechanical dynamics to achieve a complete dynamic representation

2.2.2. Simplifying assumptions

Before modeling the machine, several simplifying assumptions are made to ease analysis:

- The air gap is uniform, and notching effects are ignored.
- Magnetic saturation, hysteresis, and eddy currents are neglected.
- Winding resistances are constant (no temperature variation), and skin effect is neglected.
- Mutual leakage inductance between the two star circuits is neglected.
- The machine operates under balanced conditions. [22]

2.2.3. The dual star induction machine model

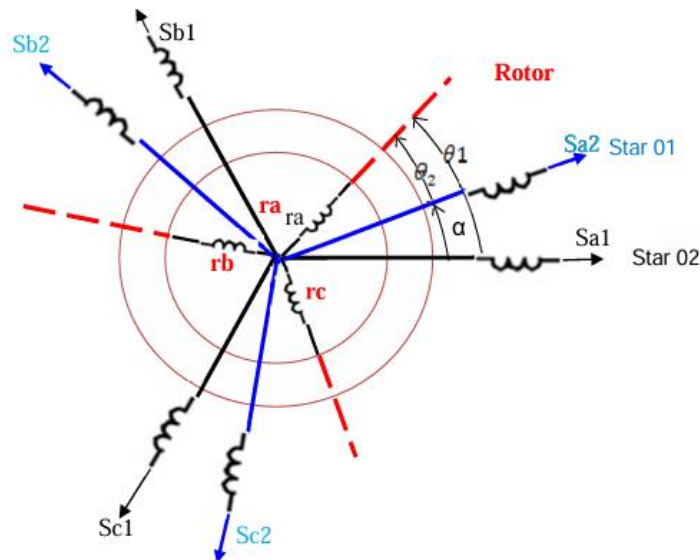


Figure 2.6: Representation of the windings of the dual-star induction machine. [23]

2.2.3.1. Electrical equations of the DSIM(voltage)

The DSIM voltage equations represent for each winding the sum of the ohmic drop and inductive drop owing to the flux. Applying the OHM rule to each of the two stator and rotor states yields the following composition: [24] [25]

▪ *Star N°1*

$$\begin{cases} [V_{sa1}] = [R_{sa1}][I_{sa1}] + \frac{d}{dt}[\phi_{sa1}] \\ [V_{sb1}] = [R_{sb1}][I_{sb1}] + \frac{d}{dt}[\phi_{sb1}] \\ [V_{sc1}] = [R_{sc1}][I_{sc1}] + \frac{d}{dt}[\phi_{sc1}] \end{cases} \quad (2.15)$$

▪ *Star N°2*

$$\begin{cases} [V_{sa2}] = [R_{sa2}][I_{sa2}] + \frac{d}{dt}[\phi_{sa2}] \\ [V_{sb2}] = [R_{sb2}][I_{sb2}] + \frac{d}{dt}[\phi_{sb2}] \\ [V_{sc2}] = [R_{sc2}][I_{sc2}] + \frac{d}{dt}[\phi_{sc2}] \end{cases} \quad (2.16)$$

▪ *Rotor*

$$\begin{cases} [R_{ra}][I_{ra}] + \frac{d}{dt}[\phi_{ra}] = 0 \\ [R_{rb}][I_{rb}] + \frac{d}{dt}[\phi_{rb}] = 0 \\ [R_{rc}][I_{rc}] + \frac{d}{dt}[\phi_{rc}] = 0 \end{cases} \quad (2.17)$$

The rotor voltages equal to zero because in a squirrel-cage induction machine, the rotor does not have external electrical connections. The rotor windings are short-circuited internally at both ends.

Assume:

$$R_{s1} = R_{sa1} = R_{sb1} = R_{sc1} ; R_{s2} = R_{sa2} = R_{sb2} = R_{sc2} \text{ et } R_r = R_{ra} = R_{rb} = R_{rc}$$

In matrix form, we have:

$$\begin{cases} [R_{s1}] = R_{s1}[ID]_{3 \times 3} \\ [R_{s2}] = R_{s2}[ID]_{3 \times 3} \\ [R_r] = R_r[ID]_{3 \times 3} \end{cases} \quad (2.18)$$

Therefore:

$$\begin{cases} [V_{s1}] = [R_{s1}][I_{s1}] + \frac{d}{dt}[\phi_{s1}] \\ [V_{s2}] = [R_{s2}][I_{s2}] + \frac{d}{dt}[\phi_{s2}] \\ 0 = [R_r][I_r] + \frac{d}{dt}[\phi_r] \end{cases} \quad (2.19)$$

Such as:

- $V_{sa1}, V_{sb1}, V_{sc1}$: Phase voltages of the stator winding 1 (phases a, b, and c).
- $V_{sa2}, V_{sb2}, V_{sc2}$: Phase voltages of the stator winding 2.
- $I_{sa1}, I_{sb1}, I_{sc1}$: Phase currents of stator winding 1.
- $I_{sa2}, I_{sb2}, I_{sc2}$: Phase currents of stator winding 2.
- I_{ra}, I_{rb}, I_{rc} : Rotor phase currents.
- $\phi_{sa1}, \phi_{sb1}, \phi_{sc1}$: Magnetic flux linkages in stator winding 1
- $\phi_{sa2}, \phi_{sb2}, \phi_{sc2}$: Magnetic flux linkages in stator winding 2.
- $\phi_{ra}, \phi_{rb}, \phi_{rc}$: Rotor flux linkages.
- $R_{sa1}, R_{sb1}, R_{sc1}$: Phase resistances of stator winding 1.
- $R_{sa2}, R_{sb2}, R_{sc2}$: Phase resistances of stator winding 2.
- R_{ra}, R_{rb}, R_{rc} : Rotor resistances.

2.2.3.2. Magnetic equation of the DSIM

From the matrix $[L(\theta)]$ we obtain the equations for flux versus currents. [26]

$$[L_{\theta}] = \begin{bmatrix} L_{s1s1} & L_{s1s2} & L_{s1r} \\ L_{s2s1} & L_{s2s2} & L_{s2r} \\ L_{rs1} & L_{rs2} & L_{rr} \end{bmatrix} \quad (2.20)$$

$$\begin{cases} [\phi_{s1}] = [L_{s1s1}][I_{s1}] + [M_{s1,s2}][I_{s2}] + [M_{s1,r}][I_r] \\ [\phi_{s2}] = [L_{s2s2}][I_{s2}] + [M_{s2,s1}][I_{s1}] + [M_{s2,r}][I_r] \\ [\phi_r] = [L_{r,r}][I_r] + [M_{r,s1}][I_{s1}] + [M_{r,s2}][I_{s2}] \end{cases} \quad (2.25)$$

The stator and rotor flux equations in matrix form given by:

$$\begin{bmatrix} \phi_{s1} \\ \phi_{s2} \\ \phi_r \end{bmatrix} = \begin{bmatrix} L_{s1s1} & L_{s1s2} & L_{s1r} \\ L_{s2s1} & L_{s2s2} & L_{s2r} \\ L_{rs1} & L_{rs2} & L_{rr} \end{bmatrix} \times \begin{bmatrix} I_{s1} \\ I_{s2} \\ I_r \end{bmatrix} \quad (2.26)$$

Such as:

- L_{s1s1} : Inductance matrix of star 1.
- L_{s2s2} : Inductance matrix of star 2.
- L_{rr} : Inductance of rotor.
- L_{s1s2} : Mutual inductance matrix between star 1 and star 2.
- L_{s1r} : Mutual inductance matrix between star 1 and rotor.
- L_{s2s1} : Mutual inductance matrix between star 2 and star 1.
- L_{s2r} : Mutual inductance matrix between star 2 and rotor.
- L_{rs1} : Mutual inductance matrix between rotor and star 1.

- L_{rs2} : Mutual inductance matrix between rotor and star 2.

The usual simplifying assumption for the sinusoidal representation of m.m.f. gives us the following expressions:

$$[L_{s2,s2}] = \begin{bmatrix} L_{s2} + L_{ms} & -\frac{L_{ms}}{2} & -\frac{L_{ms}}{2} \\ -\frac{L_{ms}}{2} & L_{s2} + L_{ms} & -\frac{L_{ms}}{2} \\ -\frac{L_{ms}}{2} & -\frac{L_{ms}}{2} & L_{s2} + L_{ms} \end{bmatrix} \quad (2.27)$$

$$[L_{r,r}] = \begin{bmatrix} L_r + L_{mr} & -\frac{L_{mr}}{2} & -\frac{L_{mr}}{2} \\ -\frac{L_{mr}}{2} & L_r + L_{mr} & -\frac{L_{mr}}{2} \\ -\frac{L_{mr}}{2} & -\frac{L_{mr}}{2} & L_r + L_{mr} \end{bmatrix} \quad (2.28)$$

$$[M_{s1,s2}] = L_{ms} \begin{bmatrix} \cos(\alpha) & \cos\left(\alpha + \frac{2\pi}{3}\right) & \cos\left(\alpha + \frac{4\pi}{3}\right) \\ \cos\left(\alpha + \frac{4\pi}{3}\right) & \cos(\alpha) & \cos\left(\alpha + \frac{2\pi}{3}\right) \\ \cos\left(\alpha + \frac{2\pi}{3}\right) & \cos\left(\alpha + \frac{4\pi}{3}\right) & \cos(\alpha) \end{bmatrix} \quad (2.29)$$

$$[M_{s1,r}] = M_{sr} \begin{bmatrix} \cos(\theta_r) & \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r + \frac{4\pi}{3}\right) \\ \cos\left(\theta_r + \frac{4\pi}{3}\right) & \cos(\theta_r) & \cos\left(\theta_r + \frac{2\pi}{3}\right) \\ \cos\left(\theta_r + \frac{2\pi}{3}\right) & \cos\left(\theta_r + \frac{4\pi}{3}\right) & \cos(\theta_r) \end{bmatrix} \quad (2.30)$$

$$[M_{s2,r}] = M_{sr} \begin{bmatrix} \cos(\theta_r - \alpha) & \cos\left(\theta_r - \alpha + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \alpha + \frac{4\pi}{3}\right) \\ \cos\left(\theta_r - \alpha + \frac{4\pi}{3}\right) & \cos(\theta_r - \alpha) & \cos\left(\theta_r - \alpha + \frac{2\pi}{3}\right) \\ \cos\left(\theta_r - \alpha + \frac{2\pi}{3}\right) & \cos\left(\theta_r - \alpha + \frac{4\pi}{3}\right) & \cos(\theta_r - \alpha) \end{bmatrix} \quad (2.31)$$

When :

- L_{s1}, L_{s2} : Cyclic self-inductance of 1st star and 2nd star
- L_r : The rotor's cyclic self-inductance
- L_{ms}, L_{mr} : The maximum value of stator and rotor mutual inductance coefficients respectively.
- M_{sr} : The maximum value of mutual inductance coefficients between a star and the rotor.
- *Magnetic energy :*

$$\omega_{mag} = \frac{1}{2} ([I_{s1}]^t [\phi_{s1}] + [I_{s2}]^t [\phi_{s2}] + [I_r]^t [\phi_r]) \quad (2.32)$$

2.2.3.3. Electromagnetic torque

Electromagnetic torque is expressed as the partial derivative of electromagnetic energy storage with respect to the geometric angle of rotation of the rotor: [27] [24]

$$T_{em} = \frac{d\omega_{mag}}{d\theta_m} = p \frac{d\omega_{mag}}{d\theta_e} \quad (2.33)$$

With:

- P: number of pole pairs.
- θ_m : Mechanical rotor angle.
- θ_e : Electrical rotor angle.

The electromagnetic torque is given by the following expression:

$$T_{em} = \frac{p}{2} \left\{ [I_{s1}] \frac{d}{d\theta_r} [L_{s1,r}] [I_r]^t + [I_{s2}] \frac{d}{d\theta_r} [L_{s2,r}] [I_r]^t \right\} \quad (2.34)$$

2.2.3.4. Mechanical equation

The solution of the fundamental equation of dynamics:

$$J \frac{d\Omega}{dt} + f\Omega = T_{em} - T_L \quad ; \quad \Omega = \frac{\omega}{p} \quad (2.35)$$

With :

- T_{em} : Electromagnetic torque produced by the motor.
- T_L : Load (resistant) torque.
- ω : Electrical angular speed.
- Ω : Mechanical angular speed.
- J: Moment of inertia.

2.2.4. Park transformation [28]

This transformation consists in converting the three-phase coils (stator and rotor) into two orthogonal two-phase coils, UV. The conversion involves transforming them into electrically and magnetically equivalent coils. The figure below illustrates the transformation of the actual (abc) coils into orthogonal UV coils. In equation form:

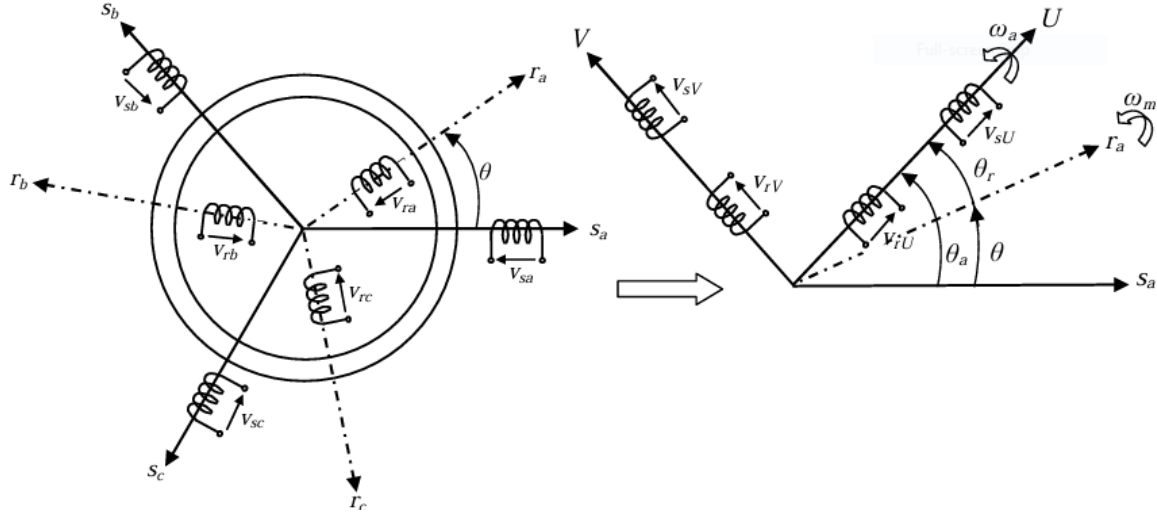


Figure 2.7: Switching from three-phase abc to two-phase UV system .

The notations used in the figure represent the following:

- θ : Electrical angle between stator and rotor axes.
- θ_a : Rotor angle in vectorial (rotating) reference frame.
- θ_r : Rotor electrical position
- ω_a : The electrical angular of the UV axis system($\omega_a = \frac{d\theta_a}{dt}$);
- ω_m : Electrical angular of the rotor($\omega_m = \frac{d\theta}{dt}$).

This transformation is defined by its transition matrix $[P(\theta_a)]$ such that:

$$[P(\theta)] = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos \theta_a & \cos(\theta_a - \frac{2\pi}{3}) & \cos(\theta_a + \frac{2\pi}{3}) \\ -\sin \theta_a & -\sin(\theta_a - \frac{2\pi}{3}) & -\sin(\theta_a + \frac{2\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (2.36)$$

The transition matrix $[P(\theta_a)]$ written in this form is orthogonal, which leads to the conservation of instantaneous power. Orthogonality allows us to write:

$$[P(\theta_a)]^{-1} = [P(\theta_a)]^T = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos \theta_a & -\sin \theta_a & \frac{1}{\sqrt{2}} \\ \cos(\theta_a - \frac{2\pi}{3}) & -\sin(\theta_a - \frac{2\pi}{3}) & \frac{1}{\sqrt{2}} \\ \cos(\theta_a + \frac{2\pi}{3}) & -\sin(\theta_a + \frac{2\pi}{3}) & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (2.37)$$

The angle θ_a takes the value θ_s for stator quantities and the value ($\theta_r = \theta_s - \theta$) for rotor quantities.

This gives: $[X_{UV}] = [P(\theta_a)][X_{abc}]$ (2.38)

- X : can represent voltage, current or flux.

When the sums of the real components (abc) are zero, the equation translating the component which is always verified as identically zero, becomes useless.

The inverse transformation is written:

$$[X_{abc}] = [P(\theta_a)]^{-1}[X_{UV}] \quad (2.39)$$

2.2.5. Applying Park's transformation to the DSIM [29]

2.2.5.1. Voltage equations

- *Star N°01*

$$\begin{bmatrix} V_{ds1} \\ V_{qs1} \\ V_{os1} \end{bmatrix} = \begin{bmatrix} R_{s1} & 0 & 0 \\ 0 & R_{s1} & 0 \\ 0 & 0 & R_{s1} \end{bmatrix} \begin{bmatrix} I_{ds1} \\ I_{qs1} \\ I_{os1} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \phi_{ds1} \\ \phi_{qs1} \\ \phi_{os1} \end{bmatrix} + \frac{d\theta_{coor}}{dt} \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \phi_{ds1} \\ \phi_{qs1} \\ \phi_{os1} \end{bmatrix} \quad (2.40)$$

- *Star N°02*

$$\begin{bmatrix} V_{ds2} \\ V_{qs2} \\ V_{os2} \end{bmatrix} = \begin{bmatrix} R_{s2} & 0 & 0 \\ 0 & R_{s2} & 0 \\ 0 & 0 & R_{s2} \end{bmatrix} \begin{bmatrix} I_{ds2} \\ I_{qs2} \\ I_{os2} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \phi_{ds2} \\ \phi_{qs2} \\ \phi_{os2} \end{bmatrix} + \frac{d(\theta_{coor} - \alpha)}{dt} \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \phi_{ds2} \\ \phi_{qs2} \\ \phi_{os2} \end{bmatrix} \quad (2.41)$$

- *Rotor*: The rotor is short-circuited, i.e. $V_{ar} = V_{br} = V_{cr} = 0$.

$$\begin{bmatrix} V_{dr} \\ V_{qr} \\ V_{or} \end{bmatrix} = \begin{bmatrix} R_r & 0 & 0 \\ 0 & R_r & 0 \\ 0 & 0 & R_r \end{bmatrix} \begin{bmatrix} I_{dr} \\ I_{qr} \\ I_{or} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \phi_{dr} \\ \phi_{qr} \\ \phi_{or} \end{bmatrix} + \frac{d\theta_{rcoor}}{dt} \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \phi_{dr} \\ \phi_{qr} \\ \phi_{or} \end{bmatrix} \quad (2.42)$$

- θ_{coor} : the angle formed by the axes A_{s1} - d.
- $\theta_{rcoor} = \theta_{coor} - \theta_m$: the angle formed by the axes A_r , d
- $\frac{d\theta_{coor}}{dt} = \omega_{coor}$: the speed of rotation of (d, q) relative to star 1.
- $\frac{d\theta_{rcoor}}{dt} = \omega_{rcoor}$: The speed of rotation of frame (d, q) relative to the rotor.

$$\begin{cases} V_{ds1} = R_{s1}I_{ds1} + \frac{d\phi_{ds1}}{dt} - \omega_{coor}\phi_{qs1} \\ V_{qs1} = R_{s1}I_{qs1} + \frac{d\phi_{qs1}}{dt} + \omega_{coor}\phi_{ds1} \\ V_{ds2} = R_{s2}I_{ds2} + \frac{d\phi_{ds2}}{dt} - \omega_{coor}\phi_{qs2} \\ V_{qs2} = R_{s2}I_{qs2} + \frac{d\phi_{qs2}}{dt} + \omega_{coor}\phi_{ds2} \\ 0 = R_r I_{dr} + \frac{d\phi_{dr}}{dt} - (\omega_{coor} - \omega_r)\phi_{qr} \\ 0 = R_r I_{qr} + \frac{d\phi_{qr}}{dt} + (\omega_{coor} - \omega_r)\phi_{dr} \end{cases} \quad (2.43)$$

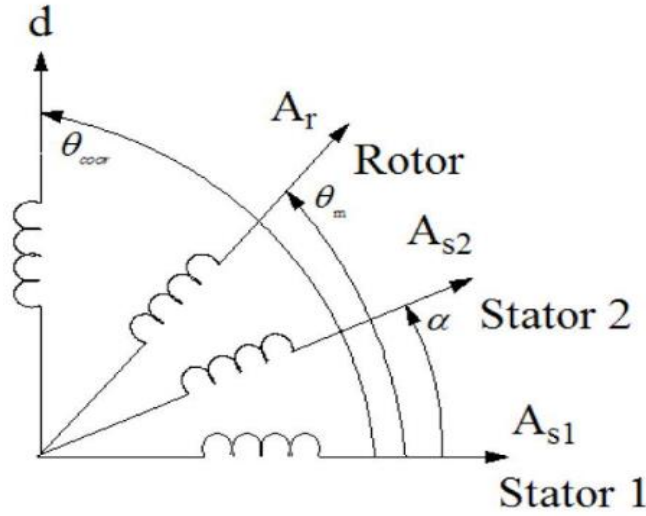


Figure 2.8: Representation of axes in Park transformations.

2.2.5.2. Flux equations [28] [26]

As with the application of Park's transformation to the voltage equations, we apply this transformation to the flux equations, obtaining:

$$\begin{cases} \phi_{ds1} = L_{s1}I_{ds1} + \frac{3}{2}L_{ms}I_{ds1} + \frac{3}{2}L_{ms}I_{ds2} + \frac{3}{2}M_{sr}I_{dr} \\ \phi_{qs1} = L_{s1}I_{qs1} + \frac{3}{2}L_{ms}I_{qs1} + \frac{3}{2}L_{ms}I_{qs2} + \frac{3}{2}M_{sr}I_{qr} \\ \phi_{ds2} = L_{s2}I_{ds2} + \frac{3}{2}L_{ms}I_{ds2} + \frac{3}{2}L_{ms}I_{ds1} + \frac{3}{2}M_{sr}I_{dr} \\ \phi_{qs2} = L_{s2}I_{qs2} + \frac{3}{2}L_{ms}I_{qs2} + \frac{3}{2}L_{ms}I_{qs1} + \frac{3}{2}M_{sr}I_{qr} \\ \phi_{dr} = L_r I_{dr} + \frac{3}{2}L_{mr}I_{dr} + \frac{3}{2}M_{sr}I_{ds1} + \frac{3}{2}M_{sr}I_{ds2} \\ \phi_{qr} = L_r I_{qr} + \frac{3}{2}L_{mr}I_{qr} + \frac{3}{2}M_{sr}I_{qs1} + \frac{3}{2}M_{sr}I_{qs2} \end{cases} \quad (2.44)$$

$$\frac{3}{2}L_{ms} = \frac{3}{2}L_{mr} = \frac{3}{2}M_{sr} = L_m \quad (2.45)$$

L_m : the cyclic mutual inductance between star 1, star 2 and the rotor.

The system of the flux equations becomes as follows:

$$\begin{cases} \phi_{ds1} = L_{s1}I_{ds1} + L_m(I_{ds1} + I_{ds2} + I_{dr}) \\ \phi_{qs1} = L_{s1}I_{qs1} + L_m(I_{qs1} + I_{qs2} + I_{dr}) \\ \phi_{ds2} = L_{s2}I_{ds2} + L_m(I_{ds1} + I_{ds2} + I_{dr}) \\ \phi_{qs2} = L_{s2}I_{qs2} + L_m(I_{qs1} + I_{qs2} + I_{dr}) \\ \phi_{dr} = L_r I_{dr} + L_m(I_{ds1} + I_{ds2} + I_{dr}) \\ \phi_{qr} = L_r I_{qr} + L_m(I_{qs1} + I_{qs2} + I_{dr}) \end{cases} \quad (2.46)$$

With:

- $L_{s1} + L_m$: cyclic self-inductance of star 1;
- $L_{s2} + L_m$: the cyclic self-inductance of star 2;
- $L_r + L_m$: the rotor's cyclic self-inductance.

2.2.5.3. Mechanical equation

When the reference frame is changed, the expression for the electromagnetic torque must be found in the new reference frame.

To calculate the expression of the instantaneous torque, it is necessary to determine the instantaneous power. The instantaneous power absorbed by the double star asynchronous machine is given by the following expression: [30]

$$P_{abc} = [V_{s1}]^T [I_{s1}] + [V_{s2}]^T [I_{s2}] \quad (2.47)$$

Resulting in:

$$P_{abc} = V_{as1}I_{as1} + V_{bs1}I_{bs1} + V_{cs1}I_{cs1} + V_{as2}I_{as2} + V_{bs2}I_{bs2} + V_{cs2}I_{cs2} \quad (2.48)$$

As we indicated earlier, Park's transformation allows power to be conserved we can write:

$$P_{abc} = V_{ds1}I_{ds1} + V_{qs1}I_{qs1} + V_{ds2}I_{ds2} + V_{qs2}I_{qs2} \quad (2.49)$$

If we replace the voltages and currents of axes (d, q) in the system of this equation by their expressions in equation of voltage, we find the expression for instantaneous power consumption:

$$\begin{aligned} P_{abc} = & R_{s1}I_{ds1}^2 + R_{s1}I_{qs1}^2 + R_{s2}I_{ds2}^2 + \omega_{coor}(\phi_{ds1}I_{qs1} - \phi_{qs1}I_{ds1} + \phi_{ds2}I_{qs2} - \phi_{qs2}I_{ds2}) \\ & + \frac{d\phi_{ds1}}{dt} I_{ds1} + \frac{d\phi_{qs1}}{dt} I_{qs1} + \frac{d\phi_{ds2}}{dt} I_{ds2} + \frac{d\phi_{qs2}}{dt} I_{qs2} \end{aligned} \quad (2.50)$$

Electromagnetic power and torque can be written in the universal form:

$$P_{em} = \Omega T_{em} \quad (2.51)$$

With,

- Ω : the mechanical rotational speed of the rotor.

In the equation for acquired power (2.50), the second term represents electromagnetic power:

$$P_{em} = +\omega_{coor}(\phi_{ds1}I_{qs1} - \phi_{qs1}I_{ds1} + \phi_{ds2}I_{qs2} - \phi_{qs2}I_{ds2}) \quad (2.52)$$

By replacing expression (2.52) in (2.51) it is clear that the electromagnetic torque is of the following form:

$$T_{em} = P(\phi_{ds1}I_{qs1} - \phi_{qs1}I_{ds1} + \phi_{ds2}I_{qs2} - \phi_{qs2}I_{ds2}) \quad (2.53)$$

2.2.6. Choice of reference system

Machine variables are in the (dq) reference frame at electrical angle θ_r to the rotor. An additional angle, θ_{sa} , for the stator is introduced but unspecified. The reference frame can be fixed to the stator, rotor, or spinning magnetic field, often defined by the stator flux vector. Although called a "vector," it represents a full magnetic field, with the vector showing only phase and amplitude. The rotor flux can also be depicted by a rotor flux vector that spins with the magnetic field, but the rotor has a relative "slip" motion. Under steady-state, rotor and stator flux vectors rotate at the same speed, showing they are synchronous. [31]

2.2.6.1. Stator reference frame (α, β)

In this frame of reference, the axes (d, q) are stationary with respect to the stator $\omega_{coor} = 0$, in this situation, phases As1 and d coincide. This frame of reference is most suited to working with instantaneous quantities, as it does not require any transition to the real system. The usage of this method allows us to investigate the starting and braking cycles of AC machines.

2.2.6.2. Rotor reference frame

In this frame of reference, the frame's electrical speed (d, q) equals the rotor's mechanical pulsation $\omega_{coor} = \omega_r$. This frame of reference is fixed relative to the rotor and is used to investigate transient regimes in both induction and synchronous machines.

2.2.6.3. Reference frame linked to the rotating field

In this reference frame, the reference frame (d, q) has the same speed as the rotating fields $\omega_{coord} = \omega_s$, this frame of reference is commonly used to apply speed, torque, and other instructions because the quantities are continuous.

2.2.7. State-Space Representation of the Dual-Star Induction Machine

In this work, we use the reference frame linked to the rotating field to model and control the DSIM. In this case, the DSIM model becomes: [32]

$$\begin{cases} V_{ds1} = R_{s1}I_{ds1} + \frac{d\phi_{ds1}}{dt} - \omega_s\phi_{qs1} \\ V_{qs1} = R_{s1}I_{qs1} + \frac{d\phi_{qs1}}{dt} + \omega_s\phi_{ds1} \\ V_{ds2} = R_{s2}I_{ds2} + \frac{d\phi_{ds2}}{dt} - \omega_s\phi_{qs2} \\ V_{qs2} = R_{s2}I_{qs2} + \frac{d\phi_{qs2}}{dt} + \omega_s\phi_{ds2} \\ 0 = R_r I_{dr} + \frac{d\phi_{dr}}{dt} - (\omega_s - \omega_r)\phi_{qr} \\ 0 = R_r I_{qr} + \frac{d\phi_{qr}}{dt} + (\omega_s - \omega_r)\phi_{dr} \end{cases} \quad (2.54)$$

The two expressions for magnetizing fluxes as a function of stator and rotor currents are:

$$\begin{cases} \phi_{md} = (I_{ds1} + I_{ds2} + I_{dr})L_m \\ \phi_{mq} = (I_{qs1} + I_{qs2} + I_{qr})L_m \end{cases} \quad (2.55)$$

By introducing the magnetizing flux expressions into the system of flux equations (2.46), we obtain:

$$\begin{cases} I_{ds1} = \frac{\phi_{ds1} - \phi_{md}}{L_{s1}} \\ I_{ds2} = \frac{\phi_{ds2} - \phi_{md}}{L_{s2}} \\ I_{qs1} = \frac{\phi_{qs1} - \phi_{mq}}{L_{s1}} \\ I_{qs2} = \frac{\phi_{qs2} - \phi_{mq}}{L_{s2}} \\ I_{dr} = \frac{\phi_{dr} - \phi_{md}}{L_r} \\ I_{qr} = \frac{\phi_{qr} - \phi_{mq}}{L_r} \end{cases} \quad (2.56)$$

Replacing the currents in the system of equations (2.56) with their expression in the system of equations (2.43) gives: [30] [32]

$$\begin{cases} \frac{d\phi_{ds1}}{dt} = V_{ds1} - \frac{R_{s1}}{L_{s1}}(\phi_{ds1} - \phi_{md}) + \omega_s \phi_{qs1} \\ \frac{d\phi_{ds2}}{dt} = V_{ds2} - \frac{R_{s2}}{L_{s2}}(\phi_{ds2} - \phi_{md}) + \omega_s \phi_{qs2} \\ \frac{d\phi_{qs1}}{dt} = V_{qs1} - \frac{R_{s1}}{L_{s1}}(\phi_{qs1} - \phi_{md}) + \omega_s \phi_{ds1} \\ \frac{d\phi_{qs2}}{dt} = V_{qs2} - \frac{R_{s2}}{L_{s2}}(\phi_{qs2} - \phi_{md}) + \omega_s \phi_{ds2} \\ \frac{d\phi_{dr}}{dt} = -\frac{R_r}{L_r}(\phi_{dr} - \phi_{md}) + \omega_{gl} \phi_{qr} \\ \frac{d\phi_{qr}}{dt} = -\frac{R_r}{L_r}(\phi_{qr} - \phi_{md}) + \omega_{gl} \phi_{dr} \end{cases} \quad (2.57)$$

Using the equation (2.46) we have: $\omega_{gl} = \omega_s - \omega_r$ (2.58)

- ω_{gl} : Slip angular frequency.
- ω_s : Synchronous electrical angular speed of the stator field.
- ω_r : Rotor electrical angular speed.

$$\begin{cases} \phi_{md} = L_a \left(\frac{\phi_{ds1}}{L_{s1}} + \frac{\phi_{ds2}}{L_{s2}} + \frac{\phi_{dr}}{L_r} \right) \\ \phi_{mq} = L_a \left(\frac{\phi_{qs1}}{L_{s1}} + \frac{\phi_{qs2}}{L_{s2}} + \frac{\phi_{qr}}{L_r} \right) \end{cases} \quad (2.59)$$

And:

$$L_a = \frac{1}{\frac{1}{L_{s1}} + \frac{1}{L_{s1}} + \frac{1}{L_{s1}} + \frac{1}{L_{s1}}} \quad (2.60)$$

It is possible to obtain other expressions for the instantaneous torque by using the stator flux expressions and replacing (2.46) in (2.51), we obtain:

$$T_{em} = PL_m [(I_{qs1} + I_{qs2})I_{dr} - (I_{ds1} + I_{ds2})I_{qr}] \quad (2.61)$$

By using the equation (2.46) we have the rotor currents:

$$\begin{cases} I_{dr} = \frac{1}{L_r + L_m} [\phi_{dr} - L_m(I_{ds1} + I_{ds2})] \\ I_{qr} = \frac{1}{L_r + L_m} [\phi_{qr} - L_m(I_{qs1} + I_{qs2})] \end{cases} \quad (2.62)$$

We replace the equation (2.62) in the equation (2.61) so the instantaneous torque is:

$$T_{em} = P \frac{L_m}{L_r + L_m} [(I_{qs1} + I_{qs2})\phi_{dr} - (I_{ds1} + I_{ds2})\phi_{qr}] \quad (2.63)$$

Finally, the machine's mechanical equation can be written as follows:

$$J \frac{d\Omega}{dt} = T_{em} - T_L - K_f \Omega \quad (2.64)$$

From the replacement of the magnetizing flux expressions (ϕ_{md} , ϕ_{mq}) in (2.59) and after simplification, we find the new system of equations:

$$\begin{cases} \frac{d\phi_{ds1}}{dt} = V_{ds1} - \left(\frac{R_{s1}}{L_{s1}} - \frac{R_{s1}L_a}{L_{s1}^2} \right) \phi_{ds1} + \frac{R_{s1}L_a}{L_{s1}L_{s2}} \phi_{ds2} + \omega_s \phi_{qs1} + \frac{R_{s1}L_a}{L_r L_{s1}} \phi_{dr} \\ \frac{d\phi_{ds2}}{dt} = V_{ds2} - \left(\frac{R_{s2}}{L_{s2}} - \frac{R_{s2}L_a}{L_{s2}^2} \right) \phi_{ds2} + \frac{R_{s2}L_a}{L_{s1}L_{s2}} \phi_{ds1} + \omega_s \phi_{qs2} + \frac{R_{s2}L_a}{L_r L_{s2}} \phi_{dr} \\ \frac{d\phi_{qs1}}{dt} = V_{qs1} - \left(\frac{R_{s1}}{L_{s1}} - \frac{R_{s1}L_a}{L_{s1}^2} \right) \phi_{qs1} + \frac{R_{s1}L_a}{L_{s1}L_{s2}} \phi_{qs2} - \omega_s \phi_{ds1} + \frac{R_{s1}L_a}{L_r L_{s1}} \phi_{qr} \\ \frac{d\phi_{qs2}}{dt} = V_{qs2} - \left(\frac{R_{s2}}{L_{s2}} - \frac{R_{s2}L_a}{L_{s2}^2} \right) \phi_{qs2} + \frac{R_{s2}L_a}{L_{s1}L_{s2}} \phi_{qs1} - \omega_s \phi_{ds2} + \frac{R_{s2}L_a}{L_r L_{s2}} \phi_{qr} \\ \frac{d\phi_{dr}}{dt} = - \left(\frac{R_r}{L_r} - \frac{R_r L_a}{L_r^2} \right) \phi_{dr} + \frac{R_r L_a}{L_r L_{s1}} \phi_{ds1} + (\omega_s - \omega_r) \phi_{qr} + \frac{R_r L_a}{L_r L_{s2}} \phi_{ds2} \\ \frac{d\phi_{qr}}{dt} = - \left(\frac{R_r}{L_r} - \frac{R_r L_a}{L_r^2} \right) \phi_{qr} + \frac{R_r L_a}{L_r L_{s1}} \phi_{qs1} - (\omega_s - \omega_r) \phi_{dr} + \frac{R_r L_a}{L_r L_{s2}} \phi_{qs2} \end{cases} \quad (2.65)$$

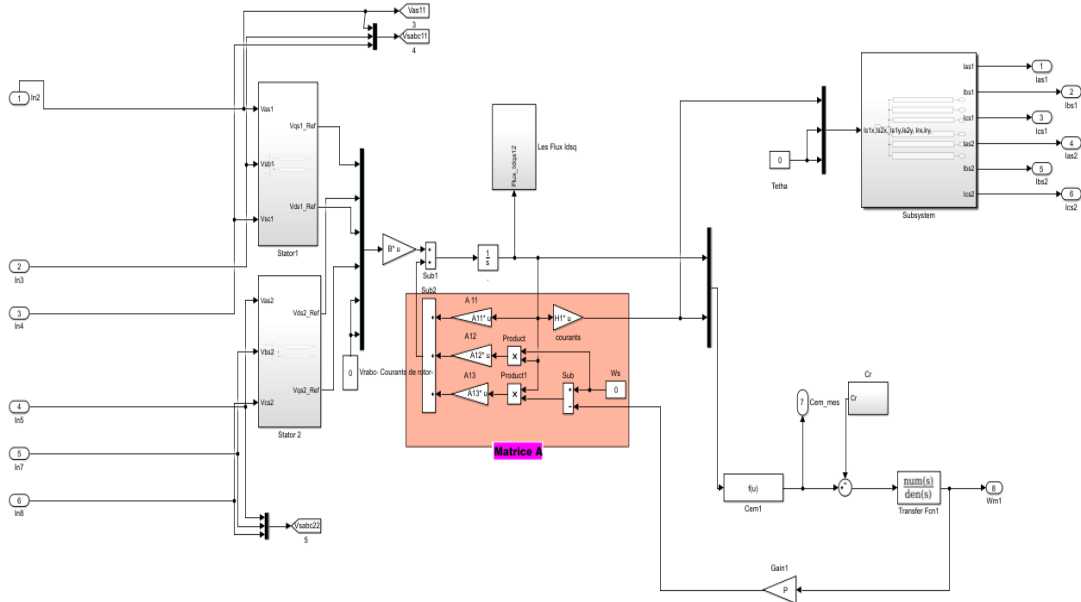


Figure 2.9: Functional block diagram of the dual-star induction machine

2.2.8. Numerical simulation and analysis of results

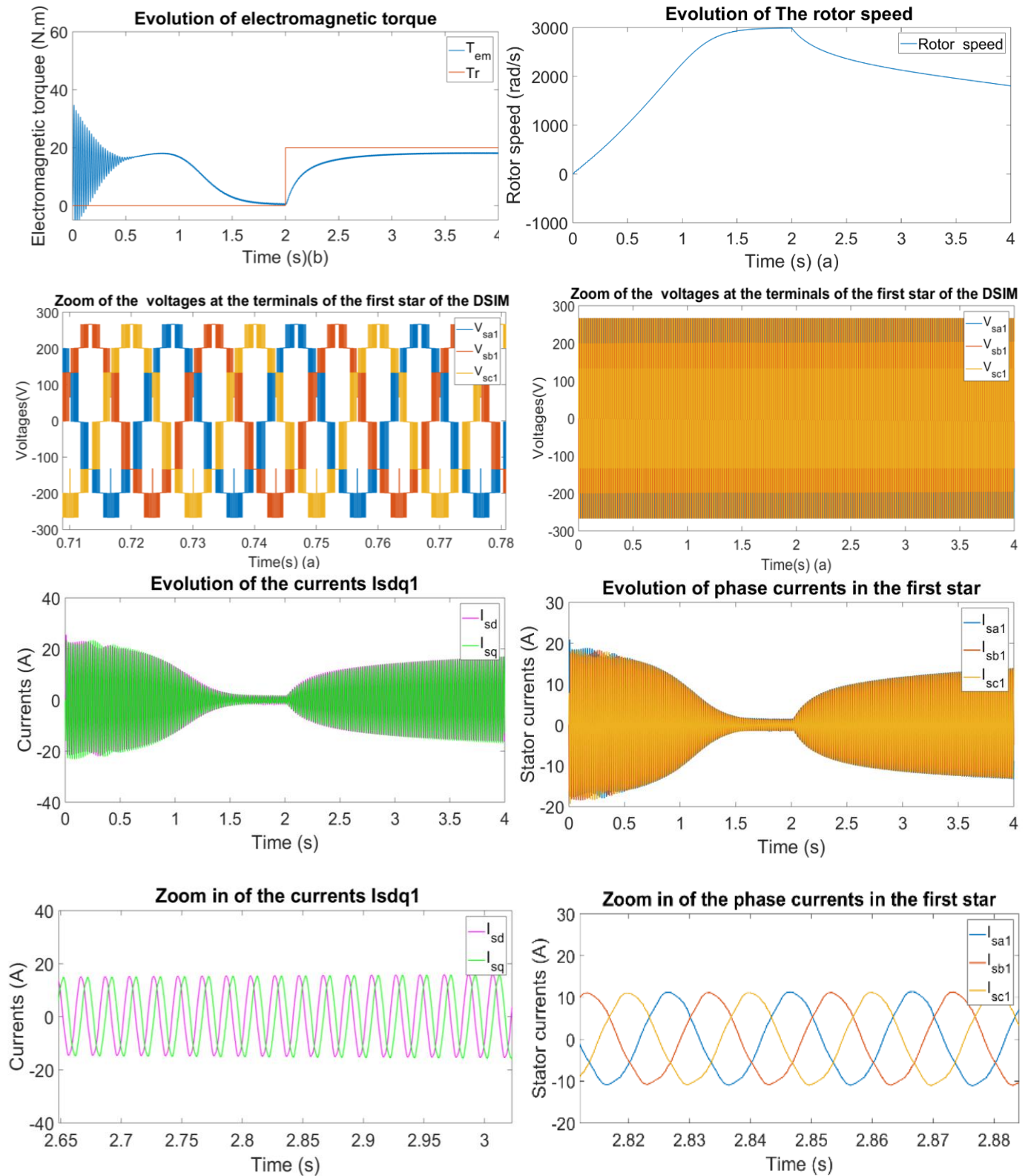


Figure 2.10: DSIM Behavior under Simulation: Torque, Speed, Voltages, and Currents Evolution

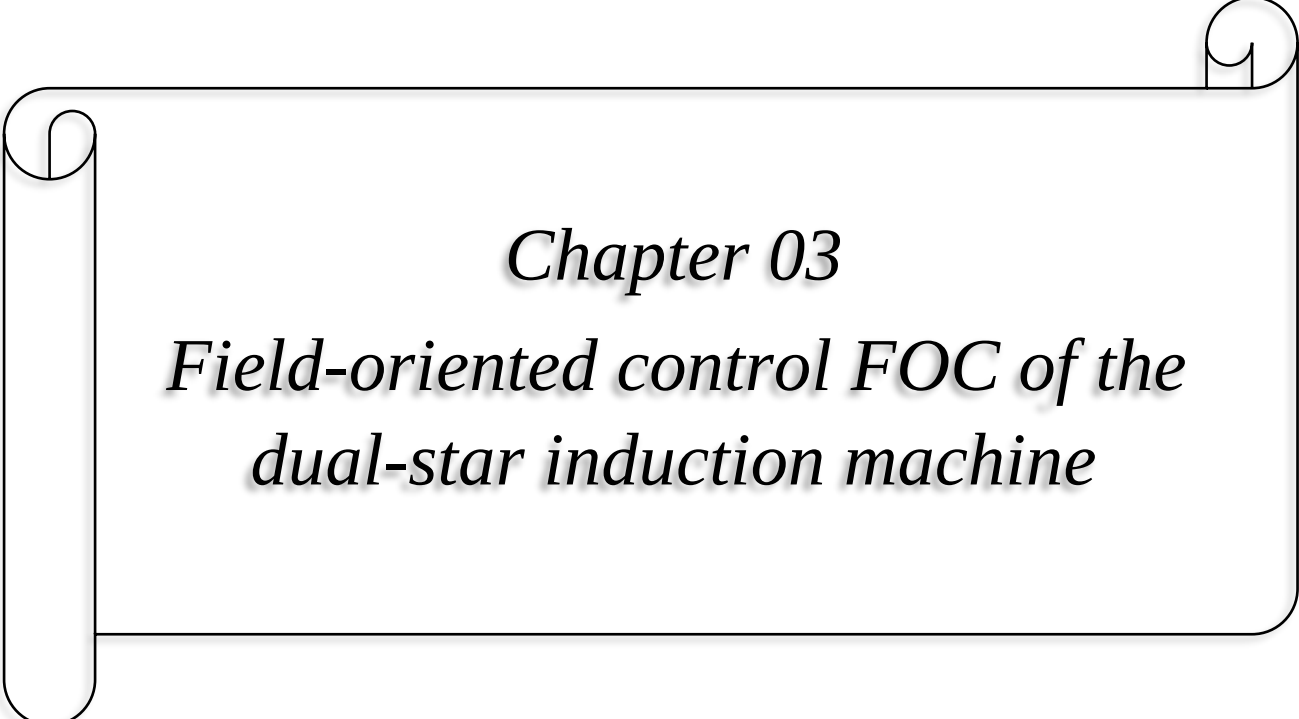
The analysis of a Dual Star Induction Machine (DSIM) powered by a three-phase three-level inverter with no control strategy reveals both the system's inherent features and its natural dynamic constraints. The stator voltages V_{sabc1} show well-defined, symmetrical steps typical of a Neutral Point Clamped (NPC) inverter, demonstrating effective harmonic reduction via multilayer PWM (MLI). The stator currents I_{sabc1} are almost sinusoidal, with acceptable spectral quality, but their stability is totally dependent on the power topology and lacks dynamic regulation. The uncontrolled interaction of flux and torque in the dq frame (I_{sdq1}) limits energy efficiency.

In the dq framework (I_{sdq1}), the absence of decoupling of the quadrature and direct components restricts the energy efficiency because of uncontrolled flux-torque interactions. The electromagnetic torque (T_{em}) experiences start-up oscillations and differs extensively from the load torque (T_r), demonstrating a lack of slip compensation. The mechanical speed ω_m achieves steady-state with a sluggish transient response in spite of reaching it. Briefly, though the three-level inverter provides a sufficiently good quality supply of voltage and current, the absence of torque control (T_{em}) and dynamic response (ω_m) calls for introducing vector or robust control methods to improve both performance and stability.

2.2.9. Conclusion

In this part, the operational principles of the Dual Star Induction Machine (DSIM) were thoroughly investigated. The entire modeling procedure was described step by step, including the development of electrical, magnetic, and mechanical equations, as well as the application of appropriate reference frame transformations.

Finally, the model was implemented in MATLAB/Simulink, and the simulation results were examined to confirm the machine's dynamic behavior.

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Chapter 03

Field-oriented control FOC of the dual-star induction machine

3.1. Introduction

Scalar control techniques, which regulate the modulus of the flux vector in the machine, have been widely adopted in industrial applications. However, these methods are known for their limited performance during transient conditions.

Since the asynchronous machine is inherently a multivariable system, controlling its speed or position and thus its torque requires the simultaneous regulation of several variables to artificially decouple flux and torque. Among the various strategies developed to achieve this decoupling, one of the most notable is vector control, introduced by Hasse in 1969 and Blaschke in 1972, by managing both the amplitude and phase of electrical quantities, vector control enables the AC machine to emulate the behavior of a separately excited DC machine, where torque and flux can be controlled independently.

3.2 Field oriented control principle

The goal of this control is to compare the behavior of the double-star asynchronous machine to that of a DC machine with independent excitation in two ways.

- Torque and machine flux can be regulated independently.
- Ensures optimal torque generation during steady-state and transient circumstances.

In a DC machine, the armature current (I_a) controls torque, while the field current (I_f) controls flux.[33]

The electromagnetic torque is given by:

$$T_{em} = k \cdot \phi_f \cdot I_a = k' \cdot I_f I_a \quad (3.1)$$

And:

- ϕ_f : Magnetic flux generated by the excitation current I_f
- I_a : The current flowing through the armature winding
- k', k : Fixed parameters

For the dual-star induction machine, vector control involves decoupling the electromagnetic torque and rotor flux components.

To accomplish this, an appropriate control law and coordinate system must be selected to ensure independent control of flux and torque.

Given that the electromagnetic torque expression depends on the stator currents and the rotor flux, if the rotor flux is aligned with the d -axis of the rotating reference frame, then:

$$\phi_{dr} = \phi_r ; \phi_{qr} = 0$$

Hence the final relationship of the couple becomes:

$$T_{em} = p \frac{L_m}{L_m + L_r} \phi_{dr} (I_{qs1} + I_{qs2}) = k \phi_{dr} (I_{qs1} + I_{qs2}) \quad (3.2)$$

$$\text{and } k = p \frac{L_m}{L_m + L_r}$$

From Equation (3.2), it is evident that the electromagnetic torque results from the interaction between a flux component and a current component. This expression is reminiscent of the torque equation for a separately excited DC machine.

Consequently, the operation of the dual-star induction machine under vector control resembles that of a separately excited DC machine, where flux and torque can be independently regulated. [35]
[34]

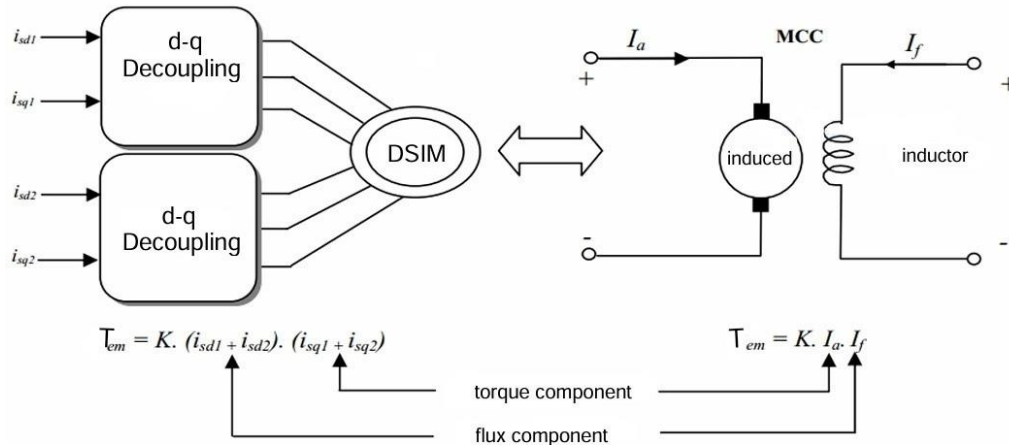


Figure 3.1: Field Oriented Control principle. [34]

3.3. Choice of flux orientation

The DSIM model operates with a voltage supply, and the reference frame is chosen to be aligned with the rotating field, denoted as (d, q). With these choices regarding the supply and the reference frame established, the next step in the control strategy is to define the direction of the flux. Three possible options are available for this selection:

- Rotor flux orientation under the conditions $\phi_{dr} = \phi_r ; \phi_{qr} = 0$
- Stator flux orientation under the conditions $\phi_{ds} = \phi_s ; \phi_{qs} = 0$

- Air gap flux orientation under the conditions $\phi_{dm} = \phi_m$; $\phi_{qm} = 0$

For the DSIM, rotor flux orientation is chosen because it enables variable speed control with independent regulation of flux and electromagnetic torque through the stator currents. [36]

3.4. Field Oriented Control methods

3.4.1. Direct method (DFOC Direct Field Oriented Control)

To determine the rotor flux, two methods are commonly used to identify both the position (θ_s) and the flux magnitude (ϕ_r):

- Sensor-based measurement: This method involves using sensors to directly measure the flux in the machine's air gap. However, flux sensors are mechanically fragile and may not perform reliably under harsh conditions such as strong vibrations or high temperatures.
- Mathematical estimation: This approach uses mathematical models to estimate the flux. While it avoids the need for physical sensors, it is sensitive to variations in machine parameters, which can affect accuracy. [37]

3.4.2. Indirect method (IFOC Indirect Field Oriented Control)

The indirect method focuses on utilizing the position of the rotor flux rather than its magnitude. The term "indirect method" refers to the approach where a flux sensor or estimator is not required, as long as a rotor position sensor is available. However, this method still relies on the presence of a rotor position sensor for accurate operation.

3.5. Indirect field-oriented control (IFOC) with closed-loop speed control

The basis of this method is to directly determine the rotor flux component from the rotor's mechanical speed of rotation using a speed sensor. This can be done for a flux-weakening block described by the following nonlinear function.

The equation describes how ϕ_{rref} is determined based on the rotor speed Ω . It is a nonlinear control law for flux weakening:

$$\begin{cases} \phi_{rref} = \phi_n & \text{if } |\Omega| \leq \Omega_n \\ \phi_{rref} = \phi_n \frac{\Omega_n}{|\Omega|} & \text{if } |\Omega| > \Omega_n \end{cases} \quad (3.3)$$

- ϕ_{rref} : The rotor flux reference.
- ϕ_n : Nominal (rated) rotor flux (maximum flux allowed).
- Ω_n : Nominal (rated) speed.

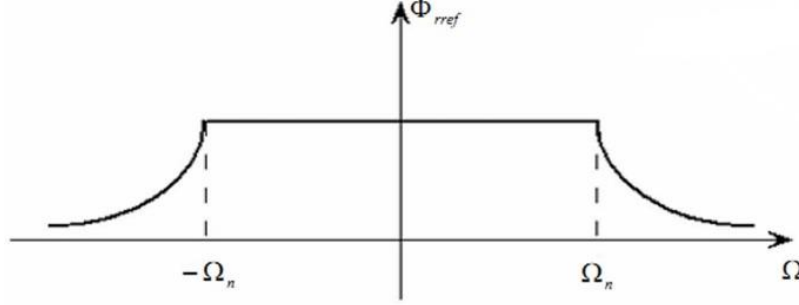


Figure 3.2: flux-weakening diagram

The flux is typically maintained at its nominal value for rotor speeds less than or equal to the machine's nominal speed Ω_n . For higher speeds, the flux is gradually reduced as the rotor speed increases, in order to limit the voltage at the machine terminals. To ensure perfect decoupling, stator current control loops are implemented ($I_{ds1}, I_{ds2}, I_{qs1}, I_{qs2}$) and ($V_{ds1}, V_{ds2}, V_{qs1}, V_{qs2}$) .[38][39]

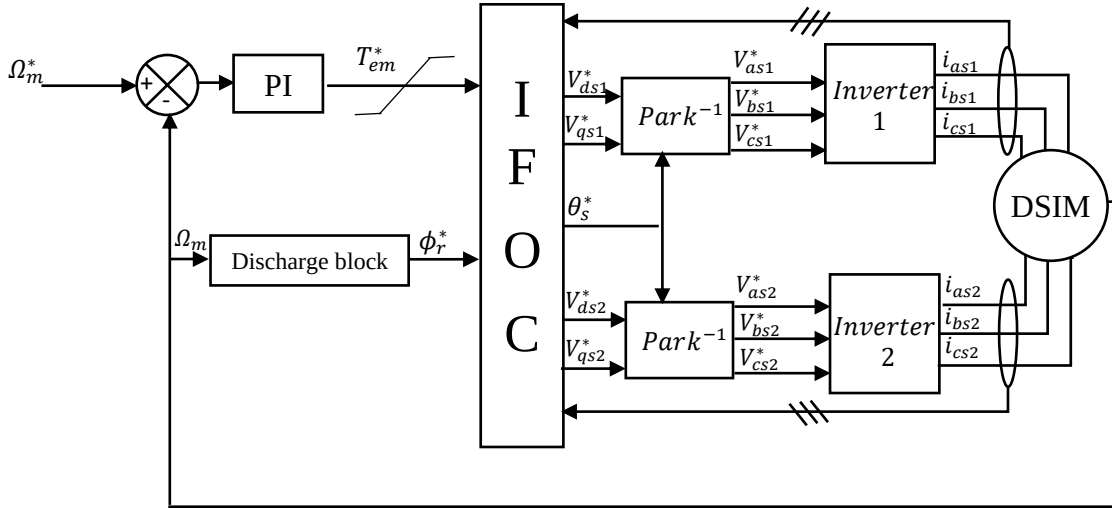


Figure 3.3: Indirect field-oriented control (IFOC) [39]

3.5.1. Identifying speed controller parameters

The diagram of the speed control loop is given in figure 3.4:

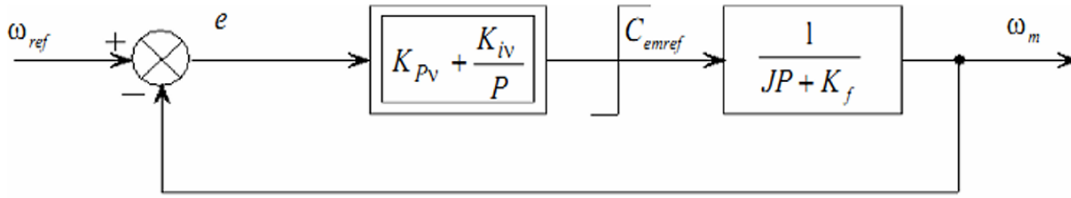


Figure 3.4: Speed regulation loop

Identification gives us:

$$\begin{cases} k_{pv} = \frac{J}{C} \\ k_{iv} = \frac{k_f}{C} \end{cases} \quad \text{we use} \quad C = C_r \quad (3.4)$$

- k_{pv} : Proportional gain.
- k_{iv} : Integral gain.
- C : Desired closed-loop time constant.
- C_r : time constant selected by the designer to ensure the desired dynamic behavior.

The control must be limited by a saturation device defined by:

$$T_{emref}(lim) = \begin{cases} T_{em} & \text{if } |T_{em}| \leq T_{emmax} \\ T_{emmax} \text{sign}(T_{emref}) & \text{if } |T_{emref}| > T_{emmax} \end{cases} \quad (3.5)$$

- T_{emref} : Reference electromagnetic torque.
- T_{emmax} : Maximum allowed electromagnetic torque. [39] [40] [41]

3.5.2. Identification of PI controller parameters [42] [43]

The representative diagram of the control loop of a first-order servo system with unitary feedback controlled by a PI is given in Figure 3.5:

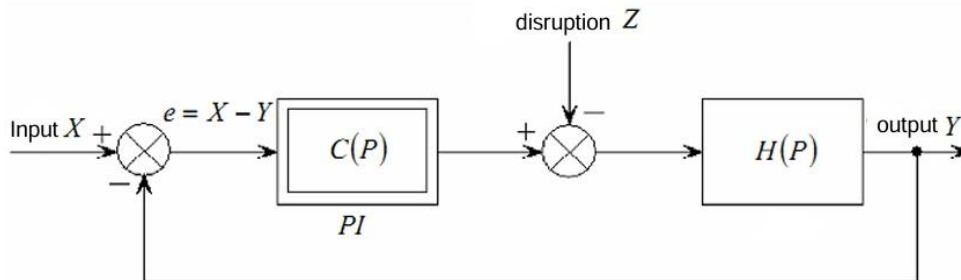


Figure 3.5: Schematic representation of a first-order system with closed-loop control via a PI regulator

The identification of PI controller parameters for systems with a first-order transfer function is of the first order, which means that:

$$H(P) = \frac{1}{aP + b} \quad (3.6)$$

The PI controller transfer function is:

$$C(P) = k_p + \frac{k_i}{P} \quad (3.7)$$

Disturbance is neglected in the controller parameter identification process.

The controlled system has the following open-loop transfer function:

$$T(P) = C(P)H(P) = \frac{k_p \cdot P + k_i}{aP^2 + bP} \quad (3.8)$$

In closed loop :

$$F(P) = \frac{T(P)}{1+T(P)} = \frac{k_p \cdot P + k_i}{aP^2 + (b+k_p)P + k_i} \quad (3.9)$$

To determine the conduct of a first-order system whose transfer function has the following form:

$$G(P) = \frac{1}{T(P)+1} \quad (3.10)$$

Simply identify (3.10) to (.9) as follows:

$$\frac{k_p \cdot P + k_i}{aP^2 + (b+k_p)P + k_i} = \frac{1}{T(P)+1} \Rightarrow k_p \cdot T_p^2 + (k_i T + k_p)p + k_i = aP^2 + (b+k_p)P + k_i \quad (3.11)$$

Thus :

$$\begin{cases} k_p = \frac{a}{T} \\ k_i = \frac{b}{T} \end{cases} ; \begin{cases} k_{ps1} = \frac{L_{s1}}{T} \\ k_{is1} = \frac{R_{s1}}{T} \end{cases} ; \begin{cases} k_{ps2} = \frac{L_{s2}}{T} \\ k_{is2} = \frac{R_{s2}}{T} \end{cases}$$

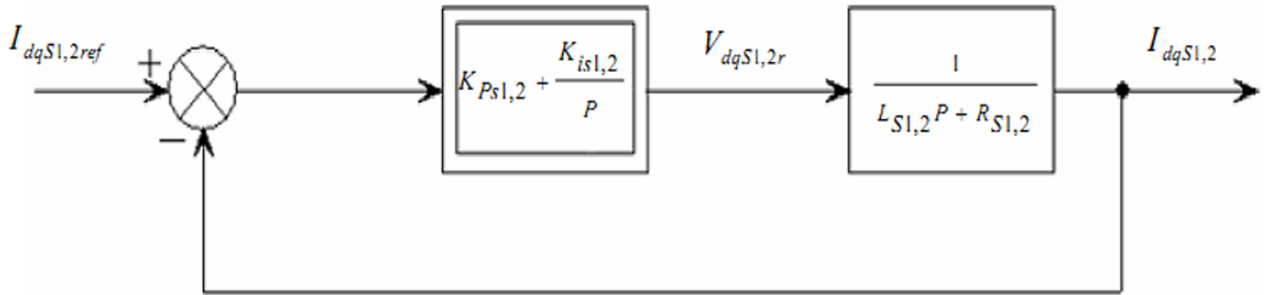


Figure 3.6: Control loop diagram for stator current regulation. [43]

3.5.3. Defluxing block

The principle of defluxing is to maintain the rotor flux at its nominal value ϕ_{rn} when operating at or below the rated speed. Ω_{rn} . [44]

$$\phi_r = \begin{cases} \phi_{rn} & \text{if } |\Omega_r| \leq \Omega_{rn} \\ \frac{\Omega_{rn}}{\Omega_r} \phi_{rn} & \text{if } |\Omega_r| > \Omega_{rn} \end{cases} \quad (3.12)$$

3.6. Numerical Simulation and Analysis of Results

- Simulation of the System Under Parametric Variations to Assess Its Robustness

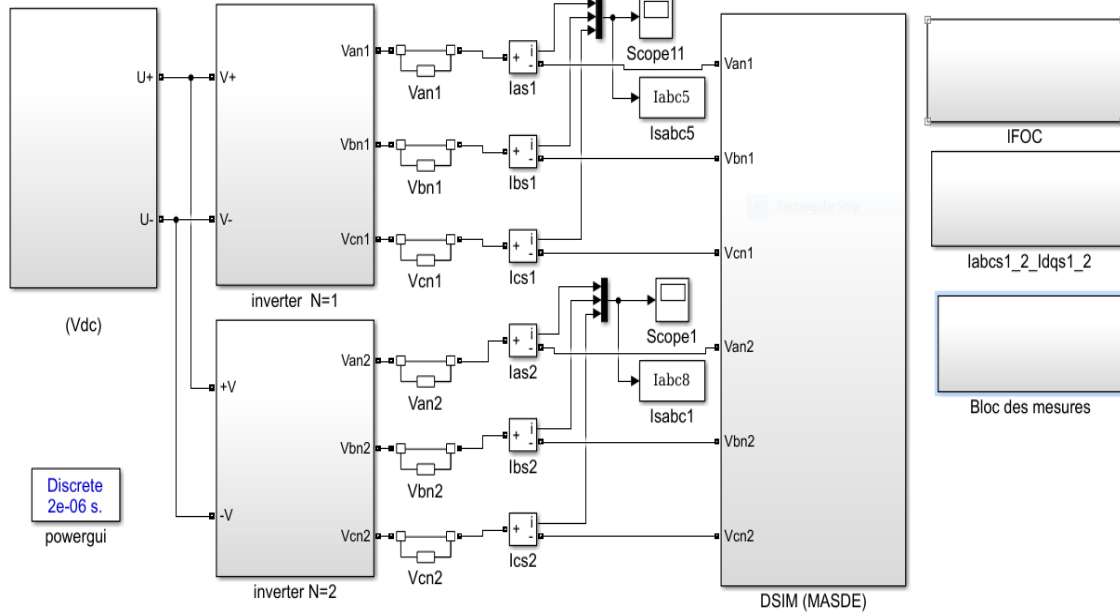


Figure 3.7: MATLAB/Simulink Model of IFOC for Dual Star Induction Machine

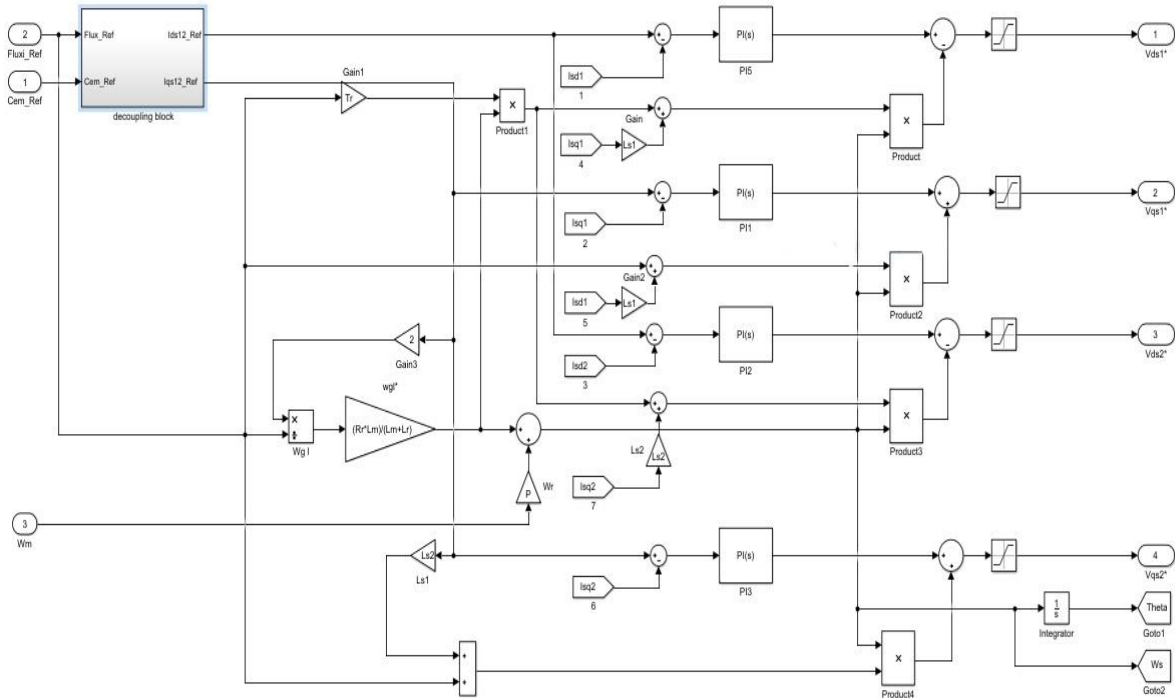
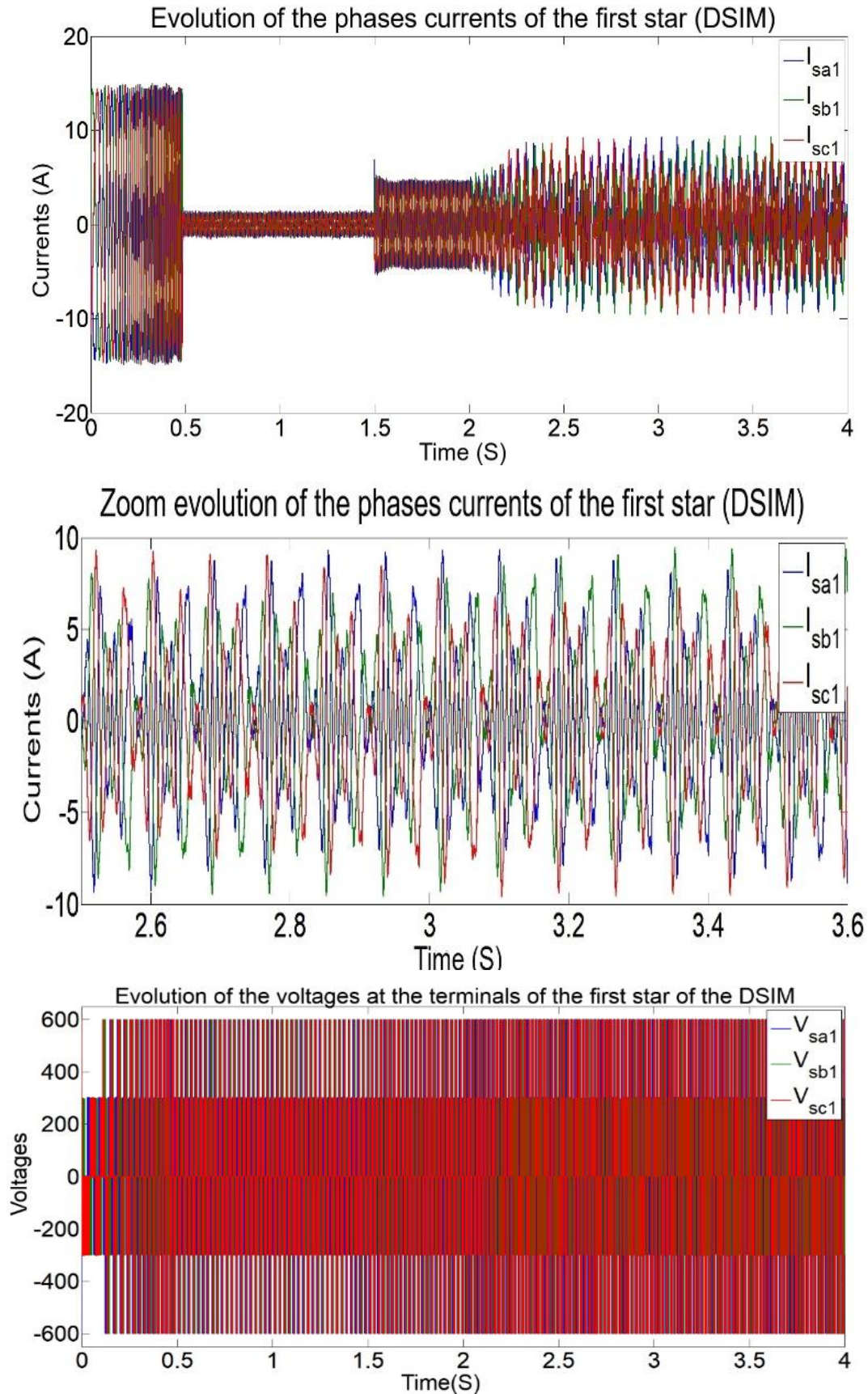
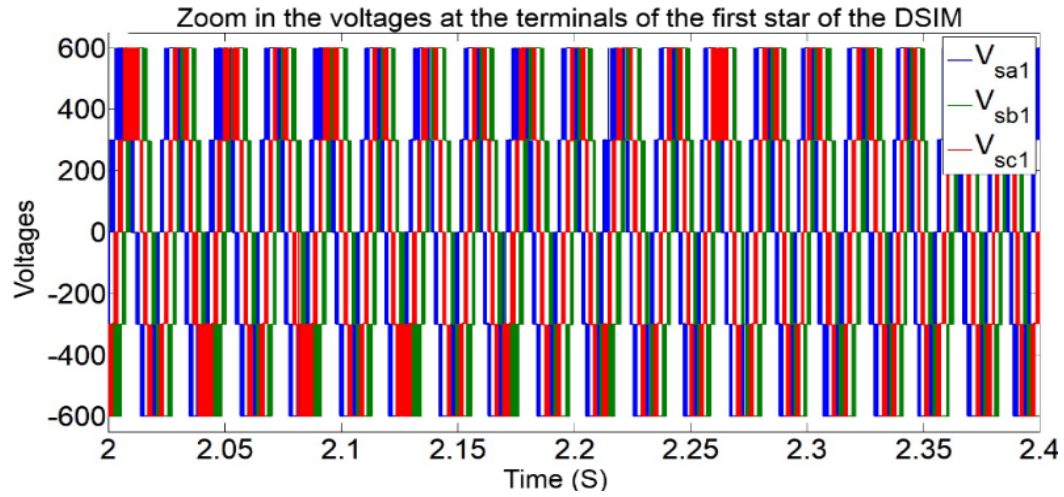
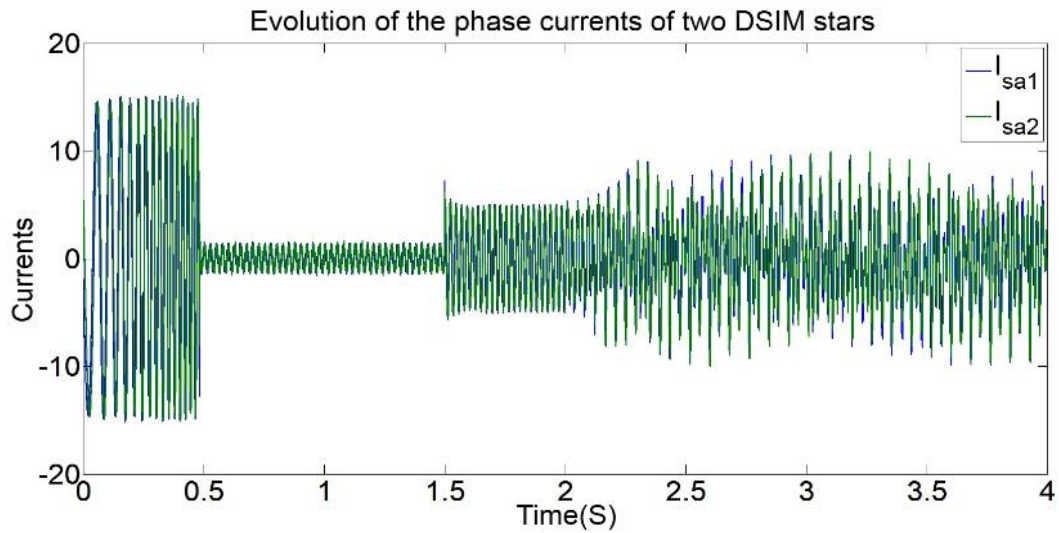
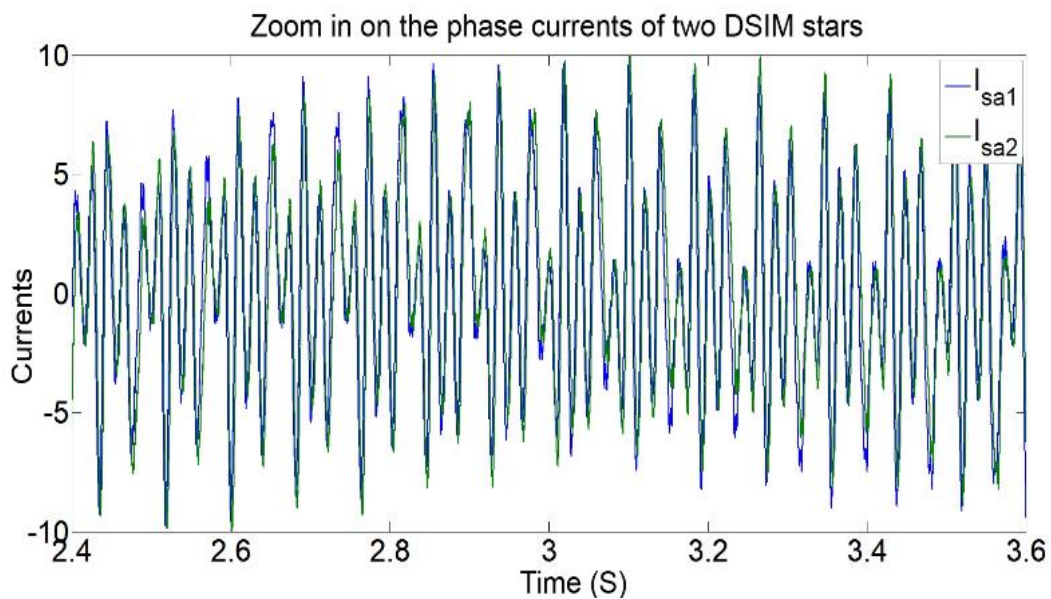


Figure 3.8: MATLAB/Simulink Block of Indirect Field Oriented Control (IFOC)





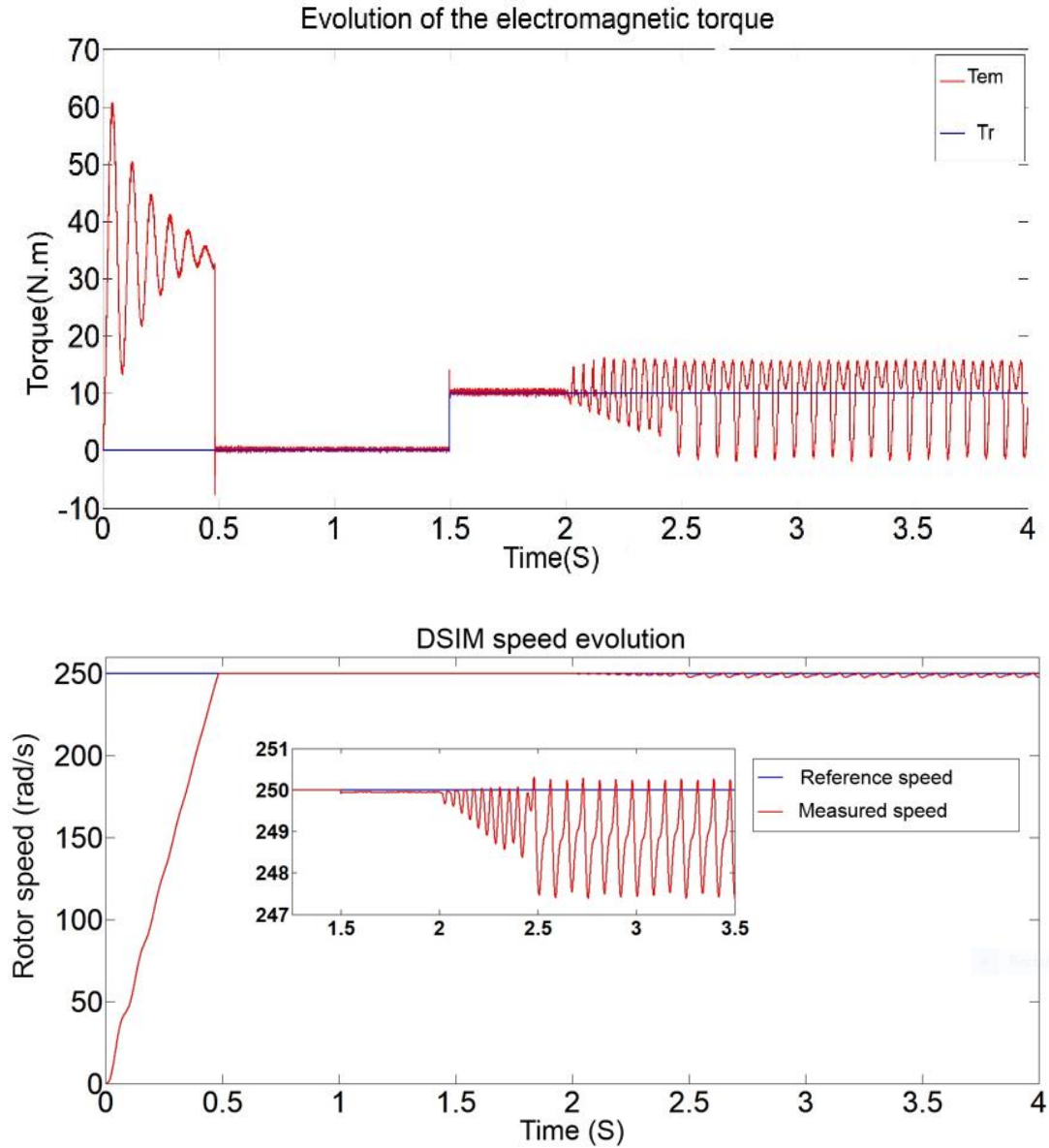


Figure 3.9: Dynamic Response of the DSIM under Indirect Field-Oriented Control (IFOC)

A robustness test was conducted on a dual star induction machine (DSIM) under vector control and fed by a three-level inverter by doubling the rotor resistance to $R_r = 4 \times R_r$. Though the system is affected by the parameter changes, it preserves its global dynamic stability.

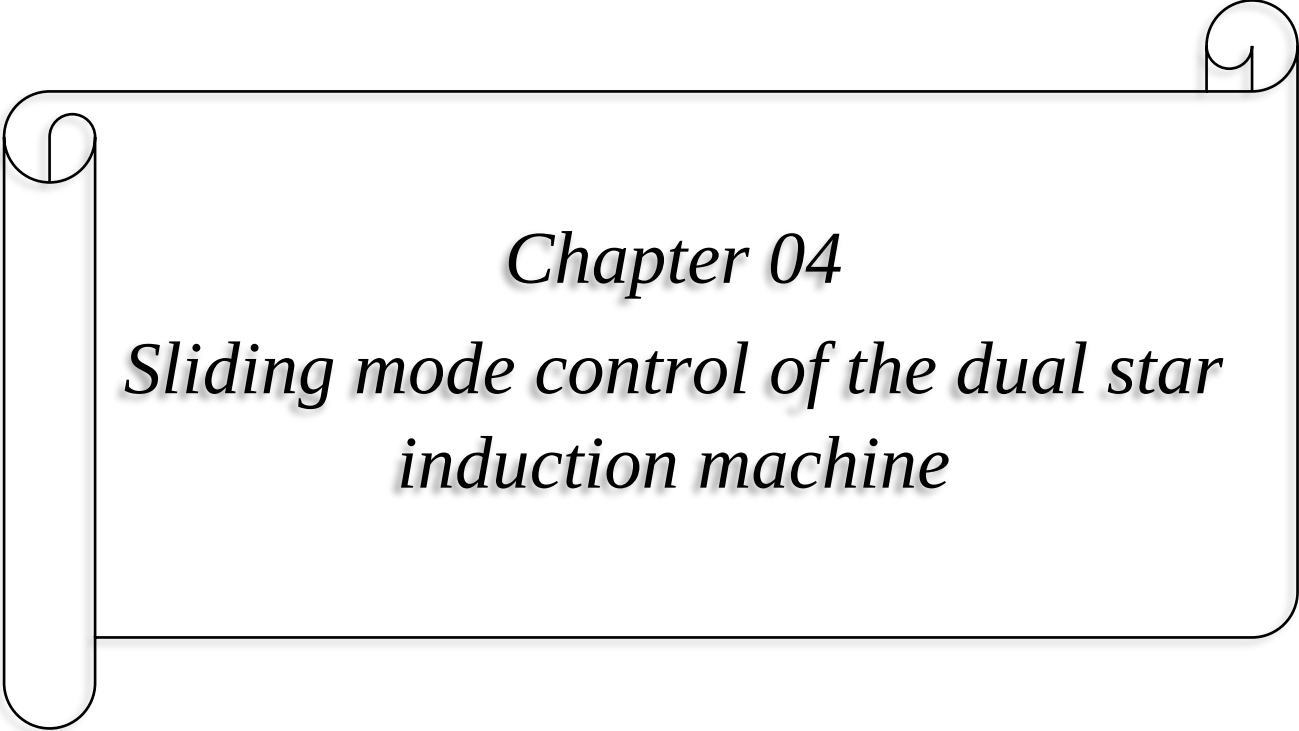
The three-phase stator currents I_{sabc} are approximately sinusoidal at ~ 50 Hz but are slightly distorted and asymmetrical, with amplitude values of ± 10.4 A, indicating sensitivity to internal variations. The I_{sd} and I_{sq} components deviate from references, especially i_{sq} , which oscillates around -6 A with large ripple, indicating partial decoupling loss.

Stator fluxes are elliptical in shape with small flattening along the direct axis and a maximum of 1 Wb, which suggests functional but less precise regulation. The speed response reaches the reference of 150 rad/s with small overshoot and negligible oscillations. The electromagnetic torque T_{em} initially reaches 28 Nm and then stabilizes around 15 Nm, however, it undergoes persistent fluctuations ± 1 Nm, indicating limited disturbance rejection capability.

In conclusion, vector control delivers adequate performance under R_r variation but is compromised by flux/current decoupling and torque precision, explaining the necessity for more resilient solutions, such as SMC.

3.7. Conclusion

In this chapter, the Field Oriented Control (FOC) strategy was studied in detail, with a focus on the Indirect FOC (IFOC) method. Key aspects included flux orientation, PI controller tuning, and implementation of defluxing. The developed MATLAB model and simulation outcomes demonstrated the efficacy of the IFOC strategy in providing precise speed control and stable machine operation under dynamic conditions.

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Chapter 04

Sliding mode control of the dual star induction machine

4.1. Introduction

These are typically found to possess some degree of robustness to parametric perturbations of the plant to be controlled while providing good dynamic response speed and accuracy. Sliding mode control is particularly appropriate for non-linear systems, being distinguished from other approaches by its lack of sensitivity to external disturbances and parametric variations. The selection of the sliding surface depends upon the specific system being considered, along with the desired performance criteria, but the selection of the control law is aimed at achieving convergence along with fulfillment of the sliding condition, which really takes care of the preference of the switching surfaces. The basic components of the sliding mode control technique are discussed in this chapter. We initially present the different sliding mode control structures, then outline the control principle and the choice of the sliding surface. This robust control method is a special kind of variable-structure control.

4.2. Theory of sliding mode control

Classical PID control laws yield good results for linear systems with fixed parameters. In the case of non-linear systems, the classical control laws may not be sufficient because they are not robust, especially when the accuracy demands and other dynamic characteristics of the system are high. The main feature of variable-structure systems is that their control laws are discontinuous.

This type of control affords a multitude of advantages which comprise robustness, accuracy, stability, simplicity and shorter response time. This makes it particularly suitable in dealing with systems whose models are not adequately known either because of parameter identification problems or because of approximations in the system model. [45]

4.2.1. Variable Structure System (VSS)

Dynamic systems with structural changes are characterized by a switching function and logic that allows them to switch between structures at any time, allowing them to possess new properties. In sliding mode control, the system's trajectory is initially forced onto a sliding surface, causing sliding motion.

The trajectory in the phase plane can be divided into three distinct modes:

1. Convergence Mode (CM): During this mode, the state variables develop from any initial condition in the phase plane towards the switching surface defined by $S(x, y) = 0$ (for two-dimensional systems). The control law and a convergence criterion govern this phase.
2. Sliding Mode (SM): When the state trajectory intersects the sliding surface, it enters the sliding mode wherein the system dynamics are confined to the surface and the state is compelled towards the origin of the phase plane. The characteristics achieved in this mode are based on the structure of the sliding surface $S(x, y)$.
3. Steady-State Mode (SSM): This mode is used to analyze the system's response in the vicinity of its equilibrium point, which is usually the origin in the phase plane. It defines the system's effectiveness and the quality of control in the steady-state mode. [46] [47]

4.3. Condition of existence of the sliding mode

After selecting the slip function, the second step is to find a control law in such a way that the state vector will move towards the surface and remain there ($S = 0$). For this, the control law should be designed in a way that S is attractive. In order to find the condition of attractiveness, the following Lyapunov function should be analyzed:

$$V(S) = \frac{1}{2} S^2 \quad (4.1)$$

A necessary and sufficient condition, called the attractiveness condition, for a sliding variable $s(x, t)$ to converge to 0 is that its time derivative is written as negative: $S \cdot S' < 0$.

When this holds, the sliding variable and its derivative have always opposite signs, and 0 is an attractive center for S . The time of convergence, also referred to as "reaching time" t_r , is directly set by the selection of S : [48]

$$S \cdot S' \leq -\mu |S|, \mu > 0$$

4.4. Principle of sliding mode control of variable structure systems

A variable-structure system is a system whose structure changes during operation. The sliding-mode control of such systems generally has two modes of operation:

- Reaching mode, or access mode, or convergence mode (MC).

- Sliding mode.

Thus, the phase trajectory, starting from any initial condition, reaches the switching surface in a finite time (non-sliding mode), then tends asymptotically towards the equilibrium point with a dynamic defined by the sliding mode. [49][50]

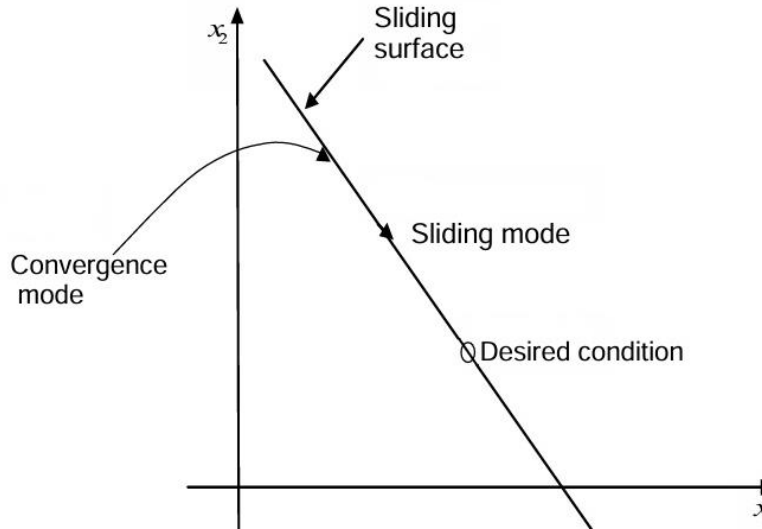


Figure 4.1: Different modes of convergence for the state trajectory

4.5. Sliding-mode control concept

Sliding-mode controller design focuses on stability and performance effectiveness, often specifying system dynamics within the convergence phase framework. It consists of continuous and discontinuous components, with the discontinuous component suppressing model inaccuracies and disturbances.

The development of this control method can be separated into three main phases:

1. Choice of surfaces,
2. Establish convergence conditions,
3. Determination of the control law.

4.5.1. Selection of the Sliding Surface

The sliding surface can be linear or non-linear, is constructed in such a way that the system always has the desired dynamics, and is generally chosen with a relative degree equal to one the

derivative of the surface $S(x)$ with respect to time explicitly shows the control U). For systems of integer order, several forms of writing are proposed in.

Consider a class of non-linear systems, where the surface vector has the same dimension as the control vector u .

$$\dot{x} = f(x, t) + g(x, t) \cdot u \quad (4.2)$$

When:

$x = (x_1, x_2, \dots, x_n)^T$: belongs to X is the system state vector.

$u = (u_1, u_2, \dots, u_n)^T$: control vector.

f, g are sufficiently differentiable vector fields defined on x .

The surface is given by:

$$S_j(x) = \sum_{i=1}^n c_{ji} \cdot e_i \quad (4.3)$$

When :

- $s_i(x) = 0$: is the switching function, which separates the state space into two disjoint parts, $s(x, t) < 0$ et $s(x, t) > 0$.
- $S(x) = [S_1(x), S_2(x), \dots, S_m(x)]$: Vector of the sliding surface.
- $x_{ref\ i} = [x_{ref1}, x_{ref2}, \dots, x_{refn}]^T$: Reference vector (desired condition).
- $e_i = x_i - x_{refi}$: The error between the variable to be controlled and its reference.
- $e_i = [e_1, e_2, \dots, e_n]^T$: Gap vector.
- C_{ij} : Sliding surface parameters.
- C : Matrix of $n \times m$ surface parameters (slope matrix of the surface line).

The surface parameters are selected in a manner that the performance specifications are achieved. Slotine provides a full equation from which the sliding surface is derived, which guarantees the convergence of a variable to its desired value.

$$S(x) = \left(\frac{\partial}{\partial t} + \lambda \right)^{r-1} \quad (4.4)$$

- λ : A positive constant.

- r : relative degree, equal to the number of times you need to derive the output to make the command appear.

S : is a linear differential equation whose only solution is: ($x = 0$, this implies that the state of the system is equal to the desired state). [50][51][52]

4.5.2. Convergence existence conditions [53]

The conditions of existing convergences are the criteria that allow in dynamics of the system to converge towards the sliding surface and to remain there independently from the perturbation. There are two considerations corresponding to the convergence mode of the system state.

a) Direct switching function

This is the first condition of convergence (Attractiveness). It was suggested and investigated by Emilyanov and Utkin. The purpose is to provide the surface with a dynamic convergent towards zero. It is established by:

$$\dot{S} > 0 \text{ if } S(x) < 0$$

$$\dot{S} < 0 \text{ if } S(x) > 0$$

This condition can otherwise be given by:

$$\dot{S}(x)S(x) < 0$$

b) Lyapunov function

The determination of the sliding domain can be reduced to an analysis of stability of the system in the sliding regime. This analysis is based on the use of the Lyapunov function. The Lyapunov function is defined as a positive scalar function ($V(x) > 0$) of the system's state variables, and it is necessary that the control law ensures a reduction of this function ($\dot{V}(x) < 0$). The underlying concept is to select a scalar function $S(x)$ in a way that the variable to be regulated gets attracted towards its desired reference value. It also requires constructing a control U in a manner that the square of the area evolves like a Lyapunov function given by:

$$V(x) = \frac{1}{2}S^2(x) \text{ by Deriving, we obtain : } \dot{V}(x) = \dot{S}(x)S(x) \quad (4.5)$$

For the Lyapunov candidate function to decay, it is sufficient to ensure that : $\dot{V}(x) < 0$

4.5.3. Determination of the control law

After the sliding mode is achieved, the system dynamics are decoupled from the control law, which is primarily used to keep the sliding conditions satisfied (i.e., the surface attractivity). Thus, the design of the sliding surface is independent of the control law. The second step involves finding the control input to compel the state trajectory towards the sliding surface and then onwards to its equilibrium point, while ensuring that the conditions for sliding mode to exist are met.

To achieve a sliding mode will typically require a discontinuous control action. The sliding surface has to be attractive from both sides. Thus, while a discontinuous control component is necessary, it does not eliminate the possibility that a continuous component may be present. Indeed, it is possible for the continuous component to effectively decrease the amplitude of the discontinuous component to any level that is desirable. In the presence of disturbances, the primary function of the discontinuous component is to provide the attractivity conditions.

In such cases, the design of a sliding mode controller involves two integral parts: one responsible for exact linearization (U_{eq}) and another one which is responsible for providing stability (U_n).

$$U = U_{eq} + U_n \quad (4.6)$$

U_{eq} is linked with the command proposed by Filipov. It is used for keeping the variable being monitored on the sliding surface $S(x) = 0$. Equivalent control is obtained under the supposition that the derivative of the surface is zero and is represented as $\dot{S}(x) = 0$, discrete control U_n is defined so that the condition of convergence is fulfilled with respect to the uncertainties in the parameters of the system model. [54] [55]

To illustrate the above development, we consider the state system. We want to determine the corresponding representation of the control U .

The derivative of the surface $S(x)$ is:

$$\dot{S}(x) = \frac{dS}{dt} = \frac{dS}{dX} \frac{dX}{dt} \quad (4.7)$$

Replacing (4.6) in (4.7) we find:

$$\dot{S}(x) = \frac{dS}{dX} [A][X] + \frac{dS}{dX} [B]U_{eq} + \frac{dS}{dX} [B]U_n \quad (4.8)$$

During sliding mode and steady state, the surface is zero, and its derivative and the discontinuous component also become zero. From this relation the equivalent control is obtained.

$$U_{eq} = - \left\{ \frac{dS}{dX} [B] \right\}^{-1} \left\{ \frac{dS}{dX} [A][X] \right\} \quad (4.9)$$

For the equivalent command to take a finite value, it must be: $\frac{dS}{dX} [B] \neq 0$

During the convergence mode, and by replacing the equivalent command by its expression in (4.8), we obtain the new expression of the derivative of the surface:

$$\dot{S}(x) = \frac{dS}{dX} [B] U_n \quad (4.10)$$

and the condition $S(x)\dot{S}(x) < 0$ become : $S(x) \frac{dS}{dX} [B] U_n < 0$

In order to satisfy the condition, the sign of U_n must be opposite to that of : $S(x) \frac{dS}{dX} [B]$

The simplest form of discrete control is that of a sign function.

$$U_n = K_x \text{sign } S(x) \quad (4.11)$$

The sign of K_x must be different from that of $\frac{dS}{dX} [B]$.

However, the latter generates a phenomenon called chattering on the sliding surface, which is generally undesirable because it adds high-frequency components to the control spectrum. [56]

▪ Chattering phenomenon

In practical applications, the ideal gliding regime is not possible due to the finite limits of the switching frequency of control devices. That is, there is no switching device that operates at infinite frequency since it would imply that one should be capable of providing infinite energy.

The inherent discontinuity in the control system produces a definite dynamic behavior in the vicinity of a boundary layer on the sliding surface that is usually termed chattering. The periodic behavior observed at the surface can be addressed by the imperfections of the switching elements or by certain technological and physical constraints, like switching time delays or hysteresis phenomena, that might produce previously unaccounted (non-modeled) high-frequency dynamics.

Chattering can degrade system performance and even lead to instability. Chattering also implies high mechanical loads on actuators, increasing energy consumption which can damage electrical power circuits. [57]

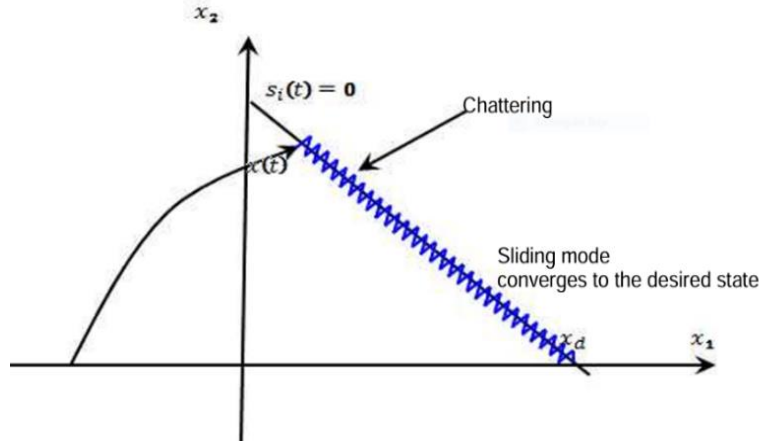


Figure 4.2: Chattering phenomenon

But it is feasible to minimize (or get rid of altogether) the interference by substituting the sign function with a reasonable saturation function which has the effect of reducing such high-frequency components. One example of a saturation function:

$$sat(S(x)) = 1 \text{ si } S(x) > \xi$$

$$sat(S(x)) = -1 \text{ si } S(x) < -\xi$$

$$sat(S(x)) = (S(x))/\xi \text{ si } |S(x)| < \xi$$

The result is the following command:

$$U_n = \begin{cases} K_x \text{ sign } S(x) & \text{if } |S(x)| \geq \xi \\ \frac{K_x}{\xi} S(x) & \text{if } |S(x)| < \xi \end{cases} \quad (4.12)$$

It can also be replaced by a function of class C1. Below is an example of such a function:

$$\text{smooth}(S(x)) = \tanh(S(x)) = \frac{e^x - e^{-x}}{e^x + e^{-x}} \quad (4.13)$$

Hence, we get the following command:
$$U_n = K_x \frac{S(x)}{\xi + |S(x)|} \quad (4.14)$$

4.6. Cascade adjustment of speed and rotor flux by sliding mode

In the case of vector control of an asynchronous motor, the use of a cascade control strategy together with sliding mode control makes it possible to realize the control system in a number of hierarchical levels. Every level is connected with one important variable that characterizes the system state: stator currents, rotor flux, and rotor speed.

The basic idea of sliding mode control is based on the description of sliding surfaces (also called switching surfaces), along which the system state dynamics are forced to evolve. When the system reaches these surfaces, its behavior is totally insensitive to uncertainties and disturbances.

4.6.1. Stator current control surface

In a vector-controlled induction motor, the stator currents in the synchronized reference frame (d-q) are directly proportional to the rotor flux (d axis) and electromagnetic torque (q axis). Their control is therefore required in order to guarantee system performance.

A- Direct current control surface

The I_d current is mostly attributed to the rotor flux. Its control keeps the flux at a predefined, usually constant, value, and thus the linearization of the electromagnetic model is provided:

$$S_d(t) = i_d(t) - i_d^*(t) \quad (4.15)$$

- $i_d(t)$: direct measured current (d axis).
- $i_d^*(t)$: direct current set point, generally derived from the flux set point ϕ_r^* .

When : $i_d^* = \frac{\phi_r^*}{L_m}$

B) Quadrature current control surface

The current i_q creates torque. The control surface cooperates with this in order to complement the setpoint torque, and this is normally determined by the speed loop:

$$S_q(t) = i_q(t) - i_q^*(t) \quad (4.16)$$

- $i_q(t)$: measured quadrature current (q axis).
- $i_q^*(t)$: quadrature current setpoint, supplied by the speed controller. [58] [59]

4.6.2. Rotor flux control surface

The rotor flux ϕ_r is not directly controlled in current sensor-only systems. Nevertheless, it is either viewed or estimated by way of a flux observer. It is typically referred to in terms of id in concepts of vector control. [60]

$$\phi_r(t) = L_m i_d(t) \Rightarrow S_\phi(t) = \phi_r(t) - \phi_r^*(t) = L_m(i_d(t) - i_d^*(t)) \quad (4.17)$$

Consequently, the rotor flux control surface is combined with direct current $S_d(t)$. In addition, the flux surface can be represented explicitly for control methods reliant on observers:

$$S_\phi(t) = \phi_r(t) - \phi_r^*(t) \quad (4.18)$$

- $\phi_r(t)$: rotor flux estimate.
- $\phi_r^*(t)$: flux setpoint (often constant for nominal operation).

4.6.3. Rotor speed control surface

The main variable monitored in drive systems is rotor speed. This variable is regulated in an external control loop, which in turn sets the setpoint for the torque current.

The slip surface corresponding to this loop is demarcated in the following way:

$$S_\omega(t) = i_\omega(t) - i_\omega^*(t) \quad (4.19)$$

- $\omega(t)$: real rotor speed (measured or estimated).
- $\omega^*(t)$: setpoint speed.

In some formulations, in order to improve dynamics and stability, a first-derivative dynamic sliding surface is defined:

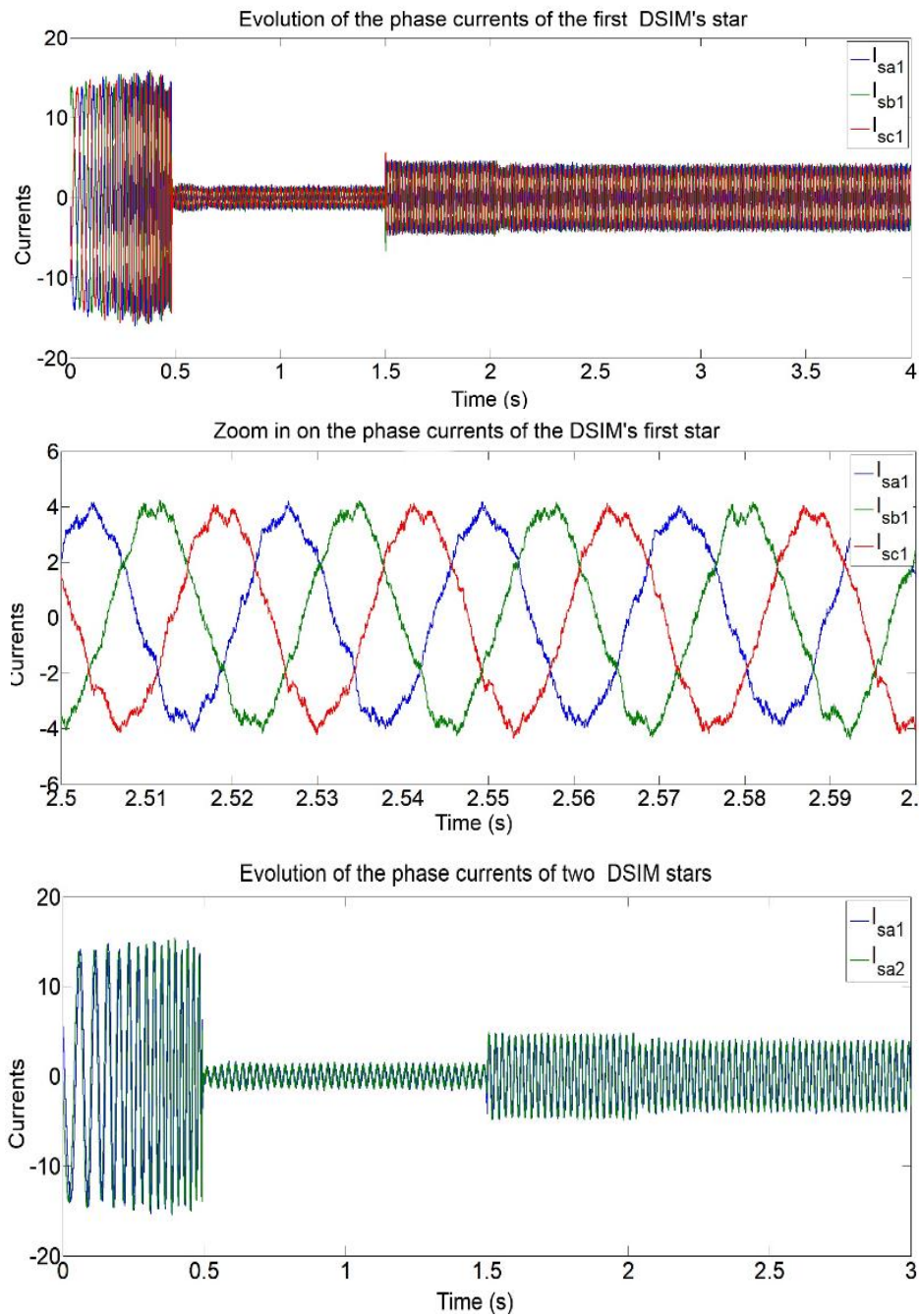
$$S_\omega(t) = \dot{e}_\omega(t) + \lambda_\omega e_\omega(t) \quad \text{with } e_\omega(t) = \omega(t) - \omega^*(t) \quad (4.20)$$

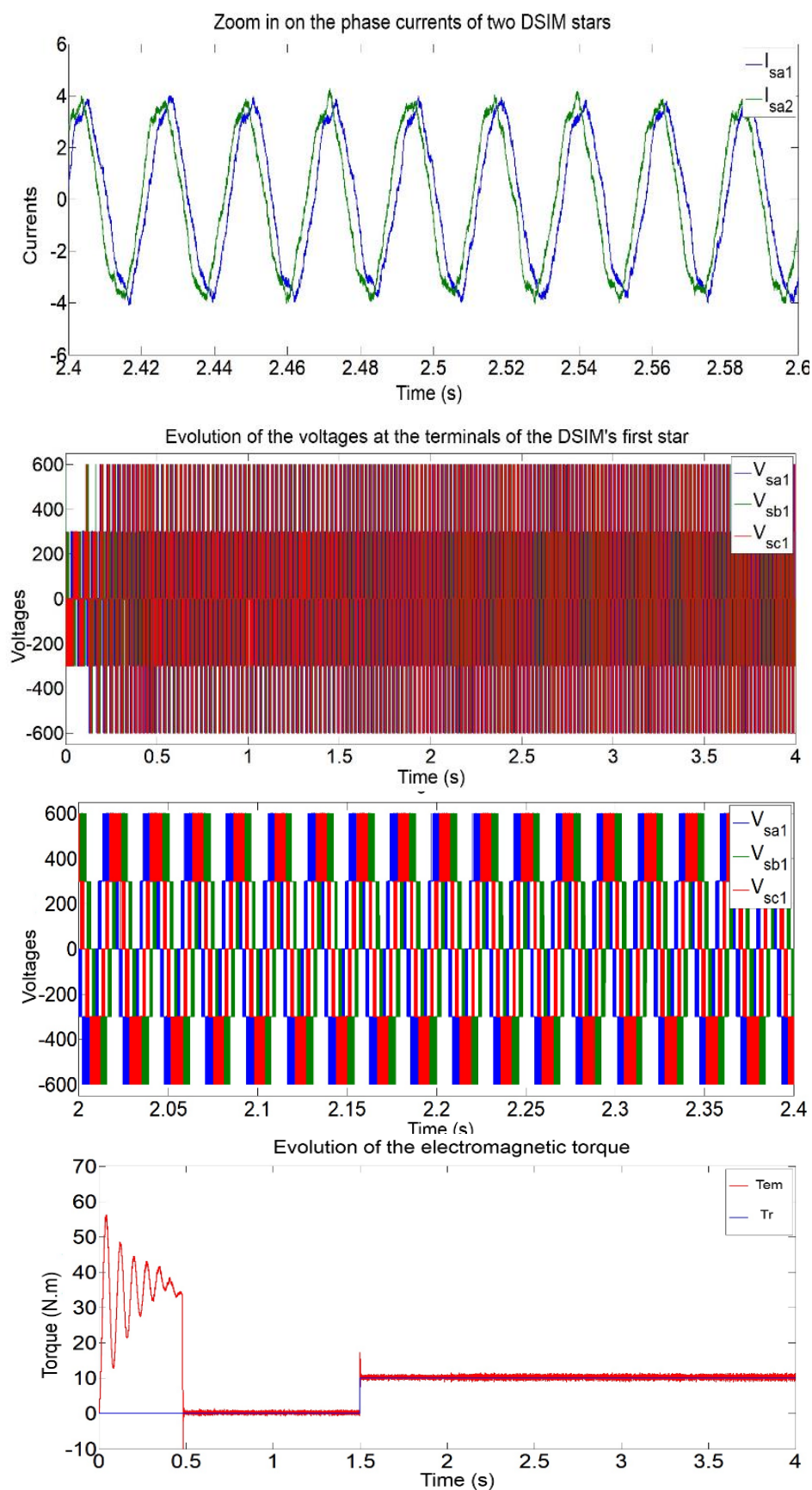
- $\lambda_\omega > 0$: convergence constant (gain).

This version allows you to control both speed and its evolution, and is used in second-order sliding mode controls or to improve tracking quality. [61]

4.7. Simulation and results interpretation

- Simulation of the System Under Parametric Variations to Assess Its Robustness





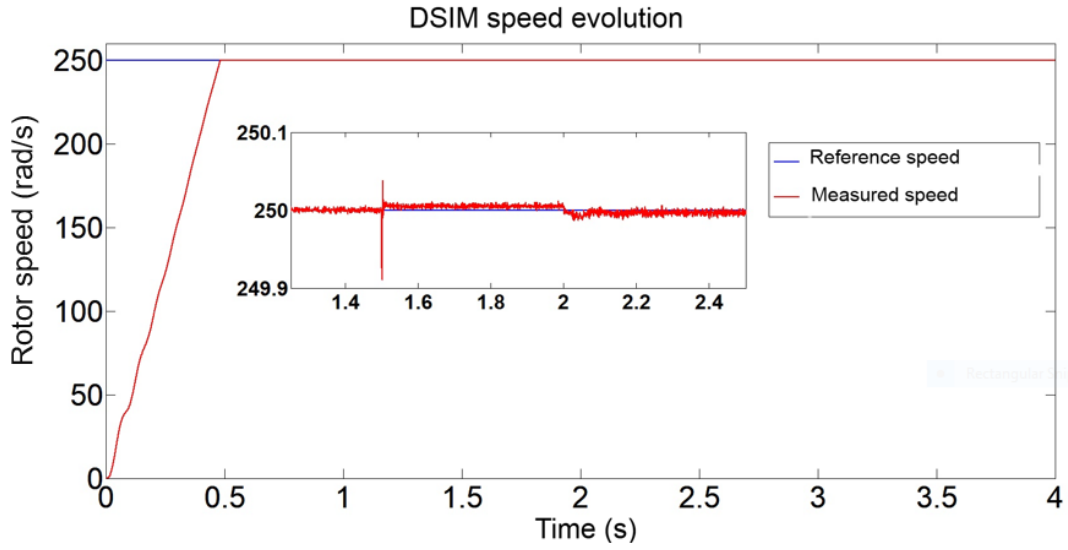


Figure 4.3: Dynamic Response of the DSIM under Sliding Mode Control (SMC)

The figures present a detailed quantitative analysis of system performance. The A1 and A2 stator currents, demonstrate symmetrical oscillation around ± 10 A, indicating balanced and steady-state operation. The electromagnetic torque experiences a moderate fluctuation around $15 \text{ N} \cdot \text{m}$ with a minimum of 13 and a maximum of $17 \text{ N} \cdot \text{m}$ ($\pm 13\%$), thereby confirming the effectiveness and stability of the Sliding Mode Controller (SMC). The rotor speed achieves 100 rad/s in a time interval of less than 0.08 seconds, with minimal overshoot and no high-level oscillations, indicating an effectively damped dynamic response.

The rotor flux is maintained constant at approximately 1 Wb, establishing correct flux orientation with the control strategy applied. The three-phase inverter voltage exhibits clear level transitions at -100 V, 0 V, and +100 V, characteristic of a three-level Neutral Point Clamped (NPC) inverter topology, which plays a significant role in Total Harmonic Distortion (THD) reduction. The control signals exhibit high-frequency switching behavior, characteristic of sliding mode operation.

The sliding surface S remains near zero-bound (amplitude < 0.05), indicating that the system trajectories move towards and remain on the sliding manifold, ensuring robust stability. Accurate speed tracking with a steady-state error of less than 1 rad/s, which translates to an accuracy of 99%. Overall, the outcomes confirm the efficacy of the SMC approach in providing

dynamic performance, high accuracy, and energy efficiency for the Dual Star Induction Machine (DSIM) operated using a multilevel inverter.

4.8. Comparison between SMC and FOC based on simulation results

The simulation findings reveal that SMC outperforms FOC in terms of both stability and precision. FOC obtains 150 rad/s with substantial overshoot, whereas SMC achieves almost flawless speed tracking with no overrun. FOC's torque peaks at 28 Nm , with $\pm 1 \text{ Nm}$ fluctuations. SMC peaks at 30 Nm with reduced $\pm 0.5 \text{ Nm}$ oscillations.

Stator currents under FOC show peaks of $\pm 10.4 \text{ A}$ and poor decoupling, whereas SMC maintains $i_{sd} \approx 1.5 \text{ A}$ steady and stable i_{sq} . Both keep flux near 1 Wb , but SMC's flux is more stable versus FOC's flattened flux. Current distribution is more balanced in SMC than FOC. Overall, SMC proves more robust under severe parameter changes compared to FOC.

4.9. Conclusion

In this chapter we have provided theory and applications of Sliding Mode Control (SMC) for the dual-star induction machine. It has reviewed the necessary background of SMC, introduced the basics of Variable Structure Systems, outlined the necessary conditions for establishing sliding mode existence, and outlined the practical approach for designing the control law and sliding surfaces for stator currents, rotor flux, and speed.

A cascade control structure was developed to achieve robust and decoupled control of the dynamic variables of the machine. The simulation results demonstrated the use of SMC and the successful achievements of fast response, high accuracy, and effective robustness against disturbances. The comparison with Field-Oriented Control (FOC) supported the better implementation of the SMC controller and consistently better performance, particularly when there are modifications to parameters and loads.

General conclusion

This study performs a comprehensive review of the dual star induction machine (DSIM) fed by a three-level NPC (Neutral Point Clamped) inverter, emphasizing on system modelling, control strategies, and performance aspects. The research progresses from theoretical analysis to simulation, presenting an organized way to define and investigate the scope of a DSIM, developing improved methods of control, reduce harmonics and investigate the performance.

To begin, the DSIM's characteristics such as greater fault tolerance, greater torque density, and greater thermal and electrical characteristics, are important for difficult cases and applications. The three-level NPC inverter is explicitly chosen priory based on the recommended voltage quality performance parameters, and it offers reduced harmonic requirements and less switching losses, all of which are critical for multiphase machines.

A complete mathematical model was developed, made possible by a state-space representation using Park's transformation. This model allowed for the accurate simulation of both the machine and the inverter dynamics. This model was then used to test the control strategies in different conditions, the focusing on two (2) control methods of interest, Indirect Field Oriented Control (IFOC) by PI regulator and by Sliding Mode Control (SMC) all of which rely on decoupling the torque control from the flux control using a different vector. IFOC with PI regulator offers semi-decoupled torque and flux control with decent dynamics performance but did have issues with parameter variations as well as external disturbances that it was sensitive to. SMC is a nonlinear, robust control that was tested to address these concerns. The simulation results showed that in ideal conditions IFOC provides a smooth output, but SMC provided a better tracking output, faster output response time, and better disturbance rejection performance, particularly when there was uncertainty in the system parameters and load changes.

The thesis demonstrates that SMC is a more reliable alternative for controlling a Dual-star Induction Machine (DSIM) than conventional IFOC, especially in harsh conditions. This work also lays a solid groundwork for future research, including real-time implementation and applications in fields such as renewable energy and electric mobility.

Appendices

Parameter	Description	Value	Unit
R_{s1}	Stator resistance of the first stator winding	3.72	Ω
R_{s2}	Stator resistance of the second stator winding	3.72	Ω
R_r	Rotor resistance	2.12	Ω
r_r	Rotor resistance (alternate notation)	2.12	Ω
L_{s2}	Stator inductance of the first stator winding	0.022	H
L_2	Stator inductance of the second stator winding	0.022	H
L_r	Rotor inductance	0.006	H
L_m	Magnetizing (mutual) inductance	0.3672	H
L_{s1}	Stator leakage inductance	0.032	H
f_s	Supply frequency	50	Hz
β	Phase displacement angle between stator windings	$\pi/6$ (≈ 0.5236)	rad
p	Number of pole pairs	1	—
J	Moment of inertia of the rotor	0.0662	$\text{kg}\cdot\text{m}^2$
f	Mechanical friction coefficient	0.001	—
ω_n	Natural frequency	296.98	rad/s
k_{pd}	Proportional gain for derivative action	0.78	—
k_{id}	Integral gain	2.48	—
L_ω	Stator winding inductance or leakage reactance	0.0826	H
P_n	Nominal power	4.5	kW

Table 01: DSIM parameters and referenced units

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ملخص

نقدم في هذا البحث دراسة محاكاة لالة كهربائية مزدوجة النجم اللاتزامنية من اجل الحصول على أداء جيد للماكينة تم استعمال تغذية بواسطة اثنين من المحولات متعددة المستويات بالتحديد ثلاثية المستويات، تم في البدء اختبار أداء الماكينة بواسطة التغذية فقط دون تحكم. بالنسبة للتحكم تم تطبيق التحكم الكلاسيكي وهو التحكم الشعاعي الغير مباشر من اجل تقديم أداء أفضل وتحسين عمل الماكينة تم بعدها اقتراح طريقة أخرى أكثر فعالية وتطور للتحكم وهو التحكم الانزلاقي بهدف التحصل على نتائج أفضل من استراتيجية التحكم السابقة تم تطبيقها على الماكينة ومراقبة النتائج. كما تم ادراج اختبار المتانة والأداء المتمثل في تغيير قيم الماكينة لكل طريقة تحكم ومن ثم مقارنة النتائج مع بعضها والتعليق على كلا الأداءين المتحصل عليهما.

Résumé

Dans ce travail, nous réalisons une étude de simulation d'une machine asynchrone double étoile, alimenté par deux onduleurs multiniveaux, en particulier à trois niveaux., Afin d'obtenir de bonnes performances de la machine.

Au début, on évalue le comportement de la machine en fonctionnement libre, c'est-à-dire sans commande. à partir de là, on implémente une commande classique qu'est la commande vectorielle indirecte pour augmenter les performances et le comportement dynamique de la machine, Ensuite, une commande plus robuste et performante, la commande par mode glissant, a été implémentée dans le but de surpasser les résultats obtenus avec la commande vectorielle indirecte et d'évaluer ses performances dynamiques. Un test de robustesse et de performance sont effectués en faisant varier les paramètres de la machine pour chaque stratégie de commande. Les résultats obtenus sont ensuite comparés et analysés dans le but de mettre en exergue les atouts de chacune des approches ainsi que leurs limites.

Abstract

In this work, we present a simulation study of a dual star induction machine powered by two multilevel inverters, specifically of three levels, with the aim of improving its performance. At first, the characteristics of the machine were evaluated in free operation, that is, without any control. then, a classic control method was used, which is Indirect Field Oriented Control to enhance the performance and dynamics of the machine. Subsequently, a more advanced control strategy sliding mode control was applied to the machine to evaluate its performance in comparison with the conventional control method. Robustness and performance tests were undertaken for each control strategy by varying machine parameters. The results obtained were then compared and assessed to show pros and cons of each method.