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**Amélioration des Performances de La Commande DPC D'un
Actionneur Double Etoile Fonctionnement Moteur Génératrice
Alimenté par des Convertisseurs Matriciels**

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**The Performance Improvement of DPC Control of
Dual Stator Induction Machine (Motor /Generator)
Supplied by Matrix Converters**

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ملخص: الهدف الرئيسي من هذه الأطروحة يركز على تعزيز وتحسين أداء التحكم المباشر في عزم الدوران لآلة مزدوجة غير متزامنة تعمل كمحرك ومولد، والتي يتم تغذيتها بموجات مصفوفية. في الجزء الأول من هذه الأطروحة، قمنا بتقديم نبذة حول الموجات المصفوفية ومختلف الاشكال وطرق التحكم المقترحة في البحوث ووضع نماذج لها أولاً. بعد ذلك قدمنا تقنية التحكم المباشر في عزم الدوران لصنفين مختلفين من المموج المصفوفي يعرفان بالمموج المصفوفي المباشر وغير المباشر لتشغيل الآلة. تم إجراء دراسة مقارنة بين نتائج المحاكاة التي تم الحصول عليها. بعد ذلك، تم اقتراح تقنية ذكية تعتمد على المنطق الضبابي، حيث تم استبدال جدول التبديل والتحكم التخلفي بنظام الاستدلال الضبابي. لإظهار أداء الطريقة المقترحة، تم تقديم نتائج المحاكاة. من أجل تحسين أداء مخطط التحكم مثل تقليل تموجات عزم الدوران والتدفق، وتقليل التشوه التوافقي المنخفض الذي يظهر في تيار الإدخال والذي ينتج عن عدم وجود التزامن بين مرحلتي المموج المصفوفي الغير مباشر، تم اقتراح التقنية التكيفية للتحكم المباشر في عزم الدوران التي تعتمد على التحكم الشعاعي والمحسن باستخدام خوارزمية الخفافيش. بالإضافة إلى ذلك، تم تحسين أداء نظام التحكم عن طريق إدخال وحدة تحكم تأزرية قوية في الحلقة الخارجية لتنظيم السرعة. في الجزء الثاني من هذه الأطروحة، تم تطوير تحكم قوي يعمل على تحويل طاقة الرياح استناداً إلى مولد مزدوج لا تزامني والموصولة بالشبكة عن طريق المموج المصفوفي المتعدد المستويات باستخدام التحكم المباشر الجزئي. بالإضافة إلى ذلك، ومن أجل تحسين نقل الطاقة في أنظمة الرياح، تم اقتراح وحدة تحكم تكيفية جزئية. تظهر النتائج التي تم الحصول عليها أن مخطط التحكم المقترح يظهر أداء عال في دقة التحكم من خلال نتائج المحاكاة.

كلمات مفتاحية: المموج المصفوفي، التحكم المباشر في عزم الدوران، التحكم الشعاعي، محرك مزدوج غير متزامن، مولد مزدوج غير متزامن، التحكم المباشر الضبابي في عزم الدوران، التحكم الشعاعي في عزم الدوران، التحكم التأزري، التحكم الكسري، التحكم التكيفي.

Abstract: The main objective of this thesis is to improve the performance of Direct Torque Control (DTC) for a Dual Stator Induction Machine (DSIM) operating as a motor and generator fed by a Matrix Converter (MC). In the first part of this thesis, a state-of-the-art on matrix converter outlines its various topologies, models, and modulation techniques has been conducted. Then, we have given the DTC of two different topologies of MC known as Direct and Indirect MC to feed the DSIM. A comparative simulation study between two topologies was carried out. Therefore, an intelligent DTC based on fuzzy logic control of DSIM was presented, where the switching table and hysteresis controller have been replaced by a fuzzy inference system. To show the performance of the proposed method, simulation results were provided. In addition, in order to enhance the control scheme performances such as reducing the torque and flux ripples, and minimizing the low harmonic distortion that appears in the input current and which is caused by the lack of synchronization between the two stages of the IMC, an adaptive optimized DTC-SVM using Metaheuristic Bat Algorithm (MBA) was proposed. To increase the robustness of the control scheme, a robust Synergetic Controller (SC) was inserted in the outer loop for speed regulation. The second part of this thesis is to develop a robust control of wind energy conversion (WEC) based on a Dual Stator Induction Generator (DSIG) by using a fractional DTC-SVM for a multilevel matrix converter. Moreover, to optimize the transfer of energy in wind systems, a fractional Adaptive Synergetic Controller (FASC) was suggested. The obtained results show that the suggested control scheme makes it possible to obtain significantly improved control performance.

Keywords: Matrix Converter (MC), Direct Torque Control (DTC), Space Vector Modulation (SVM), Double Star Induction Motor (DSIM), Double Star Induction Generator (DSIG), DTC-Fuzzy control, DTC-SVM Control, Synergetic Control (SC), Fractional Control, Adaptive Control.

Résumé : L'objectif principal de cette thèse est d'améliorer les performances du Contrôle Direct du Couple (DTC) pour une Machine Asynchrone Double Etoile (MASDE) fonctionnant en tant que moteur et génératrice alimentés par un Convertisseur Matriciel (MC). Dans la première partie de cette thèse, un état de l'art sur les convertisseurs matriciels en décrivant les différentes topologies, modèles techniques de modulation a été présenté. Ensuite, nous avons présenté la DTC pour deux topologies différentes de MC appelées Direct et Indirect MC pour alimenter la MASDE. Une étude par simulation comparative entre les deux topologies a été effectuée. Par conséquent, un DTC intelligent basé sur le contrôle par la logique floue de la MASDE a été présenté, où la table de commutation et le contrôleur d'hystérésis ont été remplacés par un système d'inférence floue. Pour montrer les performances de la méthode proposée, des résultats de simulation ont été fournis. De plus, afin d'améliorer les performances du schéma de contrôle telles que la réduction des ondulations de couple et de flux et la minimisation de la faible distorsion harmonique dans le courant d'entrée causé par le manque de synchronisation entre les deux étages de l'IMC, un DTC-SVM optimisé adaptatif utilisant l'algorithme Metaheuristique de Chauve-Souris (MBA) a été proposé. Pour augmenter la robustesse du schéma de contrôle, un contrôleur synergétique robuste (SC) a été inséré dans la boucle externe pour la régulation de la vitesse.

La deuxième partie de cette thèse est de développer un contrôle robuste de la conversion d'énergie éolienne (WEC) basé sur une Génératrice Asynchrone Double Etoile (GASDE) en utilisant un DTC-SVM fractionnaire pour un convertisseur matriciel multiniveaux. De plus, pour optimiser le transfert d'énergie dans le système éolien, un contrôleur synergétique adaptatif fractionnaire (FASC) a été suggéré. Les résultats obtenus montrent que le schéma de contrôle proposé permet d'obtenir des performances de contrôle significativement améliorées.

Mots-clés : Convertisseur Matriciel (CM), Commande Directe du Couple (DTC), Modulation Vectorielle (SVM), Machine Asynchrone Double Etoile (MASDE), Génératrice Asynchrone Double étoile (GASDE), Commande DTC-Floue, Commande DTC-SVM, Commande Synergétique (SC), Contrôle Fractionnaire, Contrôle Adaptatif.

To my parents

To my brothers and sisters

To my friends

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Nomenclature

Acronyms / Abbreviations

- MC Matrix Converter
- IMC Indirect Matrix Converter
- ITLMC Indirect Three Level Matrix Converter
- MMMC Multi modular matrix converter
- DZSMC Direct Z-source matrix converter
- DSIM Dual Stator Induction Motor
- DSIG Dual Stator Induction Generator
- DTC Direct Torque Control
- SVM Space Vector Modulation
- ISVM Indirect Space Vector Modulation
- SC Synergetic Control
- V_{dc} DC bus Voltage
- IGBT Insulated Gate Bipolar Transistor
- MPPT Maximum Power Point Tracking
- PI Proportionnel Integral
- API Adaptive Proportionnel Integral
- FPID Fractional Proportionnel Integral Derivative
- THD Total Harmonics Distortion

- TSR Tip Speed Ratio
- WECS Wind Energy Conversion System
- FASC Fractional Adaptive Synergetic Control
- TF Transfer Function
- FLC Fuzzy Logic Control
- MBA Metaheuristic Bat-Algorithm
- HC Hysteresis Controller
- BJT Bipolar junction transistor
- IGBT Insulated-gate bipolar transistor

Notations

- C_P Power Coefficient
- $C_{P_{max}}$ Optimal power coefficient
- β Pitch angle
- λ Tip speed ratio
- λ_{op} Optimal tip speed ratio
- ν_{wind} Wind velocity
- $\gamma(\gamma = 30^\circ)$ Electrical angle between the stators.
- T_{em} Electromagnetic torque
- k_f Friction Coefficient
- G Gain of Gearbox
- J Total inertia
- λ Fractional integral order
- μ Fractional derivative order
- $\varphi_{as1}, \varphi_{bs1}, \varphi_{cs1}$ Own flux circulating in the stator 1
- $\varphi_{as2}, \varphi_{bs2}, \varphi_{cs2}$ Own flux circulating in the stator 2

- $\varphi_{ar}, \varphi_{br}, \varphi_{cr}$ Own flux circulating in the rotor
- $i_{as1}, i_{bs1}, i_{cs1}$ Phase currents of star 1
- $i_{as2}, i_{bs2}, i_{cs2}$ Phase currents of star 2
- L_m Cyclic mutual inductance between stator and rotor
- L_r Inductance proper of phase of the rotor
- L_s Inductance proper of each phase of identical stars 1 and 2
- M_s Mutual inductance between two phases of star 1
- Ω_s Synchronous angular velocity
- Ω_{mec} Machine rotation speed
- ω_{coor} The rotational speed of Park's frame (d, q)
- P_m Mechanical power
- R Blade length
- ρ Volumic mass of air
- R Section of the surface swept by the blades.
- R_{ar}, R_{br}, R_{cr} Rotor phase resistances
- $R_{as1}, R_{bs1}, R_{cs1}$ resistances of stator 1
- $R_{as2}, R_{bs2}, R_{cs2}$ resistances of stator 2
- v_{g1}, v_{g2}, v_{g3} Grid voltages
- θ_{s1} The angle between the magnetic axis of the phase a_1 and the axis d
- θ_{s2} The angle between the magnetic axis of the phase a_2 and the axis d
- θ_r The angle between the magnetic axis of the phase a_r and the axis d

General Introduction

Power electronics is a technology that combines energy, electronics, and control methods to transfer energy between a source and a load [1].

Since the great progress of industry and more precisely semiconductor technology and microcomputing, power electronics continues to take over in multiple industrial applications like electrical applications [2, 3].

Indeed, various applications such as heating, railway and maritime electric traction, renewable energies, electric welding and harmonic filtering of networks, would not have been part of the modern world without the intervention of power electronics [4].

Most of these industrial applications where power electronics intervenes, are based on complex mechanisms using electrical machines to contribute to the start-up of large electrical, mechanical or thermal systems. These machines often operate under variable speed especially in the transport sector [5–7].

Along this path, the electronic control of variable-speed electric motors has brought enormous advantages to industrial processes, increasing their possibilities and performance, facilitating their automation, reducing their maintenance, energy consumption and increasing their availability rate [8, 9].

Indirect converters (buck-to-buck) were the first power converters, used to power electric motors. Nevertheless, because of their size, their bulkiness and their reduced lifespan, extensive research has been carried out in order to overcome these problems, and find other alternatives to these converters [10, 11].

The direct converters or matrix converters were the fruit of this research, they have the following advantages [8, 9]:

- reduced size and long service life (i.e. greater reliability). This is due to the lack of a DC intermediate circuit, so there is no storage element like electrolytic capacitors. Their dynamics is faster.
- functional reversibility in four quadrants.

- adjustable input power factor is independent of the load.
- High quality input and output signals

However, these converters are characterized by their limitation of ratio, moreover, the disturbances on their input influence their general behavior [3, 6, 12].

Regarding the choice of control techniques used in the control of the various structures of converters studied in this thesis, we can cite in particular the choice of the control technique called "direct torque control (DTC)". This technique has already shown its superiority compared to other techniques, especially vector control. Unlike the latter, which is often based on rigorous mathematical calculations, the DTC technique is at its origin based on a qualitative and simplified knowledge of the behavior of the machine. Often control actions are based on simple hysteresis controllers and switching tables.

The research work presented in this thesis is a continuation of the research work that is directly related to the matrix converter and the DTC. The main goal part of this work is the study, modulation, and control as well as the synthesis of the different topologies of matrix converters associated with the dual stator induction machine. Indeed, we will be interested in the three topologies of the matrix converter which are the direct, indirect, and indirect three-level topologies. We develop their modulation and control techniques by associating them with the dual stator induction machine.

This thesis is organized into 2 parts. The first part contains the control of the dual stator induction machine with a motor mode powered by the direct and indirect matrix converter, which includes chapter 1, chapter 2, chapter 3 and chapter 4. The second part includes the control of a dual stator induction machine with a generator mode integrated into the grid by a three-level matrix converter, which includes chapter 5 and chapter 6. The first chapter of this thesis is dedicated to the state of the art of the matrix converter by citing the different topologies and modulations proposed in the scientific literature. Then, we discuss its switching mode of operation, filtering and protective circuits, and modeling. The chapter ends with a brief review of the industrial implementation of this converter.

The second chapter develops the direct torque control of the matrix converter. We start by presenting and simulating direct torque control (DTC), explaining the principle of electromagnetic flux and torque control. Then we adapt the DTC algorithm for the two studied topologies of the MC (direct and indirect). A comparative analysis of the electromagnetic and electrical performances is presented for the two controls.

In the third chapter, a study of the DTC based on fuzzy logic is devoted. First, we start

by giving an overview about fuzzy logic control, then, we present the implementation of the fuzzy control to adjust the control of indirect matrix converter that feeding the dual stator induction motor. Then a numerical simulation of the same environment will be carried out and the results found will be presented and discussed at the end of the chapter.

In fourth chapter, a DTC based on Space Vector Modulation (SVM) with the synchronization between the two stages of IMC is presented. On the other hand, an adaptive PI controller based on a gradient descent algorithm using Metaheuristic Bat Algorithm (MBA) is offered to generate the desired stator voltages and the switching states of the inverter stage via the SVM. A robust Synergetic Controller (SC) for speed regulation is also added to the outer loop to improve the performance of the control system.

The fifth chapter aims to present the modeling and control of the wind turbine with a variable speed. Thus, modeling its different components is shown by introducing the notions of power coefficient and maximizing the energy extracted from the wind by MPPT control.

In the last chapter, A fractional DTC-SVM is proposed for improving the execution and competency control of wind power generation systems with a dual stator induction generator (DSIG) coupled to the grid via a multilevel matrix converter. Then, a Fractional Adaptive MPPT Synergetic Controller (FASC) for optimizing wind energy is developed. To evaluate the effectiveness of this control in the second operational zone of the wind turbine, when the pitch angle is thought to be fixed, a significant numerical simulation is done and discussed.

Finally, we enclose this work with a general conclusion that summarizes the results obtained and the perspectives for the future continuity of this work.

Scientific Contributions

Journal papers

1. Yahia Moati, Katia Kouzi, and Atif Iqbal. "Adaptive optimized DTC-SVM using metaheuristic Bat algorithm for DSIM fed by IMC based on robust synergetic speed controller." *International Journal of Transactions on Electrical Energy Systems* (2021).
2. Moati, Yahia, and Katia Kouzi. "Investigating the performances of direct torque

and flux control for dual stator induction motor with direct and indirect matrix converter.” *Periodica Polytechnica Electrical Engineering and Computer Science* 64.1 (2020): 97-105.

3. Moati, Yahia, and Katia Kouzi. ”An Efficient of Direct Torque Control of Indirect Three Level Matrix Converter Fed Dual Stator Induction Motor Based on Synergetic Controller An Efficient of Direct Torque Control of Indirect Three Level Matrix Converter Fed Dual Stator Induction Motor Based on Synergetic Controller.” *Journal Européen des Systèmes Automatisés* 53.5 (2020): 617-627.

International Conference Papers

1. Yahia Moati, Katia Kouzi, Khalil Mokhtari, and Saddam Bensaoucha: Direct Torque Control to Improve the Performances of the DSIM Powered by Indirect Matrix Converter. *International Conference on Communications and Electrical Engineering (ICCEE18)*. University El-Oued - Algeria - December 17-18, 2018.
2. Yahia Moati, Katia Kouzi, and Khalil Mokhtari: Fixed Switching DTC of DSIM Powered by NPC-Matrix Converter. *The International Conference on Sustainable, Renewable Energy Systems and Applications (ICSRESA'19)*. University of Tebessa - Algeria - December 04-05, 2019.
3. Yahia Moati, Katia Kouzi, and Khalil Mokhtari: Improvement of DTC of Wind Energy Conversion System Based on DSIG using Matrix Converter. *The International Conference on Sustainable, Renewable Energy Systems and Applications (ICSRESA'19)*. University of Tebessa - Algeria - December 04-05, 2019.
4. Yahia Moati, Katia Kouzi, Aissa Bencherif, and Saddam Bensaoucha: New Design of Neural Direct Power Control of DSIM Fed by Indirect Matrix Converter. *International Conference in Artificial Intelligence in Renewable Energetic Systems (IC-AIRES 2018)*. Tipasa - Algeria - 24 November 2018.

National Conference Papers

1. Yahia Moati, Katia Kouzi, Messaoud Linani, Nouredine Gazzam, Oussama Maroufi: Intelligent Direct Torque Control of Six Phases Induction Motor Powered by tow stage Matrix Converter. *The 1th National Conference on Electrical Engineering and its Applications (NCEEA'01)*. Guelma- Algeria-15 November 2022.

Part I

Control of Dual Stator Induction Motor Fed by Matrix Converter

Chapter 1

Generality and State of the Art on Matrix Converters.

“ Seek knowledge, and if you are unable, then love those who have it, and if you do not love them, then do not hate them. ”

Abu Al-Darda

1.1 Introduction

The matrix converter (MC) is a direct AC to AC converter. It was proposed in 1976 by Guygyi-Pelly [13, 14]. Also is called a forced commutated converter, it invented on the basis of cyclo-converters. It uses a $n \times m$ matrix of bidirectional power switches to directly connect the n input phases with the m output phases. Its switches are bidirectional so they can conduct current in both directions and block voltages in both polarities, which can generate sinusoidal AC outputs controllable in terms of magnitude and frequency from an AC source [14, 15].

This type of converter has been the subject of research and development for around thirty years because of its possibility of replacing conventional indirect converters (AC-DC-AC) which use storage capacitive DC bus, particularly for on-embedded systems . It has several advantages over the latter [7]. It is more reliable and less bulky due to the absence of a DC capacitor and it can be encapsulated in a semiconductor module. Its command is instantaneous and it is controllable at the input (current phase shift with

respect to the input voltage) and at the output (output voltage). On the other hand, it has two major problems which are the limit of the voltage ratio (for normal modulations) and the influence of the input on the output [6, 16].

Bellow, we briefly present the state of the art on the different topologies, the modeling of this converter with the connection functions, and the modulations available through the study of the switching as well as the protections of this converter.

1.2 State of the art on Matrix Converter Topologies

In this part, a review of the main topologies of the matrix converter proposed in the literature is presented. Figure 1.1 shows an overall diagram of the main families of AC-AC converters.

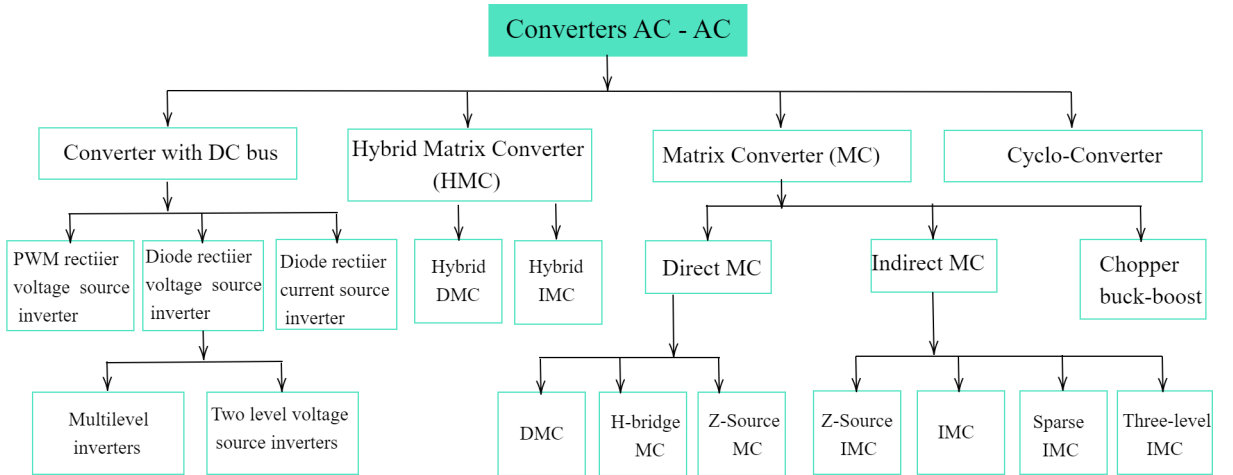


Figure 1.1: Classification of AC-AC converters.

The first topology, which was tested in 1976 by Jones and Boss, is the direct MC (3x1) topology based on BJT [14]. In 1980, Venturini and Alesena proposed the first direct MC (3x3). They described it as a matrix of bidirectional switches (figure 1.2) and named it “matrix converter” [13].

There is another structure of direct MC that alternates a current source with a voltage source (figure 1.3). The objective is to control the amplitude and frequency of the output current and allow to have voltage ratios greater than 1 [17].

This topology has not been considered in the literature because of the difficulty of realizing a current source, overvoltages caused by the input inductance (which is very large), and the dependence between the output voltage and current [9, 17].

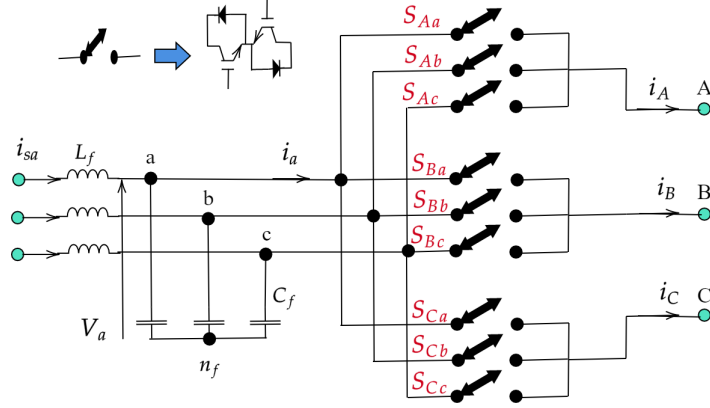


Figure 1.2: Direct matrix converter.

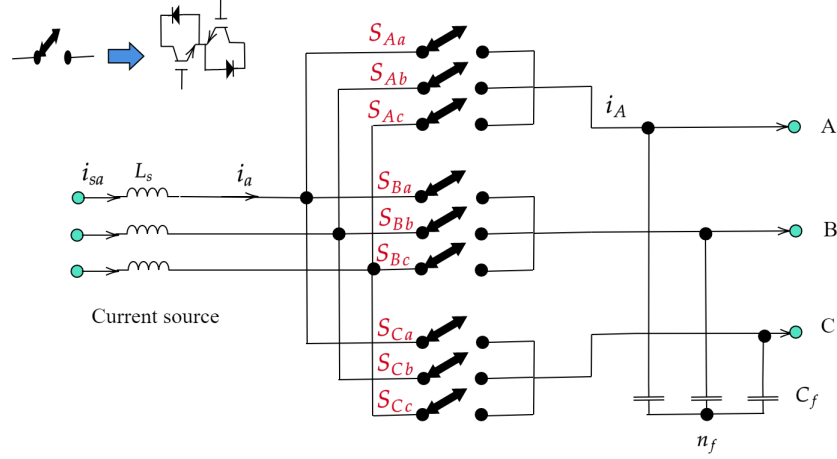


Figure 1.3: Direct current MC.

In 1989 the indirect matrix converter is suggested by Holtz and Boelkens [18]. In 1995 the topology of IMC is used, Huber et al [19]. Showed the possibility of obtaining input and output signals similar to those obtained by a direct topology, this topology is the physical implementation of the indirect model of MC. It contains two stages, the first stage is the rectifier stage which is a matrix converter [3x2] formed by 6 bidirectional switches which offer the possibility of operating in four-quadrant like the direct MC. The second stage is a conventional voltage inverter (figure 1 .4).

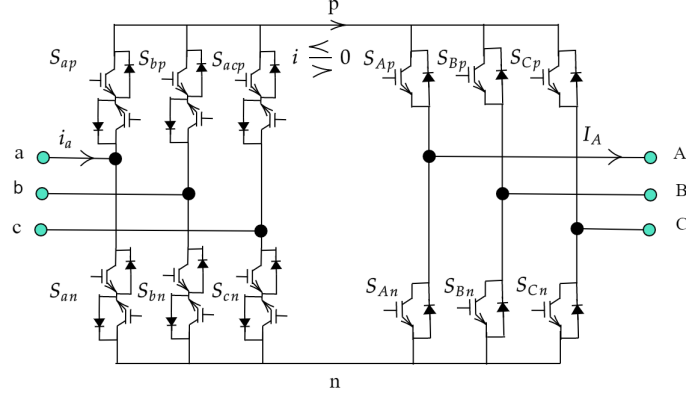


Figure 1.4: Indirect matrix converter.

At each instant, two transistors and two diodes conduct in the rectifier. This topology offers the same advantages and fewer disadvantages as the direct MC converter. Indeed, the switching, as well as the protection circuit (clamp circuit) also are simpler than those of the direct MC. Moreover, it consists of 18 IGBTs and 18 diodes like the MC [17, 20]. From a production point of view, it is simpler because of the inverter stage (there are modules that encapsulate the inverter and specific drivers for the inverter). It has another advantage which is the possibility of using it in the case of multi-machine by adding inverters in parallel like we did in this research.

A major step was accomplished in 2001 with the development of several types of IMC following the invention of sparse IMC by Kolar et al [7, 20] (figure 1.5). This converter is an IMC but with a reduction of the semiconductors that can be controlled at the level of the rectifier stage.

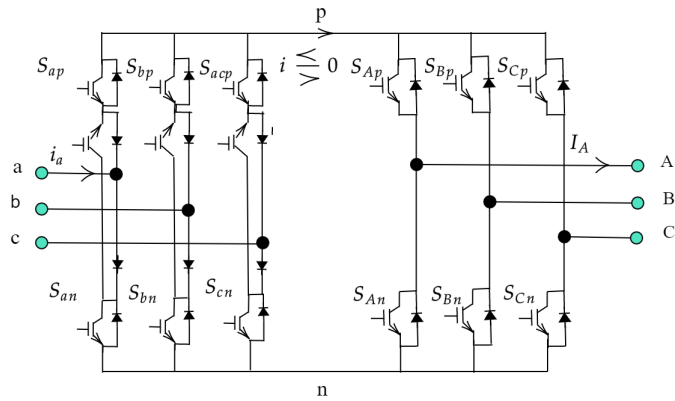


Figure 1.5: Sparse IMC.

This topology is generated by the pooling of a first switch that can be controlled from the cell **a** and a second from cell **á**, within the same arm. With this structure, a unity

power factor is achieved at the input, sinusoidal current, and an output voltage equivalent or very close to that obtained by the conventional IMC.

This structure makes it possible to eliminate three controllable components of the rectifier compared to the conventional indirect topology. On the other hand, depending on the direction of the power flow, three transistors and three diodes (power transmitted to the load) or two transistors and two diodes (power reinjected into the input network) are conducting. The conduction losses of this topology will therefore be greater than those generated by the rectifier of the "indirect matrix" solution.

Another topology named "Very-Sparse" matrix converter has been proposed and experimentally verified by Kolar [21,22] (figure 1.6). This structure is composed of six single transistor four-segment switches. This topology makes it possible to limit the number of controllable components for the rectifier to six, thus reducing their number by two compared to the rectifier of the "indirect" topology. The conduction losses of this topology are greater since each active element requires the conduction of a transistor as well as two diodes. Thus, the rectifier solicits, at each instant, two transistors and four diodes.

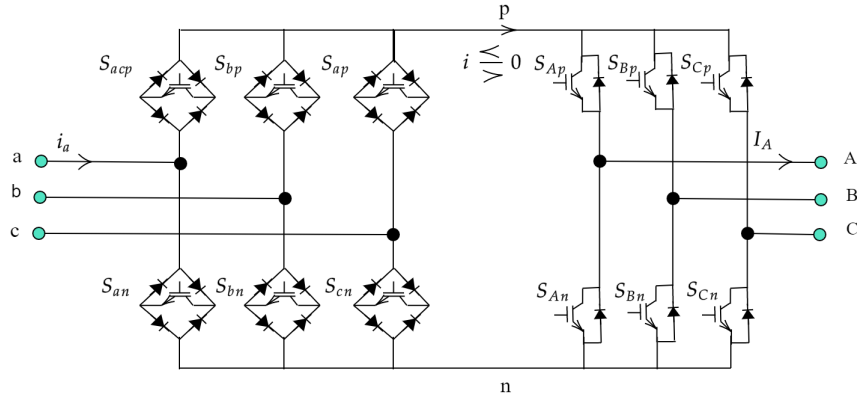


Figure 1.6: Very sparse IMC.

If the operation with DC bus current is in a single quadrant, the number of components to be controlled can be reduced to 12 (unidirectional very sparse MC) (figure 1.7) or to 9 (for the ultra sparse IMC) (figure 1.8). These latter topologies do not allow regeneration and braking [20–22].

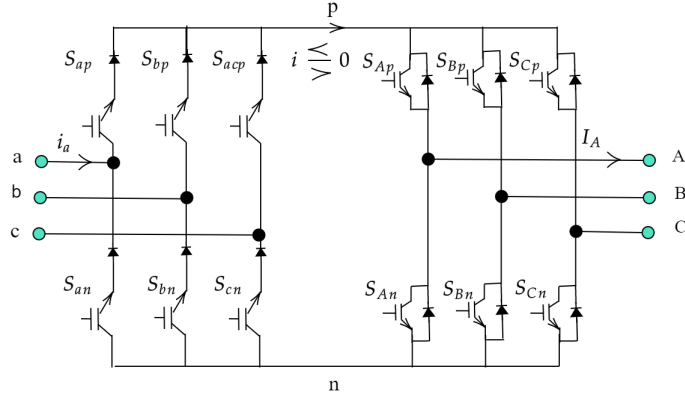


Figure 1.7: Unidirectional very sparse IMC.

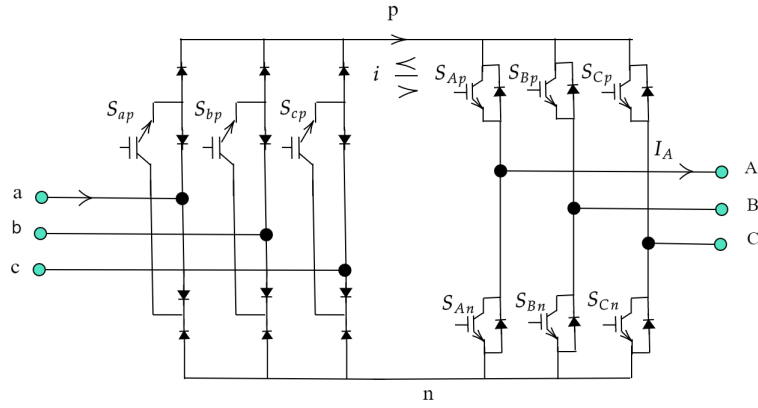


Figure 1.8: Ultra sparse IMC.

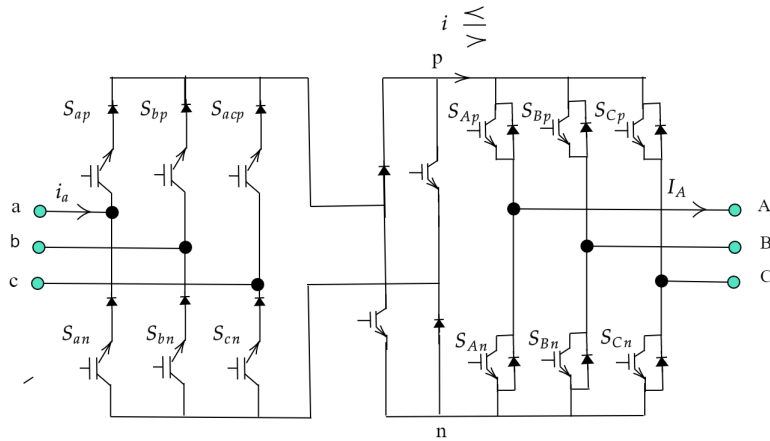


Figure 1.9: IMC with inverse stage.

Another topology of the so-called “Inverse Stage” matrix converter (figure 1.9) is composed of a rectifier with unidirectional current switches, in cascade with a second inverter

Table 1.1: Comparison between different MC topologies.

Topology	Max number of components in conduction	Bidirectional	Number of IGBTs	Number of diodes	Number of isolated power supplies for the drivers
DMC	6	yes	18	18	9(6 in the case: common collector structure switches)
IMC	7	yes	18	18	8(5 if a driver is used for a complete inverter)
Sparse IMC	9	yes	15	18	7(4 if a driver is used for a complete inverter)
Very Sparse IMC	9	yes	12	30	10(7 if a driver is used for a complete inverter)
Ultra-Sparse IMC	9	no	9	18	7(4 if a driver is used for a complete inverter)
IMC with inverter stage	9	yes	14	14	11(8 if a driver is used for a complete inverter)

stage. It has as many controllable components as the "Sparse" version. This rectifier recovers energy thanks to its inverse stage. Having to operate at the same modulation frequency as the rest of the structure, this topology generates additional switching losses as well as greater control complexity [23].

A comparative study between these structures has been summarized in Table 1.1.

A multi-level structure called H-bridge MC has been proposed, analyzed, and experimentally verified by Erikson in [24] (figure 1.10). This structure is similar to that of the direct MC but instead of using conventional bidirectional switches, a CCH-b (clamped capacitor H-bridge) is used.

This structure makes it possible to synthesize the input voltage as well as the output voltage unlike the MC where the voltage is synthesized on one side and the current on the other side. It makes it possible to exceed the limit of the voltage ratio therefore the voltage ($0 \leq V_s \leq V_{in}$). This structure aims to increase the conversion efficiency for low voltages and low powers [24] but it presents several capacities which complicate its control

and which do not allow the control of the power factor [25, 26]. This structure has been studied in detail and applied in wind power installations [24]. It is possible to increase the number of levels by cascading two H-bridges.

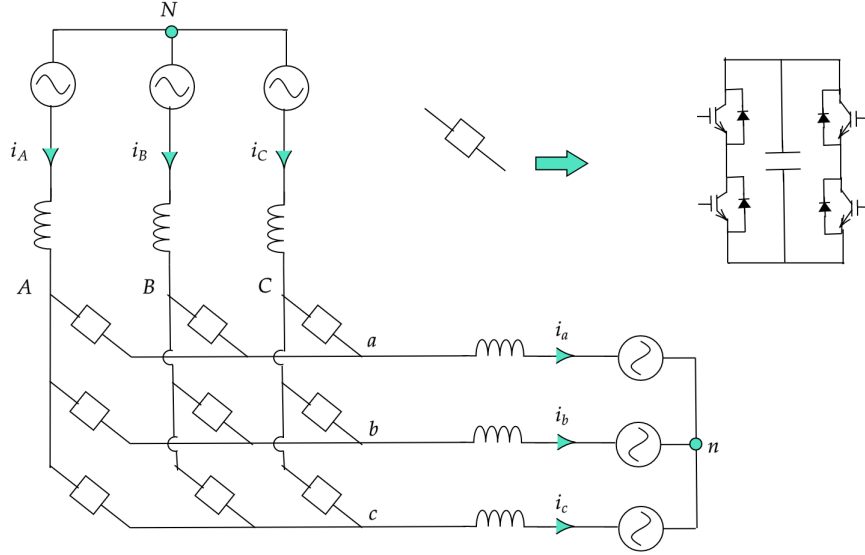


Figure 1.10: H-bridge multi-level matrix converter.

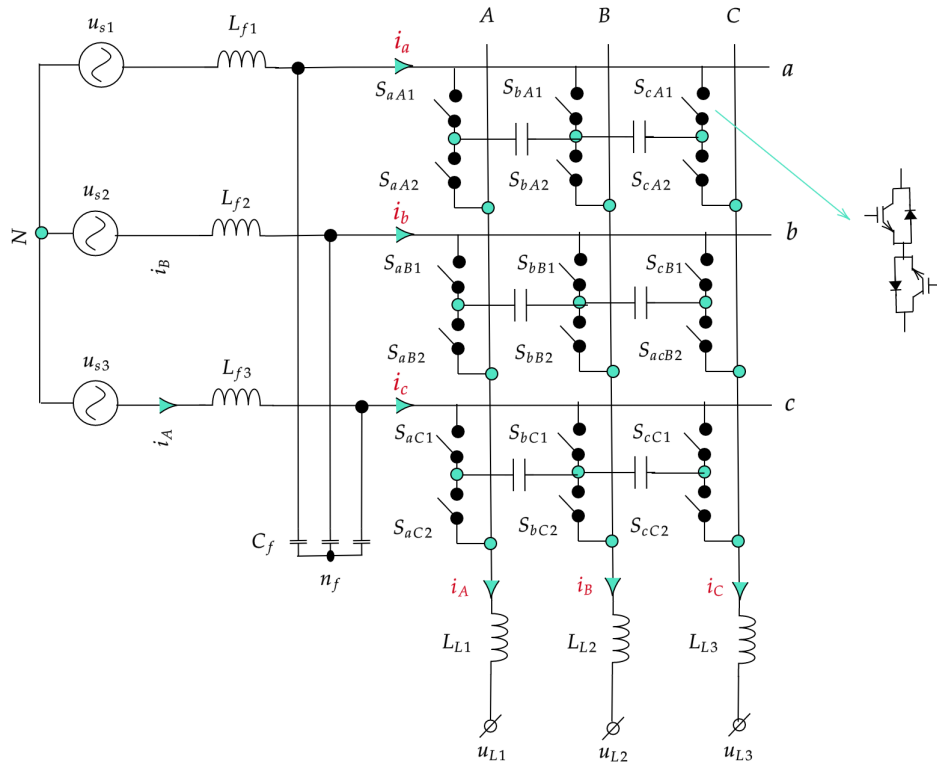


Figure 1.11: Multi-level MC with flying capacitors.

Another multi-level structure named multi-level MC with flying point capacitors has been proposed [27]. Its structure is shown in figure 1.11. This converter is obtained by replacing each switch in the forward MC with two or more series of connected switches and flying capacitors. It makes it possible to improve the output voltage but on the other hand, its control is very complex because of the voltages of the capacitors which must be controlled [3, 27].

The indirect three-level matrix structure (ITLMC) was proposed by Kolar et al [21, 22, 28]. It is similar to that of the conventional IMC by only changing the conventional two-level inverter by a three-level inverter [21, 23] (figure 1.12). This structure made it possible to have multilevel voltages and better performance than that of conventional IMC [21]. But its control is a little complex especially if we use the algorithm for balancing the voltages of the input capacitors.

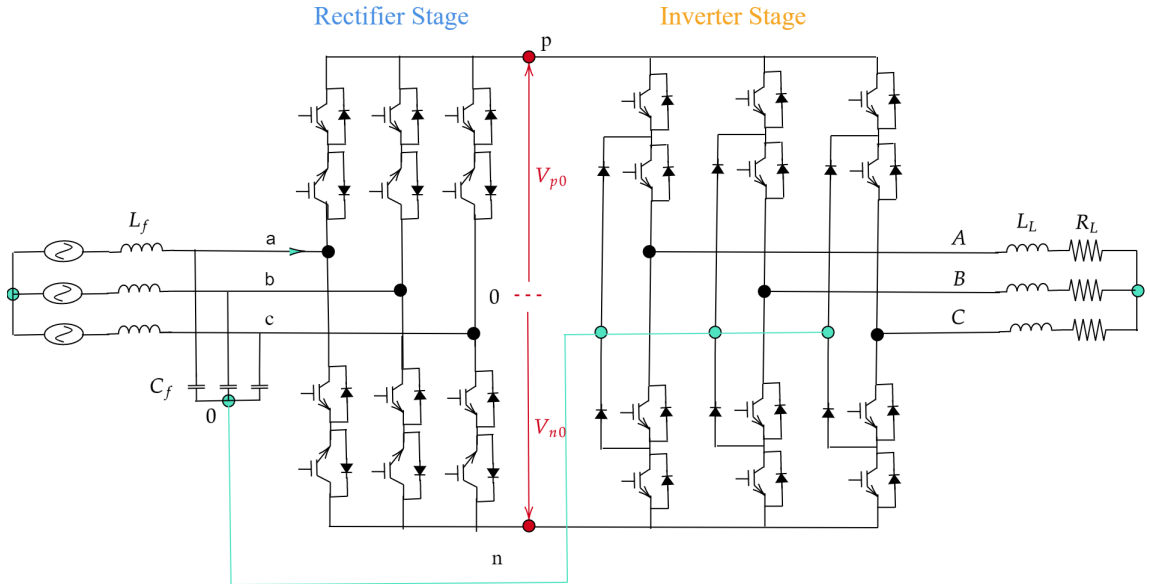


Figure 1.12: Indirect three levels MC.

In 2007 S. Ali Khajehoddin, et al. have proposed and studied a sparse multi-level IMC structure with fewer semiconductors than previous structures (figure 1.13). They used unidirectional switches in the rectifier stage and a three-level diode-clamp inverter for the inverter stage.

Another multi-level structure has been proposed by M. Y. Lee et al [21]. (figure 1.13) with fewer switches than the conventional I3LMC structure. SVM modulation has been proposed [29, 30]. This structure has been analyzed and tested experimentally [29] and it has been compared to other multilevel structures in [30]. It showed good output

performance especially for low ratios compared to the conventional IMC structure. It has the advantage of having fewer switches than the ITLMC and produces multilevel voltages, but from the point of view of the output signal quality, the ITLMC gives relatively better results. Among the disadvantages of this structure, are noted the distortions of the input current which are more important than the ones experienced in the conventional structure.

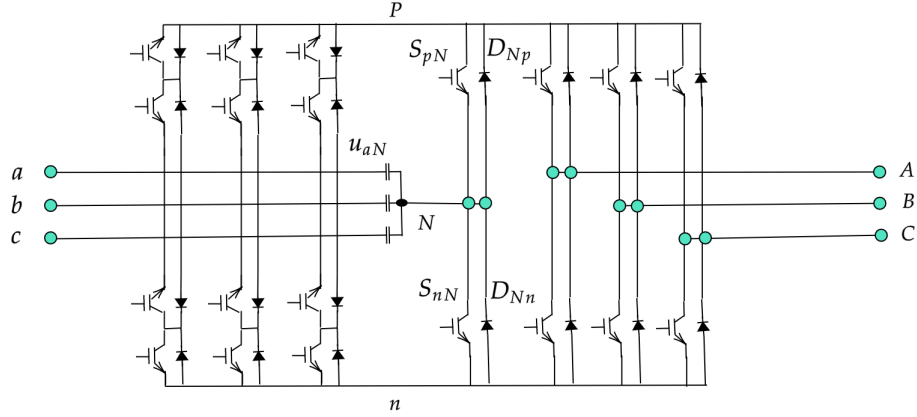


Figure 1.13: Indirect three-level sparse MC.

Another unidirectional multi-level input and output structure based on the Vienna rectifier has been proposed in [17] (figure 1.14). This structure is rarely presented in the literature. It has the advantage of having fewer switches than the ITLMC and also, the connection of the neutral point is not done at the level of the input capacitors.

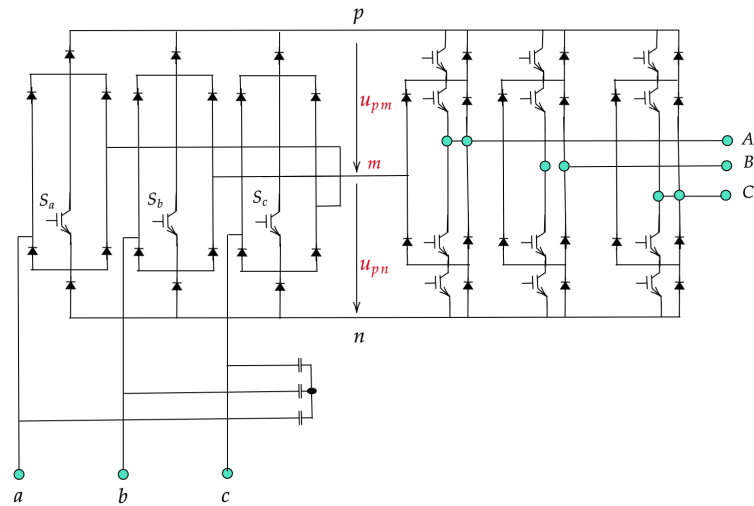


Figure 1.14: Three levels Vienna MC.

In 2001 different hybrid topologies were proposed to improve the performance of the

MC, therefore increasing the range of the output voltage and reducing the sensitivity to disturbances of the input quantities [31].

A hybrid converter (HMC) is the arrangement of two types of converters: the main converter, in our case it is the MC, and an auxiliary converter to circumvent the disadvantages of the main converter. There are two types of HMC, the direct hybrid MC [31] and the indirect hybrid MC studied by Klumpner [31]. Three structures of HMC are studied and discussed in [31]:

- In the first structure, an H-bridge is added to each output of the MC (figure 1.15). This structure can be applied to a wide range of motors and loads but it has a serious problem related to the large number of semiconductors and capacitors (large size and conduction losses).

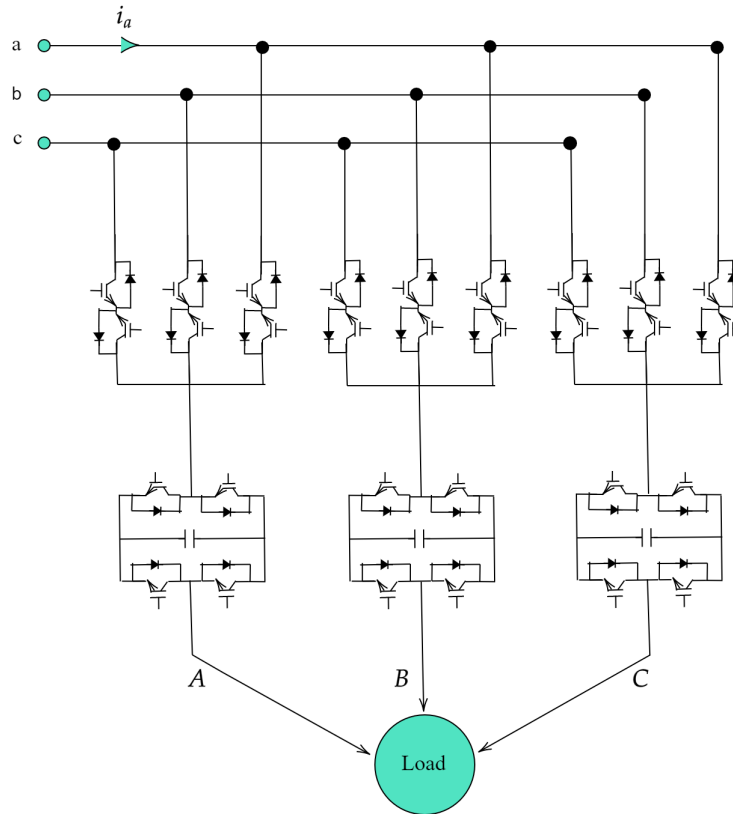


Figure 1.15: H1 topology.

- In the second structure, the outputs of the MC are connected to the inputs of load winding. The outputs of these windings are connected to a voltage inverter with a flying capacitor (figure 1.16). This structure is more efficient than the first, less expensive (fewer semiconductors and capacities) and it allows an appropriate control

to exceed the 0.866 limit of the voltage ratio (two types of control are proposed [20]). Its disadvantage is that it only applies to AC loads where the inputs and outputs of the windings are accessible (open winding).

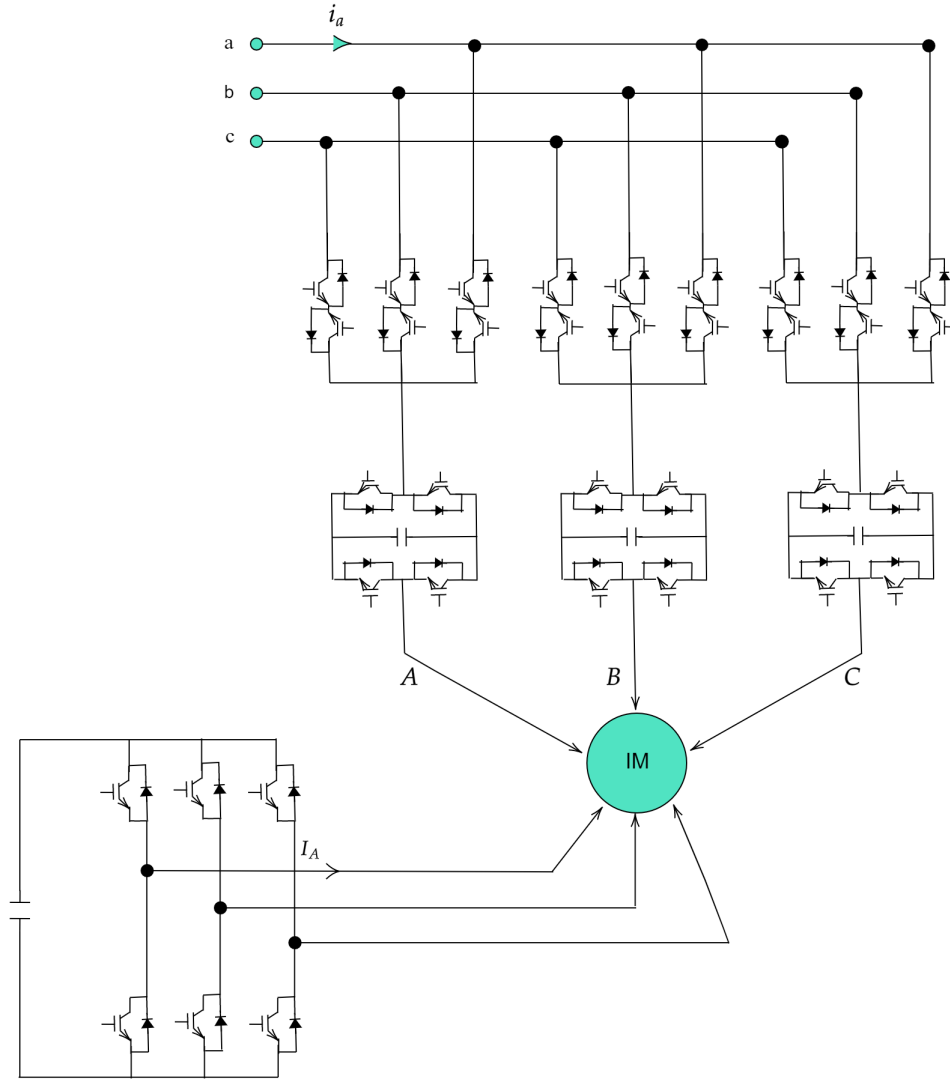


Figure 1.16: H2 topology.

- In the third structure which is based on the IMC, an H-bridge is added between the rectifier stage and the inverter stage (figure 1.17). This structure gives results close to the second structure and it also makes it possible to compensate for the imbalance effect of the input voltages [31]. Its disadvantage is that it must maintain a zero power balance (as for the second structure) which has as a consequence a limited supply voltage and which disturbs the power flow [26]. A solution to these issues has been proposed in [20, 31].

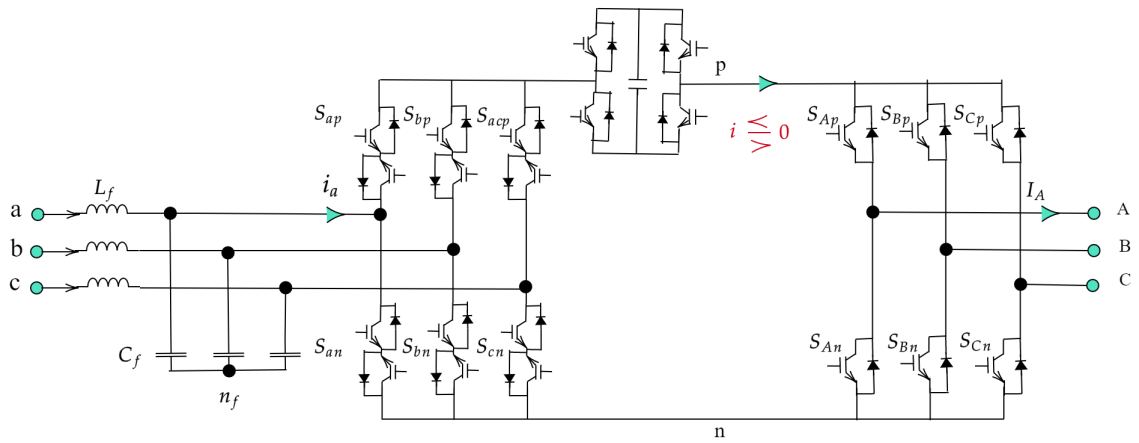


Figure 1.17: H3 topology.

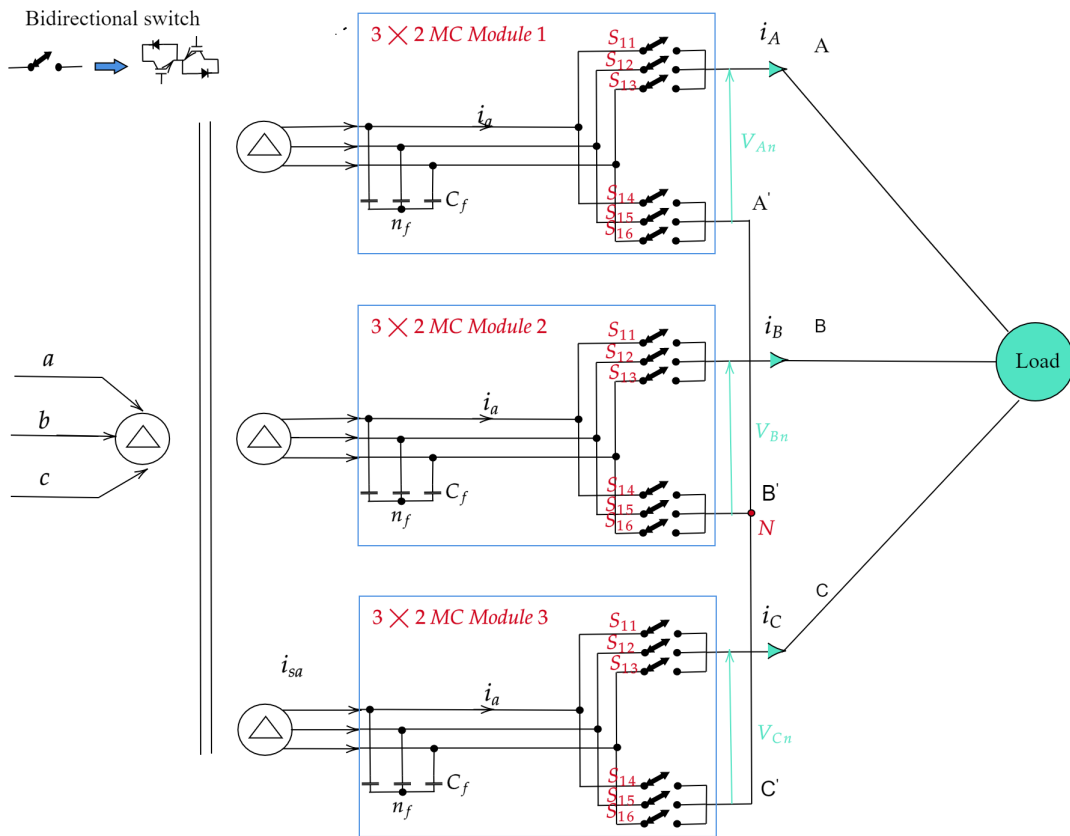


Figure 1.18: Simple MMC topology.

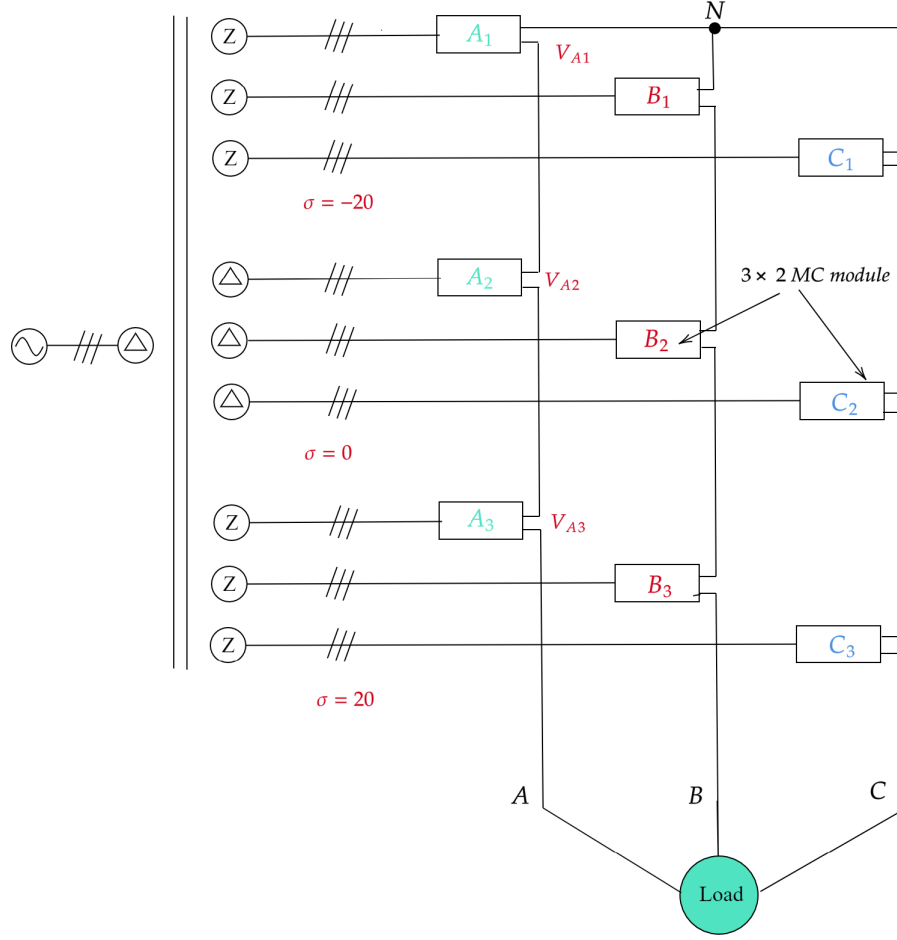


Figure 1.19: Three MMC topology.

Other multi-level structures called multi-modular have been proposed for high voltage and high power applications (electric traction). MMCs are similar to H-bridge-based converters. They have a modular design, good input and output signal quality, and are multi-level. Also, they can operate in four quadrants and can be used in regenerative applications [3, 32]. Between all the multilevel MC topologies mentioned above, the (MMCs) formed by cascades of MC modules (3×2) are the only types that have been commercialized. So far two prototypes have been developed in the industry, namely: a prototype of (3.3 kV / 3 MVA) with three modules in series per phase and another of (6.6 kV / 6 MVA) with six modules in series per phase [33]. Figures 1.18, 1.19 illustrate two examples. An indirect SVM control has been presented and experimentally tested for these two structures [34]. Other structures used for electric traction have been studied and tested experimentally in [35]. In [36] another topology of this type was proposed, it was combined with a medium frequency transformer, a prototype was simulated and then

tested experimentally.

In 2005 the Z-source matrix structures were initiated in [37], and a new topology was proposed by W. Song et al. [38]. This structure was developed to overcome the problem of limitation of the voltage ratio in the MC. It is more efficient than overmodulation control techniques with which the quality of the input current and the output voltages are degraded. It is currently at the research stage [2, 16]. In [16] a review of topologies and research on the Z-source matrix has been performed. There are two families of Z-source MC. The first family, direct, uses the MC. The simplest topologies. One of the topologies has an MC and a Z-source network at the input (figure. 1.20). The second has a quasi-Z-source network at the input of each phase (figures. 1.20 and 1.21). In [39] several topologies of DZSMC have been proposed and analyzed. A PWM control has been implemented and an experimental study of a simple topology has been carried out and verified. In [40] a PWAM (pulse width amplitude modulation) technique has been implemented and validated for a DZSMC topology.

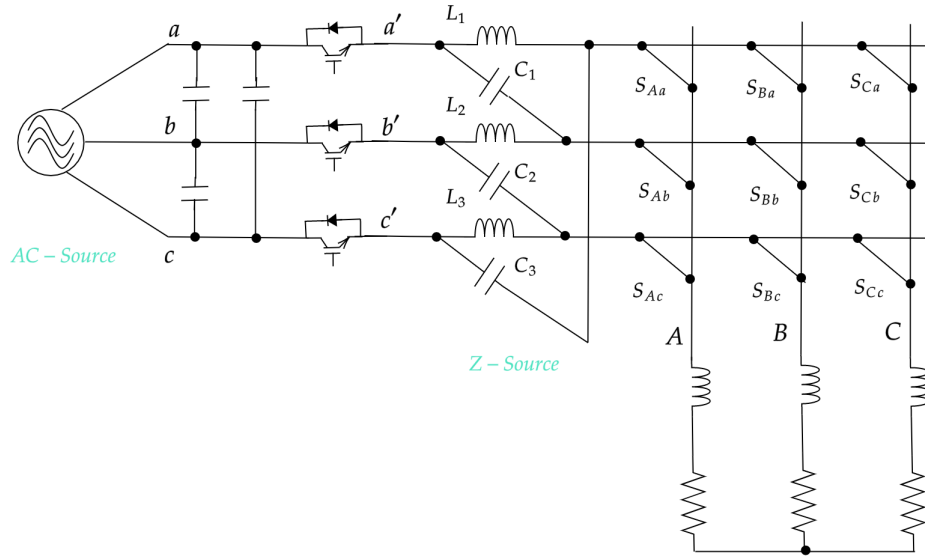


Figure 1.20: Topology with Z-source network.

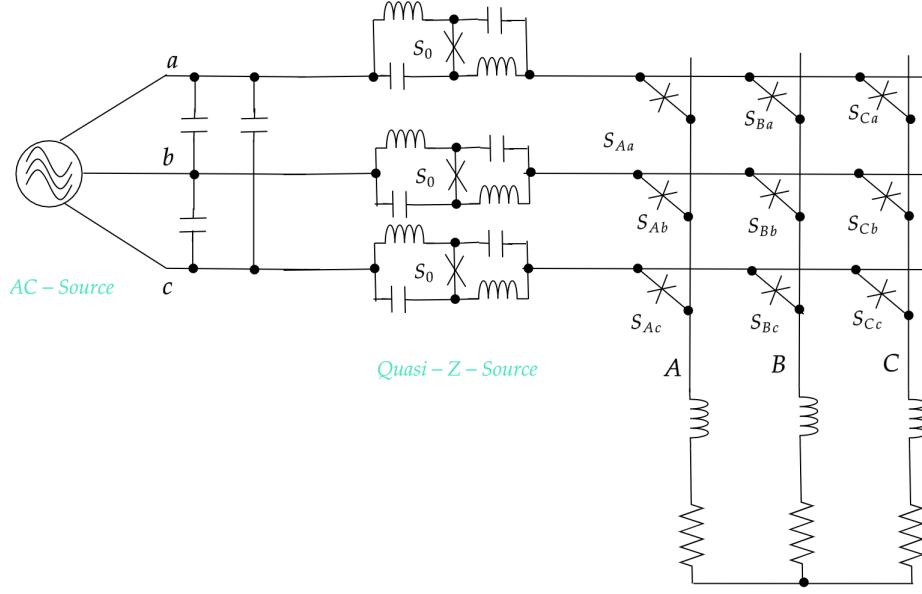


Figure 1.21: Topology with quasi Z-source..

The second family of Z-source MC is the ZSIMC (Z-source IMC) which is simpler than the first. It consists of an IMC with a small Z-source network between the two stages (figure. 1.22). The Z-source network LC parameters are similar to the input filter parameters [16, 41]. This circuit with the control loop allows amplifying of the DC bus voltage to ensure the voltage required by the load and to guarantee a unity power factor at the input. At the same time, this circuit is a harmonic filter of the input current. Modulation has been proposed and simulated in [42].

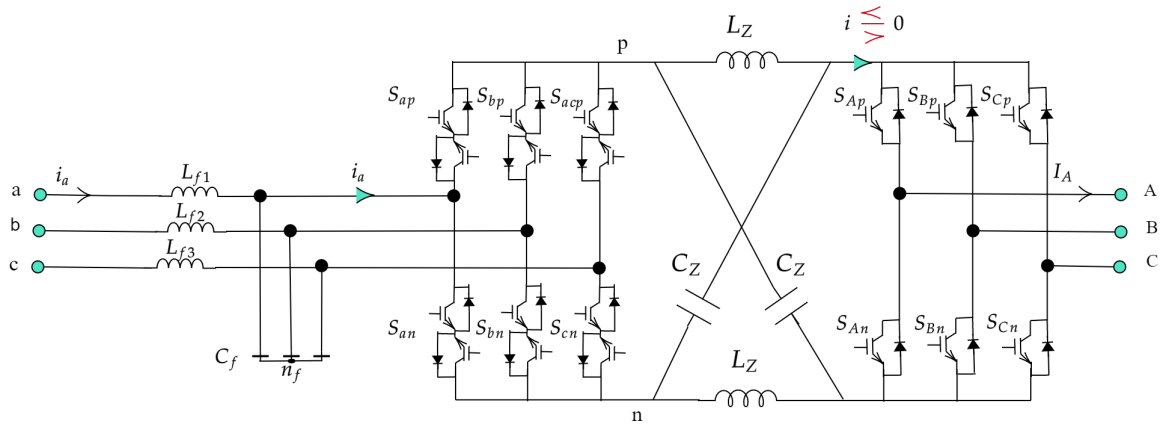


Figure 1.22: IMC topology with Z-source network.

A ZSIMC topology with a three-level NPC inverter stage was studied in [16] (figure. 1.23) where a carrier-based modulation was simulated. A comparison between two indirect

z-source topologies has been proposed in [43].

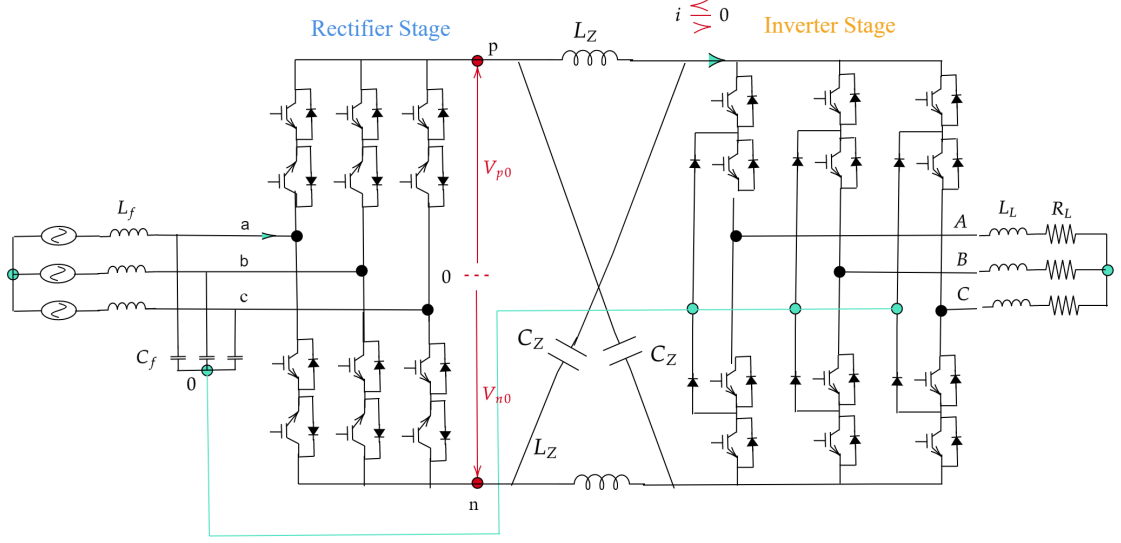


Figure 1.23: ITLMC topology with Z-source.

In [19] a new indirect type matrix topology has been proposed (figure 1.24). This topology aims to integrate two different AC sources (frequency and amplitude) into the network. With an appropriate control (in this case carrier-based), it makes it possible to amplify the output voltage while keeping a sinusoidal shape of the input and output quantities. Experimental investigations have been carried out. Other structures, proposed in the literature, are summarized in [2, 20].

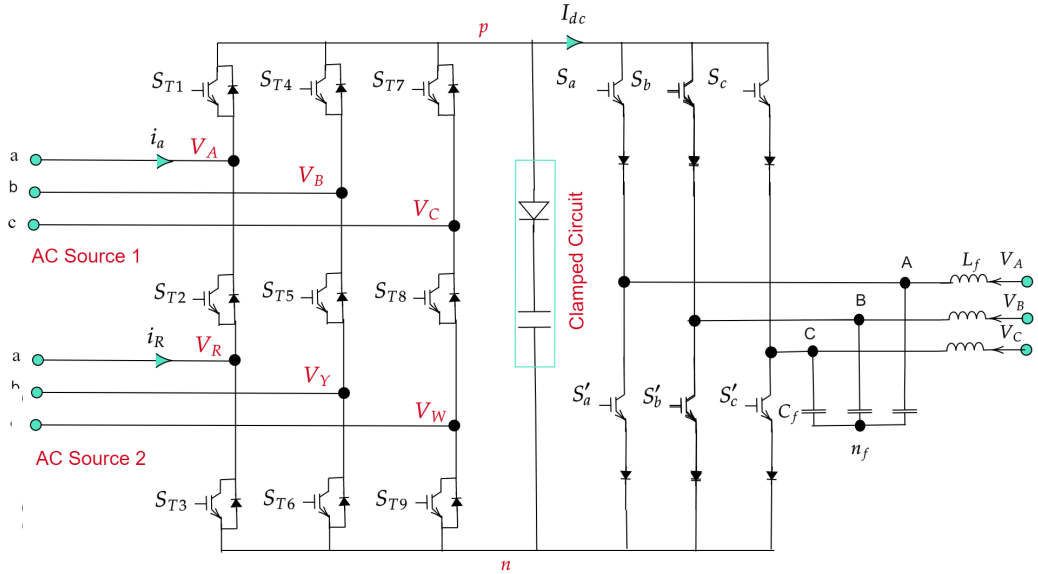


Figure 1.24: IMC topology with two inputs.

1.3 Modeling of Matrix Converter

To understand its principle, we refer to the most classic case which is the MC (3 x 3) (figure 1.25). The MC is able to connect each input phase to each output phase via nine four-segment switches with forced switching.

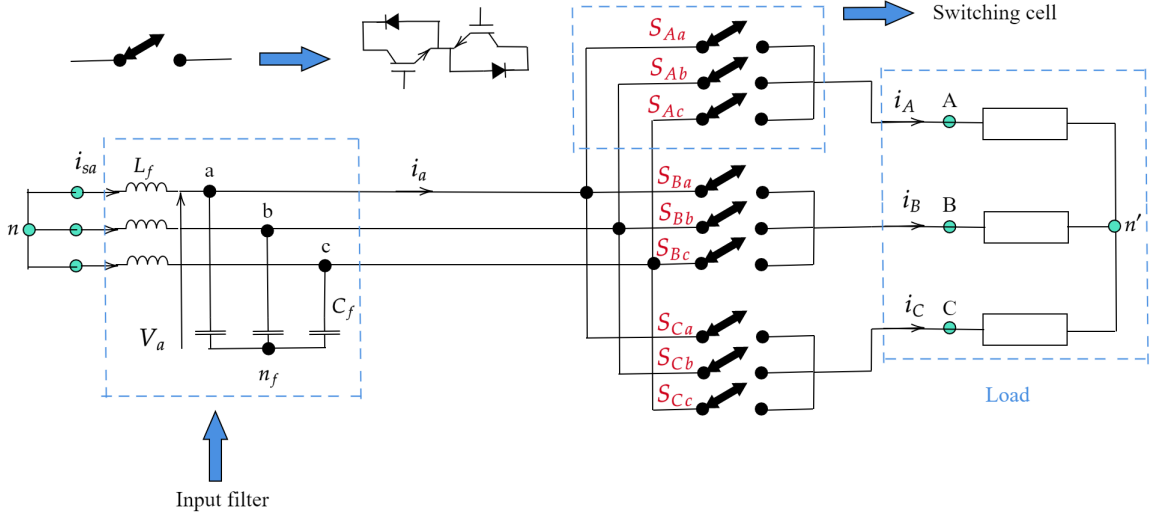


Figure 1.25: Conventional matrix converter.

For its operation, it is necessary to respect the principle of causality so as to have different sources on either side of the converter [4, 44, 45].

The output voltages are referenced to the neutral n' . A common-mode voltage between these two reference potentials named $\nu_{nn'}$, appears [23]. The selection of each input phase during a sequence forms the output signals of the converter for well-defined periods of time.

The output voltage is formed from the segments of the three input voltages. The input current is formed by the segments of the three output currents [17].

During the operation, it must be to avoid short-circuiting of a voltage source or open-circuiting a current source. Indeed, if two input phases are short-circuited, an infinite current passes through the switches and damages the circuit. This is due to the direct connection with the voltage source at the input. And given the inductive nature of the load and the absence of freewheeling, if a phase is open, an overvoltage due to the energy stored in the inductor damages the switches and the circuit of the converter [8, 9].

For a MC cell, only one switch must be on, so :

$$S_{ja} + S_{jb} + S_{jc} = 1, j \in \{A, B, C\} \quad (1.1)$$

Where S_{jk} represent the components of the matrix converter, the following statement defines the connection function.

$$S_{jk} = \begin{cases} 1 & \text{if the switch is closed} & j \in \{a, b, c\}, \\ 0 & \text{if the switch is open} & k \in \{A, B, C\} \end{cases} \quad (1.2)$$

So we have 27 possible combinations.

The instantaneous output and input voltages are linked by a connection matrix which represents the instantaneous state of all the switches such as:

$[T_1]$ is the matrix which connects the phase-to-phase voltages and $[T_2]$ is the matrix that connects the phase-to-neutral voltages with respect to the network neutral:

$$[T_1] = \begin{bmatrix} S_{Aa} - S_{Ba} & S_{Ab} - S_{Bb} & S_{Ac} - S_{Bc} \\ S_{Ba} - S_{Ca} & S_{Bb} - S_{Cb} & S_{Bc} - S_{Cc} \\ S_{Ca} - S_{Aa} & S_{Cb} - S_{Ab} & S_{Cc} - S_{Ac} \end{bmatrix} \quad [T_2] = \begin{bmatrix} S_{Aa} & S_{Ab} & S_{Ac} \\ S_{Ba} & S_{Bb} & S_{Bc} \\ S_{Ca} & S_{Cb} & S_{Cc} \end{bmatrix} \quad (1.3)$$

So we can write the instantaneous simple and phase-to-phase output voltages in the following form:

$$\begin{aligned} V_{sll} &= \begin{bmatrix} V_{AB} \\ V_{BC} \\ V_{CA} \end{bmatrix} = [T_1] \cdot \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \\ V_S &= \begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = [T_2] \cdot \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \end{aligned} \quad (1.4)$$

The input current is written as:

$$I_e = \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} = [T_2]^T \cdot \begin{bmatrix} I_A \\ I_B \\ I_C \end{bmatrix} \quad (1.5)$$

In the case of a three-phase load connected in delta we have:

$$I_e = \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} = [T_1]^T \cdot \begin{bmatrix} I_{AB} \\ I_{BC} \\ I_{CA} \end{bmatrix} \quad (1.6)$$

V_A, V_B, V_C are the phase-to-neutral output voltages with respect to neutral n (neutral of the input) and V_{AB}, V_{BC}, V_{CA} are the voltages between output phases, I_A, I_B, I_C are the

simple output currents, V_a, V_b, V_c are the phase-to-neutral input voltages with respect to neutral n , I_a, I_b, I_c are the input currents.

To generate a set of balanced sinusoidal voltages, the switching frequency must be very high compared to the frequency of the input voltages [46, 47].

To determine the behavior of the matrix converter during the switching period (T_s), the connection generating function is defined for each switch, which is only the average value over a switching period (T_s) of the instantaneous connection function. The period T_s is assumed to be infinitely small. They can also be considered as duty cycles for each switch:

$$m_{ij} = \frac{1}{T_s} \int_0^{T_s} S_{ij}(\tau) \cdot d\tau = \frac{t_{ij}}{T_s} \text{ avec } 0 < m_{ij} < 1 \quad (1.7)$$

where

$$m_{ja} + m_{jb} + m_{jc} = 1, j \in \{A, B, C\} \quad (1.8)$$

where t_{ij} represents the duration where the switch ij is in the *on* state during the period T_s [4, 21, 23].

One defines for the quantities between phases: $\hat{V}_{su} = [M_1] \cdot \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix}$, for the input currents

$\hat{I}_{in} = [M_{ii}]^T \cdot \begin{bmatrix} I_{AB} \\ I_{BC} \\ I_{CA} \end{bmatrix}$ and for the line phases: $\hat{V}_S = [M_2] \cdot \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix}$, and for the input

currents: $\hat{I}_{in} = [M_{ph}]^T \cdot \begin{bmatrix} I_A \\ I_B \\ I_C \end{bmatrix}$

where

$$[M_1] = \begin{bmatrix} m_{Aa} - m_{Ba} & m_{Ab} - m_{Bb} & m_{Ac} - m_{Bc} \\ m_{Ba} - m_{Ca} & m_{Bb} - m_{Cb} & m_{Bc} - m_{Cc} \\ m_{Ca} - m_{Aa} & m_{Cb} - m_{Ab} & m_{Cc} - m_{Ac} \end{bmatrix}, \text{ and } [M_2] = \begin{bmatrix} m_{Aa} & m_{Ab} & m_{Ac} \\ m_{Ba} & m_{Bb} & m_{Bc} \\ m_{Ca} & m_{Cb} & m_{Cc} \end{bmatrix}$$

These continuous functions of time can then be used to define and compare the modulation strategies and to compare the output quantities with the reference quantities. In general, with good modulations, the voltages satisfy the relation $V_S = V_{ref}$ in the linear range of the modulation [46, 47].

1.3.1 Indirect matrix converter model

With this model, the matrix converter is described as a combination of a four-quadrant current rectifier with a voltage inverter connected by a virtual DC link as shown in figure 1.26.

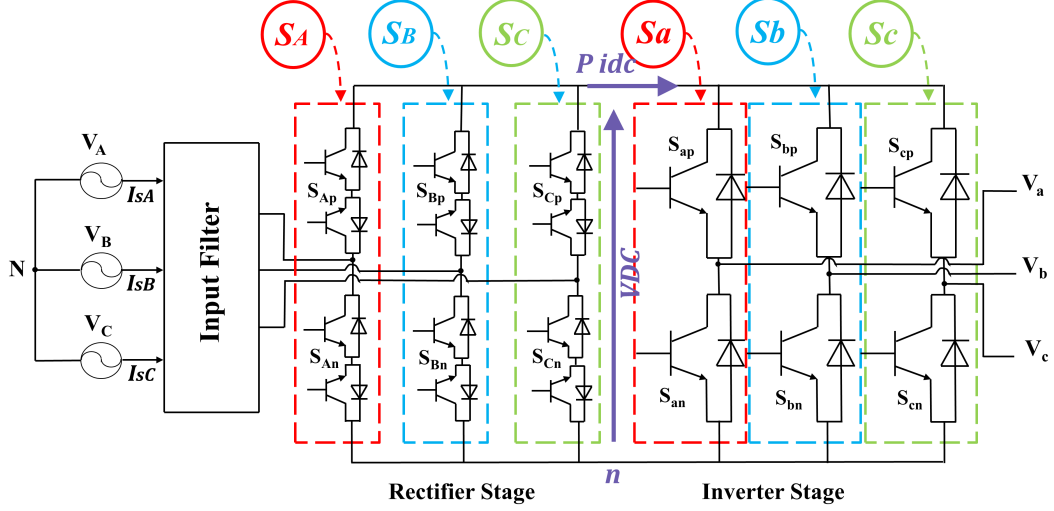


Figure 1.26: Indirect Matrix Converter.

The two power stages are directly connected by their DC sides and are bidirectional in power. The indirect model of the matrix converter makes it possible to simplify control. In addition, it makes it possible to adapt the control techniques - dedicated to the inverter to the MC. Based on the indirect matrix converter model, the conversion matrix is defined as the product of two conversion matrices on each side (inverter and rectifier).

$$\begin{aligned}
 \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} &= \begin{bmatrix} S_{ap} & S_{an} \\ S_{bp} & S_{bn} \\ S_{cp} & S_{cn} \end{bmatrix} \begin{bmatrix} S_{Ap} & S_{Bp} & S_{Cp} \\ S_{An} & S_{Bn} & S_{Cn} \end{bmatrix} \begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} \\
 &= S_{inv} \times S_{rec} \times V_{in}
 \end{aligned} \tag{1.9}$$

Where S_{jk} represent the connection function defined by the relation:

$$S_{jk} = \begin{cases} 1 & \text{if the switch is closed} & j \in \{a, b, c\}, \\ 0 & \text{if the switch is open} & k \in \{A, B, C\} \end{cases} \tag{1.10}$$

This model will be used for the control of the indirect matrix converter and even for the three-level indirect matrix converter [11, 23].

1.4 Commutation Study

The switching of four-segment switches is more restrictive than that of the switches used in voltage inverters [23, 48]. Indeed, no path for the output current is naturally available in the switches necessary for the operation of the matrix converter, and this is due to the presence of a transistor in series with a diode. As at each instant, only one switch is closed in a switching cell (figure. 1.27), the switching takes place between two switches within the cell [17].

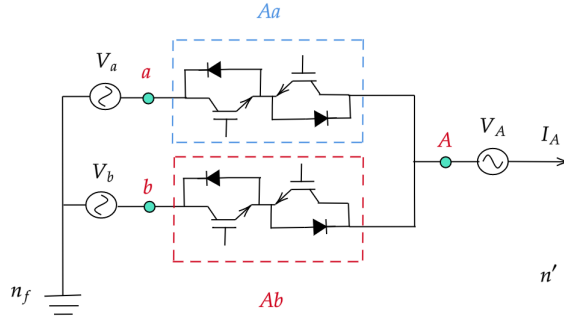


Figure 1.27: Basic diagram for the study of commutations.

The opening and closing of a switch are not instantaneous and present a certain delay on the command (figure 1.28). In the event of a strictly complementary control, there may be intervals during which the two switches are open or closed simultaneously and since the MC does not allow freewheel operation, there is a risk of short-circuiting the input voltage sources or opening the output current sources during switching [21, 23]. If the command introduces dead times, the short-circuit of the input voltage sources is avoided but the openings of the output current sources are not. This solution can be used in the case of low powers and by associating RC or RCD type switching assistance circuits in parallel with each switch to avoid overvoltages and create a channel for currents. This solution is very practical but it can lead to considerable losses, especially with the increase in dead time. The latter must be greater than the time of the risky zone [46].

Other methods are proposed, they dissociate the control of the (IGBTs) from the same four-segment switch. This dissociation makes it possible to generate particular sequencing of the control of the switching of the switches. This sequencing is controlled either by the sign of the switching cell output current or by the phase-to-phase voltage across the terminals of the two switching switches [49].

Two methods of commutation current are proposed in the literature: one uses four steps and the other uses three steps. If the sign of the current is sure, the latter can

be summarized in two steps. This type of switching requires the exact knowledge of the sign of the output current of the MC cell. Therefore it is difficult to use this sequencing controlled by the current when its amplitude is low [17, 49].

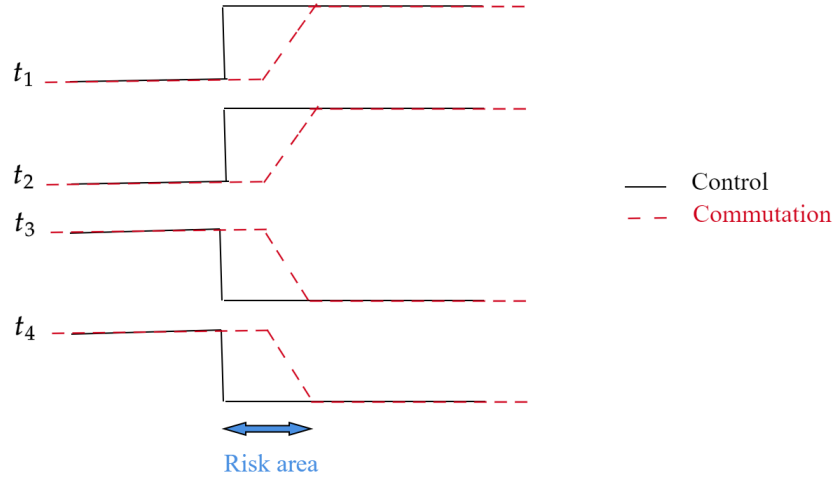


Figure 1.28: Schematization of risk during switching.

For the voltage also two types of switching are proposed, one with four steps, the other with two steps. This type of switching also requires exact knowledge of the sign of the phase-to-phase voltage across the terminals of the two switches to be switched. When the phase-to-phase voltage amplitude is low, this sequencing becomes difficult [23, 44]. There is another type of switching that combines the two parameters, voltage and current [23, 47, 50].

In the case of IMC, the switching is simpler than the MC case [2]. Indeed, the inverter stage is switched with the introduction of a dead time (no problem of overvoltage at the level of the inverter because of the antiparallel diodes). Commutation of the rectifier stage is done either in voltage or in current as for the MC [21, 48].

Current or voltage switching requires the return to call upon a logic device like CPLDs to implement the sequencing which is very fast and cannot be done by a microcontroller [2, 21].

1.5 Input Filter Design and Protection Circuit of MC

1.5.1 Input Filter

Matrix converters produce output voltages as well as input currents with high-frequency harmonics. These harmonics are likely to generate unwanted high-frequency disturbances that can affect other electronic systems. Consequently, it is necessary to add one or more filters to the converter. An input filter makes it possible to respect the limits of the current distortions taken from the network as well as the electromagnetic interference generated by the switching [44, 49].

An L-C type low-pass filter, shown in figure 1.18 for a single-phase, is added to the input of the converter [51, 52]. The resistance R_f represents the internal resistance of the inductor L_f .

The use of more complex topology such as multi-stage filters has been recommended [2, 51] in order to achieve a higher reduction of the switching frequency. But these filters cannot be industrialized given the additional cost they would incur [23].

When sizing the filter of MC (figure 1.29), five parameters must be taken into account [4, 52]:

- The quality of energy imposed;
- The harmonic content generated by the MC;
- The switching frequency of the converter;
- The modulation technique used;
- The resonance frequency of the filter.

The filter must have a cutoff frequency lower than the switching frequency. It must also absorb little reactive power, have a reduced volume and weight, and have a low voltage drop at the level of the inductor [51, 52]. If we neglect the damping resistance R_d , this filter has a transfer function (FT) as defined in the following equation:

$$TF = \frac{Ii(s)}{Ij(s)} = \frac{\frac{1}{L_f C_f}}{s^2 + \frac{r_f}{L_f} s + \frac{1}{L_f C_f}} \quad (1.11)$$

where s is the Laplace operator.

This L-C filter may cause disturbances during the power converter switch-on procedure. The inrush of current when connecting the circuit can cause surges immediately after

the L-C filter and destroy the converter. In [49] Schuster added a resistor R_d in parallel with the inductor to dampen these unwanted disturbances. This resistance also aims to attenuate the oscillations around the resonant frequency and improve the stability of the system [52]. The value of this resistance is chosen in such a way that the losses generated are as low as possible [23]. According to [21] we must ensure that $R_d \leq 2\pi f_{CE} L_f$.

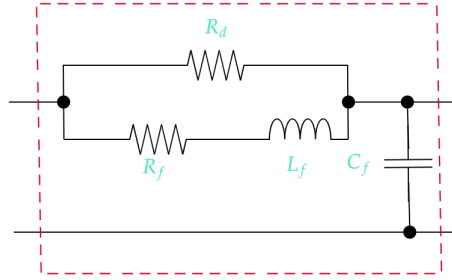


Figure 1.29: Diagram of the input filter for one phase.

1.5.2 Protection Circuit

Overcurrent protection

Protection of the switches against overload and/or short circuits must be added. This protection is done by external devices such as circuit breakers or fuses and/or by devices integrated with the driver circuit to detect switching errors that can cause a short circuit [21, 23]. In addition to these devices, an overcurrent detection loop can be added to the MC control program. This loop can stop the MC if an overcurrent is detected.

Overvoltage protection

In the case where the load is an inductive circuit like machines, overvoltages can be caused when one or more transit paths of the output currents are interrupted. The energy stored in the open load increases the voltage level at the terminals of the switches of the concerned cells. To prevent destructive surges from appearing on these switches, a path must be provided for the motor currents.

- a) **Protection with the “Clamp” circuit**

The first protection system used [2, 4] is a “Clamp” circuit [53]. It is composed of one or two capacitors (depending on voltage level) connected to all input and output potentials

through two diode-bridges as shown in figure 1.30. The rectifier bridge connected to the input is used to charge the capacitor C of this protection circuit.

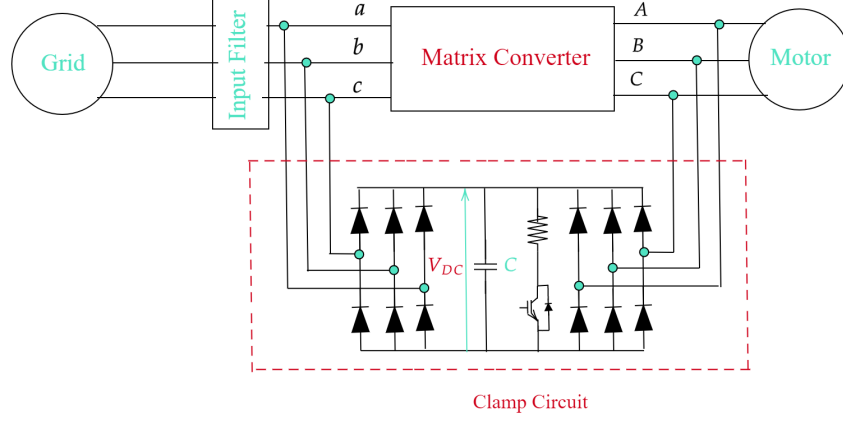


Figure 1.30: Clamp circuit for DMC.

VDC is equal to the maximum phase to phase voltage of the input network. During an overvoltage fault, the switches terminal voltages of the concerned cell increase. This will increase the output voltages of the matrix converter. At the moment when one of the output phase-to-phase voltages becomes greater than the VDC voltage, the diode-bridge connected at the output starts to conduct and limits the overvoltage to VDC. The energy stored in the motor phases is transferred into the capacitor and the voltage across the capacitor rises. A system consisting of a resistor in series with a transistor must be introduced in order to dissipate the stored energy. The clamp circuit sizing method is presented in [23, 54].

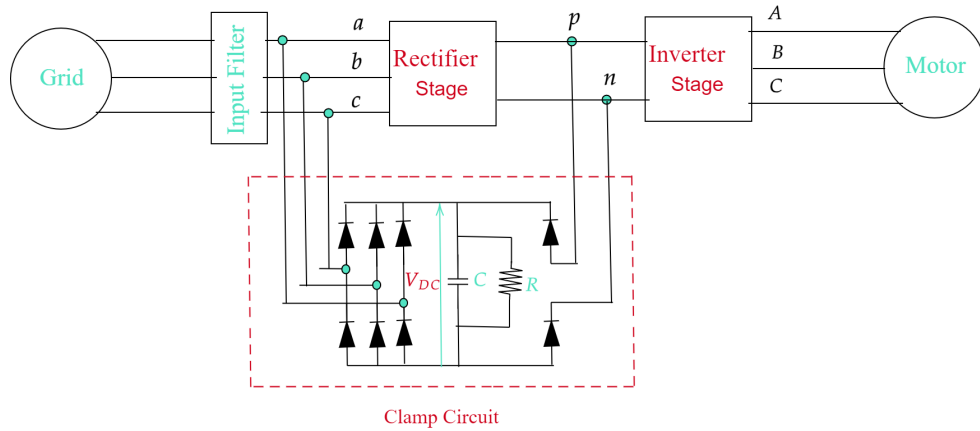


Figure 1.31: Clamp circuit for IMC.

This protection system has the advantage of being very simple and safe in all conditions of use. On the other hand, the number of semiconductor elements required increases (12 fast recovery diodes plus a transistor for the discharge).

For the indirect matrix converter topology, this circuit has fewer components [21, 46]. It is shown in figure 1.31.

- **b) Varistor protection**

A second solution [49] uses a passive protection system for low power applications. This protection is based on the use of varistors which are highly non-linear electrical resistors. Its resistance depends on the voltage at these terminals.

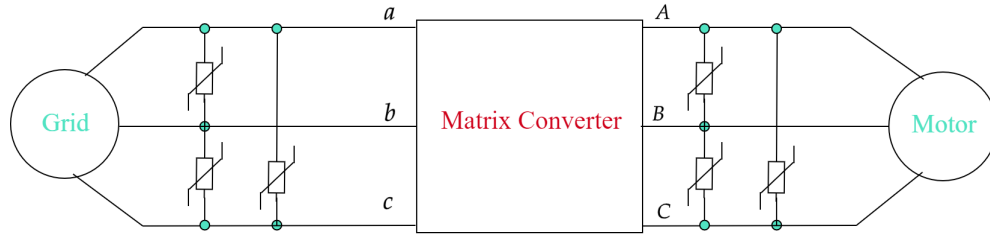


Figure 1.32: Clamp circuit for IMC.

Beyond a certain voltage threshold, the resistance of the varistor drops to allow the evacuation of the current and thus limiting the overvoltage. In normal operation, the losses caused by the varistors are negligible (very high resistance) [23]. This solution is very attractive, especially in the case of low powers. The varistors are connected at the input and at the output as shown in figure 1.32. Note that, if the overvoltage is significant, the use of varistors can cause their destruction, which generates a small explosion.

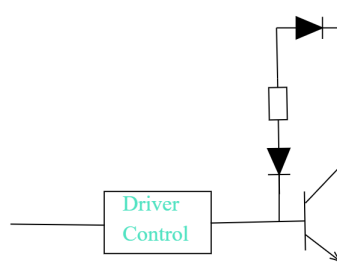


Figure 1.33: Active clamp circuit.

The energy from the load in the case of overvoltages can be dissipated in the converter itself by switching the IGBTs to linear operation. This can be achieved by a simple circuit called active clamping (figure. 1.33). It consists of a Zener diode, a diode, and a resistor placed in series between the collector and the base of each IGBT [23,49].

This solution is not yet widespread because few reliability studies have been carried out on this protection system [23]. In [51] a new protection circuit has been tested. This circuit includes a start-up circuit to ensure correct power sequencing and avoid short circuits during start-up. A resistor has been added to the clamp circuit to limit the current flowing through the capacitor.

1.6 Choice of Bidirectional Switches

The matrix converter requires the use of nine four-segment switches. These switches can be controlled both on initiation and on blocking. They must have the ability to block a voltage and conduct a current regardless of its sign. Due to their very complex implementation, only small productions of monolithic four-segment switches are possible today [2, 20]. These four-segment switches are therefore created by the association of several discrete elements, unidirectional in voltage and current [23,46].

There are four main arrangements for two-way switches. They are shown in figure 1.34.

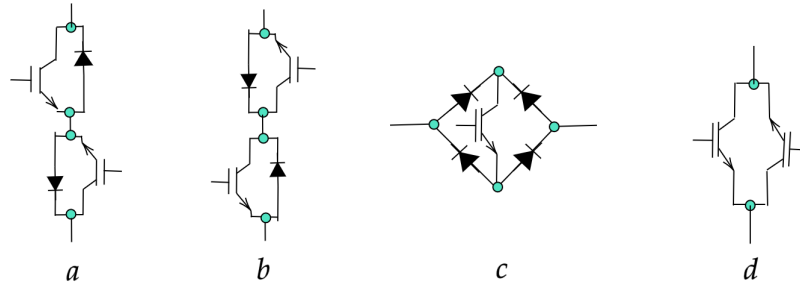


Figure 1.34: Different bidirectional switch topology (a) common emitter structure, (b) common collector structure, (c) transistor connected to diode bridge (d) RB-IGBT.

The first proposed topology is the simplest possible association. It consists of an IGBT connected to the center of a diode bridge (figure. 1.34.c).

Its main advantage is that it has only one controllable component. This reduces the cost of the power circuit as well as the complexity of the control and the driver circuit. Its main problem is related to conduction losses which are higher compared to other topologies (three switches conduct) [4,20,21].

Table 1.2: Comparative study between different structures of bidirectional switches.

Association of semiconductors	Simplicity of control	Voltage drop in conduction	Isolated power supply required for the matrix converter [3x3]
IGBT and diode bridge	simple	high	9
IGBT and diode in antiparallel common emitters	more complex	medium	9
IGBT and diode in antiparallel common collectors	more complex	medium	6
RB-IGBT	more complex	low	6

The two most commonly used configurations are the common-emitter topology and the common-collector topology. These two structures have two controllable switches and two antiparallel diodes. The diodes remain essential, they provide the reverse voltage blocking capacity.

One of the advantages of these two structures is to offer an additional degree of freedom for the management of the switching of the switch. Conduction losses are also reduced compared to the first topology. The difference between them is the number of isolated power supplies for the MC. If the topology with a common emitter is used (figure. 1.3.a), 9 isolated power supplies are necessary for the driver circuit of the MC. If the topology with a common collector is used (figure. 1.34.b) six isolated power supplies are sufficient due to the setting at the same potential of certain emitters of the IGBTs belonging to different switches [4, 17, 21].

Some companies, such as STMicroelectronics, Fuji electric, and IXYS are beginning to offer integrated bidirectional voltage switches (behavior similar to an IGBT in series with a diode (figure. 1.34.d), named RB-IGBT (Reverse Blocking IGBT) [1, 3]. The typical voltage drop of this type is then of the order of 1.6 to 2.2 V. So the advantage of this topology compared to other topologies is the reduction in the number of components and conduction losses. This topology is widely used in industry [4, 33]. A comparison between these structures is summarized in Table 1.2.

The preferred transistor is the IGBT which combines the advantages of two types of transistors which are the bipolar transistor and the MOSFET [48]. It is voltage controlled and has low voltage drop. It is widely used in high and especially medium power and medium frequency applications [23, 47].

1.7 Industrial Implementation

For the industrial implementation of the matrix converter and for this application to be more compact, it is preferable to use switch arrangements in the form of integrated semiconductor modules as in the case of inverters (the diode or thyristor bridge) [4].

Bidirectional switch modules are developed in the industry [24, 25]. In general, these modules use the RB-IGBT structure [4, 17, 47] and include three bidirectional switches. For example, we can cite the EMP-M50P12 module built by IRCI (International Rectifier Corporation Italy) which contains an MC cell and which can conduct a current of 50A and support a voltage of 1200V. We also have the Dynex DIM600EZM17-E000 which can drive a current of 600A and block a voltage of 1700V [4] (figure. 1.35). We also cite the module proposed by Fuji Electric in 2011. The latter is based on RB-IGBTs and comprises three cells of the MC (3x3)(figure. 1.36).

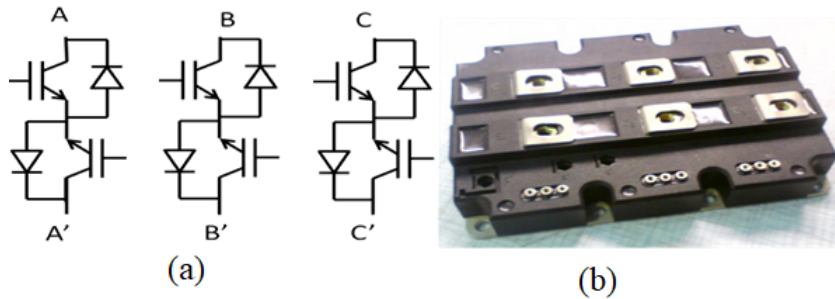


Figure 1.35: Dynex DIM600EZM17-E000 module (600A-1700V): (a) internal diagram, (b) external module.

Other modules used in industry are reported in [2, 4, 33]. Regarding the industrial realization of the MC, there are few prototypes currently on the market [2, 4]. Industrial prototypes were marketed in 2005. We can cite Varispeed AC from the Japanese company Yaskawa based on RB-IGBT 1200V. The latter was applied with a vertical axis wind turbine [4].

We also mention Frenic-FX from the Japanese company Fuji which has the same characteristics as the first. We also mention Yaskawa's FSDrive-MX1S intended for high

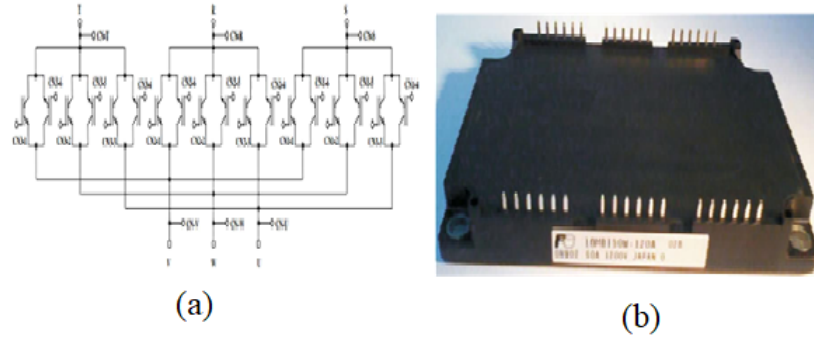


Figure 1.36: Module FUJI Electric MC (3x3) (1200V, 50A): (a) internal diagram, (b) external module.

power. It has been proposed for a power of 2.5 MW and 5 MW for a voltage of 3.3 kV and 6.6 kV respectively. It is based on a multi-modular structure [2, 4, 33]. The AC7 (5 kW–93.2 kW) from Yaskawa (figure 1.37(a)) is the first variable speed drive based on the direct structure of the low voltage (230V / 460V) matrix converter (3x3). It can operate either in constant V/f or in vector control with/or without sensors.

The U1000 (developed for powers of 3.7 kW – 600kW (figure 1.37(b))) is the latest solution from the Japanese company Yaskawa (2016) based on the matrix converter (the exact topology has not been specified). It is a compact all-in-one solution that allows operation at ultra-low THD ($< 5\%$) over the entire operating range and also allows regeneration and near unity power factor operation. This solution aims at a very good quality of energy and at the same time saving energy and space. Additionally, the U1000 offers great flexibility and motor control performance to meet a wide variety of application requirements. It can operate either in constant V/f or in vector control with or without a sensor. The company was able to win the 2016 Engineers Choice Award in **The Motion Control —Drives, Servo Drives branch** [17, 47].



Figure 1.37: Industrial prototypes of the MC, (a) Yaskawa AC7 (b) Yaskawa U1000 (2016).

1.8 State of the Art on Modulations of MC

The modulation generally makes it possible to synthesize the references of the output voltages and input currents from the input voltages and output currents. Reviews of the main modulations developed for the matrix converter have been made in [2], [4], and [9]. Since its introduction in 1976, different modulation strategies have been proposed and they can be broken down into two types:

- Scalar modulation.
- Vector modulation.

These methods consist in first finding a conversion matrix $[M_2]$, corresponding to the average model, and the image of connection times relating to a given period for each switch [9, 23]. Then these duty cycles are transformed into nine connection functions so as to be able to control the switches while respecting the constraints linked to the operation of the switching cells.

The first modulation strategy was proposed by A. Alesina and M. Venturini in 1980 [13, 45]. It solves the system defined by equations (1.4) and (1.5), with output reference voltages without the zero sequence term. They determined the connection generating function which has this expression:

$$m_{ij} = \frac{1}{3} \times \left(1 + \frac{2V_j \cdot V_k}{Ve^2} \right) \quad (1.12)$$

where j is input phase numbered ($a = 1, b = 2, c = 3$) and k output phase noted ($A = 1, B = 2, C = 3$).

This method generates sinusoidal output voltages by sequentially selecting input voltages during calculated switching cycles. The major problem of this method is the limit of the

transformation ratio between the output and the input voltage (r) which did not exceed 0.5. Moreover, the input power factor depends on the output power factor so one of the advantages of MC has been removed [55].

To overcome these problems Alesina and M. Venturini have optimized this strategy [15] and the maximum output voltage has been increased to 86% of the input voltage by modifying the desired output voltage with including the third-order harmonic of the input voltage and that of the output [15, 46]. The prescribed output voltage curve covers almost the entire envelope of the input voltage. The third-order harmonic will be eliminated in a three-phase load. The new form of the connection generator function is:

$$m_{ij} = \frac{1}{3} \left(1 + \frac{2vj.vk}{Ve^2} + \frac{4r}{3\sqrt{3}} \sin(2\pi ft + \beta) \sin(6\pi ft) \right) \quad (1.13)$$

where β is the respective phase shift of the given phase, f is the reference frequency, and r is the modulation rate.

This method remains very limited because the frequencies of the input and output networks must be perfectly known. The frequency introduced into the calculation must be exactly equal to the frequency of the input network. This method does not achieve the imbalances or pollution of the input network and its implementation requires a PLL (Phase Locked Loop) to synchronize the command with the input network [15, 23].

In 1987, G. ROY [56] proposed another scalar modulation strategy. The modulation algorithm is based on the instantaneous ratio of the measured input phase voltages. The conversion matrix of the matrix converter is calculated as shown in [23, 46]. Again, a common-mode voltage of frequency 3 times the input and output frequency is added to achieve a ratio r equal to 86.7 % [15]. Although these last two modulations (scalar modulation of Roy [56] and Alesina and Venturini strategy [15]) are different, the results are similar [23].

In 2006, Ph. Delarue [57] showed that it was possible to overcome the major drawbacks of these two approaches, i.e. the use of a PLL and the determination of the sinusoidal terms linked to the voltages of input only by measuring the phase-to-neutral input voltages. This method achieves the maximum transformation ratio r (0.86). This modulation strategy makes it possible to maintain a sinusoidal voltage system at the output and a constant instantaneous power when the input network is unbalanced and/or polluted by harmonics; the input currents then have the same waveform as the input voltages.

Another modulation called PWM three intervals was proposed by Schuster [57] and verified experimental in [49]. It uses the indirect model of the MC which comes in two stages (rectifier stage and inverter stage). Based on this modulation, the input voltages

are rectified with a rectification function $[M_{rec}]$ to build a virtual DC voltage using in addition a modulation function τ to give a sinusoidal form to the input current.

The inverter stage generates the output voltages with a function $[M_{inv}]$ which is a function of the reference voltages. Then calculate the matrix $[M_{inv}] \cdot [M_{rec}]$ which represents the connection generator matrix for the nine switches of the MC. The latter is the input of a PWM modulator which generates the instantaneous connection functions. This method is also called undershoot modulation. To maximize the transformation ratio r , the rectifier stage selects the largest negative input voltage and the largest positive input voltage to have the maximum DC bus voltage. This method can work in over-modulation but has a cost of input distortions [49, 58].

There are other scalar methods like the one proposed by Ishiguro [59], the calculation of the conversion matrix is based on the construction of two positive phase-to-phase output voltages from the two largest positive phase-to-phase input voltages. This method is detailed in [23]. Another method called calculated PWM, uses the matrix converter control model [58]. Connection generating functions are generated based on reference voltages and input voltages, depending on the angle of the input voltage. This technique is also detailed in [23].

In order to have the connection functions, PWM modulators are offered. The simplest is the one proposed by Ishiguro [59]. It is based on the comparison of two duty cycles of a switching cell at the same carrier.

The carrier-based modulation or triangular-sinusoidal PWM is a well-known method, especially for controlling inverters. It has been applied for the matrix converter with several variants, for different topologies of the matrix converter [60]. Firstly this method was applied for the single-phase matrix converter [9] and it was named SPWM (Sinusoidal PWM). Then, this technique was generalized for the different configurations of MC.

In [60] Yoon and Sul analyzed and tested this technique on the direct MC in different cases, such as the case of over-modulation. For this purpose, they used an offset voltage and based on the variation of the slope of the carrier in order to synthesize a sinusoidal input current for a power factor and prescribed output voltages. The state of the switches is determined after comparison between the reference voltages with an imaginary carrier. The switching times are determined from the instants of intersection of the reference voltages with the carrier. This technique was tested in over-modulation and the voltage ratio exceeded 1.

The concept of Space Vector Modulation (SVM), known and traditionally used for controlling inverters, was extended to matrix converters by L. Huber and D. Borojevic in

1989 [19, 50]. This approach makes it possible to synthesize spatially and temporally the voltages to be applied to the load, but also the currents to be taken from the input network. It also allows to control the input power factor independently of the output power factor [23, 50]. In [44] the SVM method has been modified to adapt to the case where the input voltages are unbalanced. Three strategies are proposed.

In [61] a study of the input current was carried out for direct SVM modulation in the case where the input voltages are unbalanced. Two input current vector control strategies are proposed. The first strategy consists in having the input current vector follow the input voltage vector at each instant. The second solution consists in following the input current vector dynamically around the input voltage vector. A comparison between these two strategies has been made and the content of the input current harmonics has been analyzed in both cases. It is shown that strategy 2 minimizes the harmonic content without a large increase in the *RMS* value of the input current. These results were confirmed by an experimental study.

Wang et al. [34, 62] proposed a method that takes into account disturbances at the MC input. This method is based on the calculation of the ratio and the phase shift of the input current from the measurement of the module and the phase shift of the input voltage vector. This method made it possible to eliminate the influence of disturbances without using an additional circuit. These results have been validated experimentally.

In [63] a comparative study between SVM modulation and model Predictive Current(MPC) for a direct MC topology was developed. A comparison between their performances under the same conditions (same input quantities and same loads) was made with simulation and experimentally. This study showed that the MPC is more attractive than the SVM especially from the input current point of view and particularly in the case where the input voltages are distorted.

The Indirect Space Vector Modulation (ISVM) was proposed in 1995 by L. Huber [19] who described the matrix converter by an equivalent circuit combining a current rectifier with a voltage inverter connected by a virtual continuous link. The two power stages are directly connected by their continuous sides. The SVM is applied for each stage. It uses two active vectors and a null vector to synthesize the two reference vectors which are the output voltage and input current vectors (V_S and I_{in}). Then a synchronization between the two stages is operated. The results obtained are identical to those of direct SVM [19, 50].

For the IMC structure, most of the modulations applied to the MC are applicable to it except for the one that uses a rotational vector [2]. In [64] an ISVM modulation has been

modified to adapt to the case where the input voltages are unbalanced. This method is based on compensation with a reliable adjustment of the modulation index of the inverter stage, this was achieved by a simple mathematical derivation. In the beginning, the phase shift remains constant then the mv index is adjusted according to the disturbances of the DC bus voltage. A simulation and experimental results were presented. In [65] the implementation of SVM modulation for the IMC using a real-time DSP-FPGA interface has been studied.

In [66] a new SVM modulation has been proposed for IMC. This method aims to reduce the Common Mode Voltage (CMV) especially in the case of large voltage ratios. Its principle is to choose adequate active vectors and then arrange them in the inverter stage. This method reduces the common-mode peak-to-peak voltage without adding another circuit; moreover, it makes it possible to reduce the THD of the output voltages and the switching losses at the level of the inverter stage. Other (CMV) reduction techniques have been developed in the literature [66–68].

In [22] J.W. Kolar et al. studied different sparse type indirect topologies and they developed an SVM modulation combined with the zero current type switching method. This last method minimizes the switching losses in the switches. It made it possible to generate sinusoidal output voltages and currents but always with a limitation of the ratio r .

In [54] a new multi-level sparse IMC structure was developed with an adapted SVM modulation. This modulation was derived from the ISVM and the SVM applied to the 3-level inverter. The experimental verification of this modulation was carried out in [29]. The latter made it possible to generate multi-level voltages while keeping a sinusoidal shape of the output voltages and input currents. These results were compared with an indirect matrix converter and with a conventional multilevel converter (ITLMC) to identify the advantages and disadvantages. It is deduced that the performance of the ITLMC structure is better than that of the sparse structure [30].

In [21] a detailed study of sparse topology and ITLMC topology has been presented. SVM has been applied, analyzed and simulated on both topologies, experimental results have been presented.

In [69] Casadei, et al. have developed generalized modulation. This method is based on the Duty-Cycle Space-Vector (DCSV) approach, which consists in representing the state of the switches by space vectors. All other modulations such as SVM or Venturini and Alesina are considered special cases of this modulation. Three degrees of freedom are available to define the modulation law which allows control of the output voltages and the

power factor. In conclusion, it is noted that the SVM is the best solution which makes it possible to obtain a maximum ratio of voltage and optimization of the switching by suitable use of the zero vectors. In [70] a generalized modulation based on a new model of MC (singular value decomposition (SVD)) was presented.

In [34] the SVM indirect modulation has been adapted to the multi-modular topology used for high powers. The simulation was validated by an experimental study.

In [42] an optimal modulation for the Z-source IMC allowing to guarantee sinusoidal input and output quantities, flexible buck-boost and low noise level has been studied. Two approaches are applied, one with the use of zero vectors in the rectifier stage, this method is difficult to implement. The other approach does not use zero vectors in the rectifier stage, it can be implemented by a carrier based or with the SVM approach. Experimental results are presented in [40].

In [71] an indirect Z-source topology with a modulation allowing to counter the disturbances of the input network was studied. A nonlinear PID controller was used for the effects caused by the Z-source network and to improve the dynamics and robustness of the control. This topology with its control made it possible to maintain good output and input performance in the event of input network disturbances while maintaining a high r ratio. Its disadvantage is that it introduces elements of energy accumulation.

In [72] a generalized modulation of the DCSV type was developed for single-phase MC, MC (3x3), and MC (3x4) converters. This strategy ensures control of the input current even at zero active power. It is simple to implement and can be applied with MC N arms. In [73] a study of the implementation of this strategy on the DSP-FPGA platform was proposed for the MC (3x4). In [73] an additional phase was added to the MC in order to increase safety in the event of the loss of an arm. Fault-tolerant SVM modulation has been proposed. An experimental study was presented.

In [74] an indirect approach of the 4-arm MC was used. The 3-D SVM was used in the inverter stage and the rectifier stage was controlled by the reference current. This modulation has been implemented and tested experimentally for different conditions.

In [75] modulation schemes based on the carrier-based method are applied for direct CM control (3x9). In [2] modulations adapted to the three topologies of the hybrid-type matrix converter are proposed and tested.

1.9 Conclusion

In this chapter, a state of the art on matrix converters has been presented. The study includes instantaneous modeling as well as modeling with connection generating functions. The different MC topologies are also presented. These topologies are divided into several families. Then, the switching problem which is one of the concerns of the MC is thus presented. Protection of the MC is essential for safe operation. This is because the connection between the two AC systems is direct and since there is no natural path for the current in the case of switching problems, overvoltages that can destroy the power circuit of the MC are possible. Thereafter, the industrial implementation of MC for speed and/or frequency variation was presented. The latter remains a little limited and is dominated by Japanese companies and in particular the Yaskawa company. Then the state of the art of modulations is presented, which can be divided into two categories, scalar modulations, and vector modulations. These are the most widely used today.

Finally, given the vastness of the research work on this type of converter and the beginning of its industrial integration, it is expected that it will soon replace conventional AC-AC converters.

Chapter 2

Direct Torque Control of Dual Stator Induction Motor Supplied by Matrix Converter

“ no amount of guilt can change the past, and no amount of worrying can change the future.”

Umar ibn Al-Khattab

2.1 Introduction

In this chapter, we will study the direct torque control of dual stator induction machine supplied by two different topologies of matrix converters that are DMC and IMC.

We firstly start with the modeling of the dual stator induction machine in view of the control. Then we study the direct torque control. This technique was initially proposed for the inverter, thereafter, it was adapted to matrix converter. It bases on the principle of applying a voltage vector that allows obtaining torque and flux in a very precise band through a switching lookup table. This technique is less dependent on machine parameters compared to the Field-Oriented Control (FOC) and it allows faster torque response.

This technique is simulated using the two topologies of the MC for different ranges of speed and load torque.

2.2 Dynamic Model of DSIM

Multiphase machines exist as alternatives to the conventional three-phase machines to cover various undertakings, exceptional benefits such that the power split between the different phases, high dependability, reduced rotor losses, and torque ripples. Such advantages make the multiphase machine very favored in various fields in recent years. Among these multiphase machines and the most common is the DSIM. The objective of this part is to present the mathematical modeling of the DSIM. Its complexity has been relatively reduced by using a certain number of simplifying assumptions and by making a change of reference by applying Park's transformation to get around the difficulty of determining mutual inductances [76, 77].

2.2.1 Brief description of the DSIM

The DSIM consists of a stator containing two identical three-phase windings displaced by an electric angle γ ($\gamma = 30^\circ$) and a squirrel cage rotor. Figure 2.1 gives the position of the magnetic axes of the windings forming the phases of the machine. We can note that index 1 represents the relative parameters of the first star and index 2 for the second one. The angle θ_1 expresses the position of the rotor (a_r phase) with respect to the star 1 (phase a_{s1}) and θ_2 the position of the rotor with respect to the star 2 (phase a_{s2}) [78, 79].

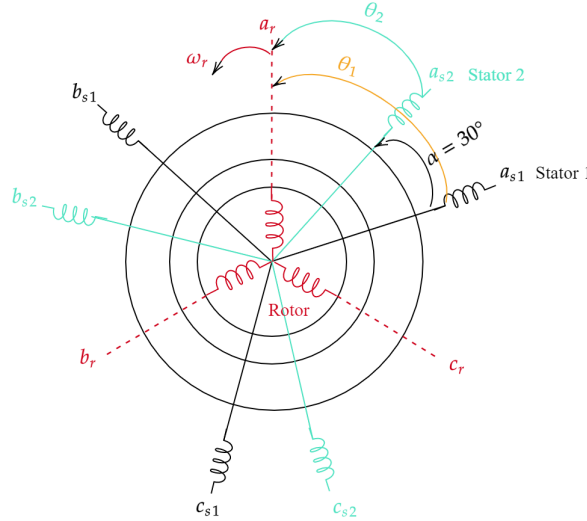


Figure 2.1: Schematic representation of a DISM

$$\theta_1 = \Omega_{mec}t + \theta_0 \quad (2.1)$$

$$\theta_2 = \theta_1 - \alpha \quad (2.2)$$

where Ω_{mec} represents the mechanical speed, and θ_0 stands for the initial position of the rotor with respect to the star 1.

2.2.2 Simplifying hypothesis

The distribution windings of the dual stator induction machine (DSIM) and its geometry is very complex for the analysis taking into account its exact configuration. Therefore, It is necessary to assume some simplifying assumptions [76, 77, 80].

- The machine is perfectly symmetrical.
- The machine is supposed to be "linear", that is to say, that the saturation and the effects of hysteresis magnetic circuits are neglected as well as the phenomena causing variations in resistance and inductance (temperatures, frequency).
- It is assumed that the two three-phase stator windings are balanced and identical.
- The spatial distribution of the magnetic fields along the air gap is sinusoidal.
- Power is supplied by an asymmetrical three-phase voltage system.

2.2.3 Three-phase model of the DSIM

Electrical equations

The electrical equations of the star 1, star 2, and rotor are respectively expressed by:

$$[v_{s1}] = [r_{s1}] [i_{s1}] + \frac{d}{dt} [\varphi_{s1}] \quad (2.3)$$

$$[v_{s2}] = [r_{s2}] [i_{s2}] + \frac{d}{dt} [\varphi_{s2}] \quad (2.4)$$

$$[v_r] = [r_r] [i_r] + \frac{d}{dt} [\varphi_r] \quad (2.5)$$

with $[v_{s1}] = \begin{bmatrix} v_{as1} & v_{bs1} & v_{cs1} \end{bmatrix}^t$, $[v_{s2}] = \begin{bmatrix} v_{as2} & v_{bs2} & v_{cs2} \end{bmatrix}^t$ and $[v_r] = \begin{bmatrix} v_{ar} & v_{br} & v_{cr} \end{bmatrix}^t$,
 $[i_{s1}] = \begin{bmatrix} i_{as1} & i_{bs1} & i_{cs1} \end{bmatrix}^t$, $[i_{s2}] = \begin{bmatrix} i_{as2} & i_{bs2} & i_{cs2} \end{bmatrix}^t$ and $[i_r] = \begin{bmatrix} i_{ar} & i_{br} & i_{cr} \end{bmatrix}^t$,
 $[\varphi_{s1}] = \begin{bmatrix} \varphi_{as1} & \varphi_{bs1} & \varphi_{cs1} \end{bmatrix}^t$, $[\varphi_{s2}] = \begin{bmatrix} \varphi_{as2} & \varphi_{bs2} & \varphi_{cs2} \end{bmatrix}^t$ and $[\varphi_r] = \begin{bmatrix} \varphi_{ar} & \varphi_{br} & \varphi_{cr} \end{bmatrix}^t$,
and $[r_{s1}] = \text{diag} \begin{bmatrix} r_{as1} & r_{bs1} & r_{cs1} \end{bmatrix}^t$, $[r_{s2}] = \text{diag} \begin{bmatrix} r_{as2} & r_{bs2} & r_{cs2} \end{bmatrix}^t$ and $[r_r] = \text{diag} \begin{bmatrix} r_{ar} & r_{br} & r_{cr} \end{bmatrix}^t$

with $r_{as1} = r_{bs1} = r_{cs1} = r_1$, $r_{as2} = r_{bs2} = r_{cs2} = r_2$, and $r_{ar} = r_{br} = r_{cr} = r_r$, where r_{as1} represents the stator resistances to the star 1, r_{as2} represents the stator resistances to the star 2, and r_r represents the stator resistances to the rotor.

Magnetic equations

The equations of stator flux (stator 1 and 2) and rotor as a function of the currents in matrix form are given by:

$$\begin{bmatrix} [\varphi_{s1}] \\ [\varphi_{s2}] \\ [\varphi_r] \end{bmatrix} = \begin{bmatrix} [L_{1,1}] & [L_{1,2}] & [L_{1,r}] \\ [L_{2,1}] & [L_{2,2}] & [L_{2,r}] \\ [L_{r,1}] & [L_{r,2}] & [L_{r,r}] \end{bmatrix} \cdot \begin{bmatrix} [i_{s1}] \\ [i_{s2}] \\ [i_r] \end{bmatrix} \quad (2.6)$$

The sub-matrices of the inductance matrix are:

$$\begin{aligned} [L_{1,1}] &= \begin{bmatrix} (L_1 + L_{ms}) & L_{ms}\cos\left(\frac{2\pi}{3}\right) & L_{ms}\cos\left(\frac{4\pi}{3}\right) \\ L_{ms}\cos\left(\frac{4\pi}{3}\right) & (L_1 + L_{ms}) & L_{ms}\cos\left(\frac{2\pi}{3}\right) \\ L_{ms}\cos\left(\frac{2\pi}{3}\right) & L_{ms}\cos\left(\frac{4\pi}{3}\right) & (L_1 + L_{ms}) \end{bmatrix} \\ [L_{2,2}] &= \begin{bmatrix} (L_2 + L_{ms}) & L_{ms}\cos\left(\frac{2\pi}{3}\right) & L_{ms}\cos\left(\frac{4\pi}{3}\right) \\ L_{ms}\cos\left(\frac{4\pi}{3}\right) & (L_2 + L_{ms}) & L_{ms}\cos\left(\frac{2\pi}{3}\right) \\ L_{ms}\cos\left(\frac{2\pi}{3}\right) & L_{ms}\cos\left(\frac{4\pi}{3}\right) & (L_2 + L_{ms}) \end{bmatrix} \\ [L_{r,r}] &= \begin{bmatrix} (L_r + L_{mr}) & L_{mr}\cos\left(\frac{2\pi}{3}\right) & L_{mr}\cos\left(\frac{4\pi}{3}\right) \\ L_{mr}\cos\left(\frac{4\pi}{3}\right) & (L_r + L_{mr}) & L_{mr}\cos\left(\frac{2\pi}{3}\right) \\ L_{mr}\cos\left(\frac{2\pi}{3}\right) & L_{mr}\cos\left(\frac{4\pi}{3}\right) & (L_r + L_{mr}) \end{bmatrix} \\ [L_{1,2}] &= \begin{bmatrix} L_{ms}\cos(\alpha) & L_{ms}\cos\left(\alpha + \frac{2\pi}{3}\right) & L_{mr}\cos\left(\alpha + \frac{4\pi}{3}\right) \\ L_{ms}\cos\left(\alpha - \frac{2\pi}{3}\right) & L_{ms}\cos(\alpha) & L_{ms}\cos\left(\alpha + \frac{2\pi}{3}\right) \\ L_{mr}\cos\left(\alpha - \frac{4\pi}{3}\right) & L_{ms}\cos\left(\alpha - \frac{2\pi}{3}\right) & L_{ms}\cos(\alpha) \end{bmatrix} \\ [L_{1,r}] &= \begin{bmatrix} L_{sr}\cos(\theta_r) & L_{sr}\cos\left(\theta_r + \frac{2\pi}{3}\right) & L_{sr}\cos\left(\theta_r + \frac{4\pi}{3}\right) \\ L_{sr}\cos\left(\theta_r - \frac{2\pi}{3}\right) & L_{sr}\cos(\theta_r) & L_{sr}\cos\left(\theta_r + \frac{2\pi}{3}\right) \\ L_{sr}\cos\left(\theta_r - \frac{4\pi}{3}\right) & L_{sr}\cos\left(\theta_r - \frac{2\pi}{3}\right) & L_{sr}\cos(\theta_r) \end{bmatrix} \end{aligned}$$

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

The sub matrices $[L_{2,1}] = [L_{1,2}]^t$, $[L_{r,1}] = [L_{1,r}]^t$ and $[L_{r,2}] = [L_{2,r}]^t$, with $L_{ms} = L_{mr} = L_{sr} = \frac{2}{3} L_m$. L_1 , L_2 , L_r : represent the maximum coefficients values of the inductance of star 1, star 2, and the rotor. L_{ms} represents the maximum value of the mutual stator inductances coefficients, L_{mr} represents the maximum value of the mutual rotor inductance coefficients, L_{sr} represents the maximum value of the mutual inductance coefficients between the star and the rotor, and L_m Cyclic mutual inductance between the star 1 and 2 and the rotor [78,81].

Electromagnetic torque expression

The electromagnetic torque is expressed by the partial derivative of the storage of electromagnetic energy with respect to the geometrical angle of the rotor rotation.

$$C_{em} = \frac{\partial W}{\partial \theta_{geo}} = p \frac{\partial W}{\partial \theta_{ele}} \quad (2.7)$$

with

$$W = \frac{1}{2} \{ [i_{s1}]^t [\varphi_{s1}] + [i_{s2}]^t [\varphi_{s2}] + [i_r]^t [\varphi_r] \} \quad (2.8)$$

The expression of electromagnetic torque is written by the following equation:

$$C_{em} = \frac{p}{2} \left\{ [i_{s1}] \frac{d}{d\theta_r} [L_{1,r}] [i_r]^t + [i_{s2}] \frac{d}{d\theta_r} [L_{2,r}] [i_r]^t \right\} \quad (2.9)$$

Mechanical equation

The equation that describes the movement of the DSIM is as follows:

$$C_{em} - C_r = J \frac{d\Omega}{dt} + K_f \Omega \quad (2.10)$$

with

$$\Omega = \frac{\omega_r}{p} \quad (2.11)$$

On the other hand

$$w_r = \frac{d\theta_r}{dt} \quad (2.12)$$

where J : Total inertia of rotating parts, K_f : The friction coefficient.

2.2.4 Modeling of the DSIM in biphas system

Park transformation

Park's transformation is often referred to as a two-axis transformation. Physically, the application of this transformation to the DSIM corresponds to a transformation of the three windings in two equivalent windings taking up the same considerations or aspects in terms of f.m.m, flux, torque, or at least an image that will be perfectly proportional to them.

From figure 2.2, we can deduce the relations between the angles as [82]:

$$\theta_{S1} = \theta_1 - \theta_r \quad (2.13)$$

$$\theta_{S2} = \theta_{S1} - \alpha = \theta_1 - \theta_r - \alpha \quad (2.14)$$

where

θ_{S1} The angle between the magnetic axis of the phase a_{S1} and d .

θ_{S2} The angle between the magnetic axis of the phase a_{S2} and d .

θ_r The angle between the magnetic axis of the phase a_r and d .

We define a unique transformation matrix for currents, voltages, and flux. It preserves the invariance of the power and is orthogonal. In addition, we assume that all the homopolar quantities are zero(balanced system). The transformation which translates the passage from the three-phase system (a, b, c) to the two-phase system (d, q) and reverses by the normalized matrices of Park [80, 83].

For the first star, it is defined as follows:

$$[P(\theta_1)] = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_{s1}) & \cos(\theta_{s1} - \frac{2\pi}{3}) & \cos(\theta_{s1} + \frac{2\pi}{3}) \\ -\sin(\theta_{s1}) & -\sin(\theta_{s1} - \frac{2\pi}{3}) & -\sin(\theta_{s1} + \frac{2\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (2.15)$$

$$[P(\theta_1)]^{-1} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_{s1}) & -\sin(\theta_{s1}) & \frac{1}{\sqrt{2}} \\ \cos(\theta_{s1} - \frac{2\pi}{3}) & -\sin(\theta_{s1} - \frac{2\pi}{3}) & \frac{1}{\sqrt{2}} \\ \cos(\theta_{s1} + \frac{2\pi}{3}) & -\sin(\theta_{s1} + \frac{2\pi}{3}) & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (2.16)$$

The three-phase voltage system for the star 1 is transformed as follows:

$$\begin{bmatrix} v_{ds1} \\ v_{qs1} \\ v_{0s1} \end{bmatrix} = [P(\theta_{s1})] \cdot \begin{bmatrix} v_{as1} \\ v_{bs1} \\ v_{cs1} \end{bmatrix} \quad (2.17)$$

For the second star 2 and the rotor, we replace in (2.15) and (2.16) θ_{s1} by θ_{s2} and then by θ_r respectively.

Model of the DSIM according to axis system (d,q)

The transformation of the DSIM according to the axes system (d, q) rotates at the speed ω_{coord} is illustrated in figure 2.2 [81, 84].

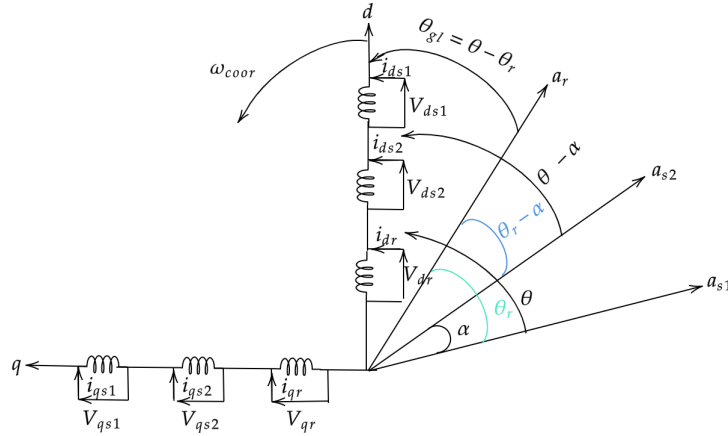


Figure 2.2: Representation of the DSIM model along the axes (d, q) .

Choice of the axis system

Three types of reference axis systems exist in practice. The choice is made according to the problem to be studied.

Axis system (α, β)

This system is stationary with respect to the stator, therefore $\theta_s = 0$.

Axis system (x, y)

This system is stationary with respect to the rotor, then $\theta_r = 0$.

Axis system (d, q)

The latter is stationary with respect to the electromagnetic field created by the stator windings, where $\omega_{\text{coor}} = \omega_s = \frac{d\theta_s}{dt}$.

Modeling of the DSIM in a coordinate axis system (d, q) linked to the rotating field

The model of the dual stator induction machine in a biphas axis (d, q) linked to the rotating field is obtained by using the normalized Park transformations in order to conserve the power. By applying these transformations to the voltages as well as to the stator fluxes, we obtain the electric and magnetic models respectively as:

$$\left\{ \begin{array}{l} V_{ds1} = R_{s1}i_{ds1} + \frac{d\varphi_{ds1}}{dt} - \omega_s\varphi_{qs1} \\ V_{qs1} = R_{s1}i_{qs1} + \frac{d\varphi_{qs1}}{dt} + \omega_s\varphi_{ds1} \\ V_{ds2} = R_{s2}i_{ds2} + \frac{d\varphi_{ds2}}{dt} - \omega_s\varphi_{qs2} \\ V_{qs2} = R_{s2}i_{qs2} + \frac{d\varphi_{qs2}}{dt} + \omega_s\varphi_{ds2} \\ V_{dr} = 0 = R_r i_{dr} + \frac{d\varphi_{dr}}{dt} - (\omega_s - \omega_r)\varphi_{qr} \\ V_{qr} = 0 = R_r i_{qr} + \frac{d\varphi_{qr}}{dt} + (\omega_s - \omega_r)\varphi_{dr} \end{array} \right. \quad (2.18)$$

$$\left\{ \begin{array}{l} \varphi_{ds1} = L_{s1}i_{ds1} + L_m(i_{ds1} + i_{ds2} + i_{dr}) \\ \varphi_{qs1} = L_{s1}i_{qs1} + L_m(i_{qs1} + i_{qs2} + i_{qr}) \\ \varphi_{ds2} = L_{s2}i_{ds2} + L_m(i_{ds1} + i_{ds2} + i_{dr}) \\ \varphi_{qs2} = L_{s2}i_{qs2} + L_m(i_{qs1} + i_{qs2} + i_{qr}) \\ \varphi_{dr} = L_r i_{dr} + L_m(i_{ds1} + i_{ds2} + i_{dr}) \\ \varphi_{qr} = L_r i_{qr} + L_m(i_{qs1} + i_{qs2} + i_{qr}) \end{array} \right. \quad (2.19)$$

where $\omega_s = \frac{d\theta}{dt}$, $\omega_r = \frac{d\theta_r}{dt}$, and $p\Omega = \frac{d\theta}{dt} - \frac{d\theta_r}{dt} = \omega_s - \omega_r$. By replacing the system of equations (2.19) in (2.18) and putting all in compact form, we will have:

$$[B][U] = [L]\left[\dot{I}\right] + \omega_{gl}[C][I] + [D][I] \quad (2.20)$$

with $[U] = \begin{bmatrix} V_{ds1} & V_{qs1} & V_{ds2} & V_{qs2} & V_{dr} & V_{qr} \end{bmatrix}$ is the command vector, $[I] = \begin{bmatrix} i_{ds1} & i_{qs1} & i_{ds2} & i_{qs2} & i_{dr} & i_{qr} \end{bmatrix}$ represents the state vector, and $[B] = \text{diag} \begin{bmatrix} 1 & 1 & 1 & 1 & 0 & 0 \end{bmatrix}$. The matrices C , L and D are given by:

$$C = \begin{pmatrix} 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & -L_m & 0 & -L_m & 0 & -(L_m + L_r) \\ L_m & 0 & L_m & 0 & (L_m + L_r) & 0 \end{pmatrix}$$

$$L = \begin{pmatrix} (L_{s1} + L_m) & 0 & L_m & 0 & L_m & 0 \\ 0 & (L_{s1} + L_m) & 0 & L_m & 0 & L_m \\ L_m & 0 & (L_{s2} + L_m) & 0 & L_m & 0 \\ 0 & L_m & 0 & (L_{s2} + L_m) & 0 & L_m \\ L_m & 0 & L_m & 0 & (L_r + L_m) & 0 \\ 0 & L_m & 0 & L_m & 0 & (L_r + L_m) \end{pmatrix}$$

$$D = \begin{pmatrix} R_{s1} & -w_s(L_{s1} + L_m) & 0 & -w_s L_m & 0 & -w_s L_m \\ w_s(L_{s1} + L_m) & R_{s1} & w_s L_m & 0 & w_s L_m & 0 \\ 0 & -w_s L_m & R_{s2} & -w_s(L_{s2} + L_m) & 0 & -w_s L_m \\ w_s L_m & 0 & w_s(L_{s2} + L_m) & R_{s2} & w_s L_m & 0 \\ 0 & 0 & 0 & 0 & R_r & 0 \\ 0 & 0 & 0 & 0 & 0 & R_r \end{pmatrix}$$

The state-space model of the system (2.20) is given by:

$$\frac{dI}{dt} = L^{-1}[BU - (w_{gt}C - D)I] \quad (2.21)$$

2.2.5 Power consumption and electromagnetic torque

The absorbed power by the DSIM in the axis system (d, q) , is expressed by:

$$P_a = v_{ds1}i_{ds1} + v_{qs1}i_{qs1} + v_{ds2}i_{ds2} + v_{qs2}i_{qs2} \quad (2.22)$$

By substituting voltages $\begin{pmatrix} v_{ds1} & v_{qs1} & v_{ds2} & v_{qs2} \end{pmatrix}$ with their expressions in equation (2.22), we find:

$$\begin{aligned} P_a = & [R_{s1}I_{ds1}^2 + R_{s1}I_{qs1}^2 - R_{s2}I_{ds2}^2 + R_{s2}I_{qs2}^2] \\ & + \left(\frac{d\varphi_{ds1}}{dt}i_{ds1} + \frac{d\varphi_{qs1}}{dt}i_{qs1} + \frac{d\varphi_{ds2}}{dt}i_{ds2} + \frac{d\varphi_{qs2}}{dt}i_{qs2} \right) \\ & + \omega_s (\varphi_{ds1}i_{qs1} - \varphi_{qs1}i_{ds1} + \varphi_{ds2}i_{qs2} - \varphi_{qs2}i_{ds2}) \end{aligned}$$

This expression consists of three terms, the first term corresponds to the losses by the Joule effect, the second term represents the change in electromagnetic energy (energy reserve), the last term is the electromagnetic power (P_{em}) [78, 82].

The expression of the electromagnetic torque is equal to

$$C_{em} = \frac{P_{em}}{\Omega_s} = p \frac{P_{em}}{\omega_s} \quad (2.23)$$

By substituting the fluxes ($\varphi_{ds1}, \varphi_{qs1}, \varphi_{ds2}, \varphi_{qs2}$) given by (2.19) in (2.24), we get:

$$C_{em} = p(\varphi_{ds1}i_{ds1} - \varphi_{qs1}i_{qs1} + \varphi_{ds2}i_{ds2} - \varphi_{qs2}i_{qs2}) \quad (2.24)$$

$$C_{em} = pL_m((i_{qs1} + i_{qs2})i_{dr} - (i_{ds1} + i_{ds2})i_{qr}) \quad (2.25)$$

From the rotor flux equations ($\varphi_{dr}, \varphi_{qr}$) expressed by (2.19), we have:

$$i_{dr} = \frac{1}{L_m + L_r} [\varphi_{dr} - L_m(i_{ds1} + i_{ds2})] \quad (2.26)$$

$$i_{qr} = \frac{1}{L_m + L_r} [\varphi_{qr} - L_m(i_{qs1} + i_{qs2})] \quad (2.27)$$

By substituting (2.26) and (2.27) in equation (2.25), we will have the relation of the electromagnetic torque expressed as a function of the stator currents and the rotor fluxes in the Park coordinate system (d, q) as :

$$C_{em} = p \frac{L_m}{L_m + L_r} (\varphi_{dr}(i_{qs1} + i_{qs2}) - \varphi_{qr}(i_{ds1} + i_{ds2})) \quad (2.28)$$

2.3 Direct Torque Control

Research on direct torque control was initially presented by "Takahashi and Depenbrock," in the middle of 1980 to develop the control of induction machines and to alternative the classical field-oriented control (FOC) [85]. This technique is characterized by the simple configuration and presents a remarkable dynamic performance as well as good robustness to variations in machine parameters. It's based on the control of the electromagnetic torque and stator flux [12, 86]. The control quantities (flux and torque) are calculated only by the quantities linked to the stator, and this is without the intervention of the mechanical sensor.

The principle of DTC is that the stator voltage vectors must be selected directly according to the differences between the references of the torque and flux and their real values. Current controllers followed by a PWM comparator in vector control are not required in

DTC control, also the machine parameters are not used except the stator resistance [46]. Consequently, the DTC has the advantage of being less dependent on machine parameters. In what follows, the DTC for the DSIM is developed for two different topologies of the MC as known as direct and indirect matrix converters. Meanwhile, we present the development of the conventional DTC (DTC for inverter) which is the basis of the DTC for the MC.

2.3.1 Classical DTC for two-level inverters

Principle

The DTC based on the stator flux orientation. The flux can be communicated in a referential identified with the stator of the machine by the following equation [10, 85]:

$$\varphi_{S1,2}(t) = \int_0^t (V_{S1,2} - R_{S1,2} I_{S1,2}) dt + \varphi_{S1,2}(0) \quad (2.29)$$

For the situation where an active voltage vector is connected amid an interim, $[0, T_e]$ the voltage drop can be neglect $(R_{S1,2} I_{S1,2})$ with respect to V_S . This last equation is then written:

$$\varphi_{S1,2}(t) = V_{S1,2} T_e + \varphi_{S1,2}(0) \quad (2.30)$$

The flux vector moves with the value of $V_{S1,2} T_e$, so the flux vector can be settled by the application of an appropriate voltage vector.

Figure 2.3 explain the evolution of the stator flux vector with the chosen voltage vector.

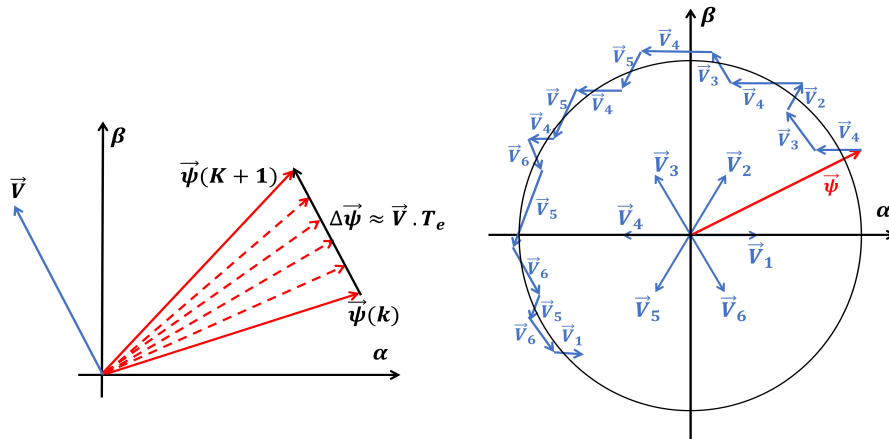


Figure 2.3: Evolution of the stator flux vector in (α, β) plan.

We can remark that the radial component of the voltage vector acts on the amplitude of the flux vector while the tangential component acts on its position.

So to increase the stator flux, it suffices to apply a voltage vector that is collinear to it and which has the same direction, and vice versa. A two-level hysteresis comparator is used for flux control as shown in figure 2.4.

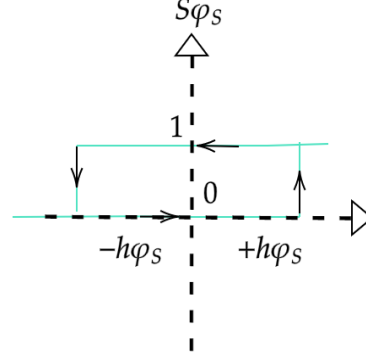


Figure 2.4: Two-level hysteresis comparator.

Its purpose is to keep the end of the stator flux vector φ_S in a circular crown. The hysteresis comparator works according to equation (2.31). This type of corrector makes it possible to obtain good dynamic performance of the flux [87, 88].

$$\left\{ \begin{array}{ll} si \Delta\varphi_S \succ h_{\varphi_S} & S_\varphi = 1 \\ si 0 \leq |\Delta\varphi_S| \leq h_{\varphi_S} et \frac{d\Delta\varphi_S}{dt} \succ 0 & S_\varphi = 0 \\ si 0 \leq |\Delta\varphi_S| \leq h_{\varphi_S} et \frac{d\Delta\varphi_S}{dt} \prec 0 & S_\varphi = 1 \\ si \Delta\varphi_S \prec -h_{\varphi_S} & S_\varphi = 0 \end{array} \right. \quad (2.31)$$

where $\Delta\varphi_S = |\varphi_S|^* - |\varphi_S|$ is the flux error, h_{φ_S} is the hysteresis band.

The electromagnetic torque of the asynchronous machine can be written in the following form:

$$C_{em} = p * \frac{Lm}{\sigma L_s * L_r} * \bar{\varphi}_S \cdot \bar{\varphi}_r \quad (2.32)$$

This equation can be written as follow:

$$C_{em} = p * \frac{M}{\sigma L_s * L_r} \|\varphi_S\| * \|\varphi_r\| * \sin(\gamma) \quad (2.33)$$

where γ is the angle between the stator flux and rotor flux.

The rotor flux is written as a function of the stator flux as follows:

$$\varphi_r = \frac{M}{L_s} * \frac{\varphi_S}{1 + j\omega\sigma Tr} \quad (2.34)$$

where σTr represents the speed of variation of the angle γ between the two fluxes. So if we can control the stator flux and its position, we can control the rotor flux and γ .

This can be done when the condition $T_e \ll \sigma T_r$ is respected. Therefore, the control of torque depends directly on the control of the rotation of the stator flux vector. In order to increase the torque value as much as possible, it suffices to apply a voltage whose vector is in advance quadrature to the stator flux vector [89]. Conversely, the reduction of the torque in terms of algebraic can be obtained quickly by applying a voltage vector with a strong lagging quadrature component as illustrated in Figure 2.5. To control the torque,

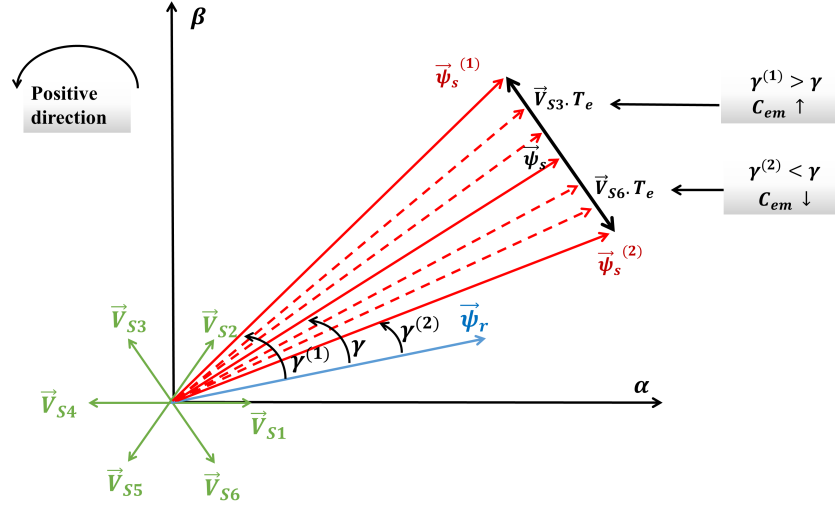


Figure 2.5: Adjustment of the electromagnetic torque by acting on the voltage vectors.

a three-level hysteresis comparator is used which is applied to control the motor in both directions of rotation as shown in Figure 2.6. The switching frequency of the switches is

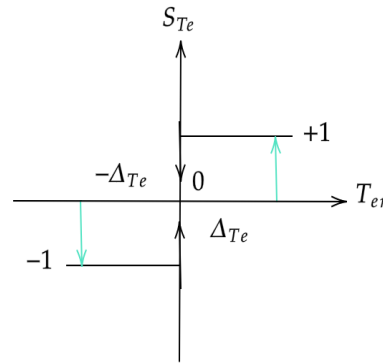


Figure 2.6: Three level hysteresis comparator.

not controlled, it varies according to the operating point and the hysteresis band. At low speed, the switching frequency is low and the acoustic noise increases.

Torque and Stator Flux Estimation

Stator flux

The amplitude and phase of the stator flux are estimated by the equation:

$$\begin{cases} \varphi_{S\alpha_{1,2}} = \int_0^t (V_{S_{1,2}} - R_{S_{1,2}} I_{S\alpha_{1,2}}) dt \\ \varphi_{S\beta_{1,2}} = \int_0^t (V_{S_{1,2}} - R_{S_{1,2}} I_{S\beta_{1,2}}) dt \end{cases} \quad (2.35)$$

The phase and magnitude of stator flux are described as follows:

$$|\varphi_{S_{1,2}}| = \sqrt{\varphi_{S\alpha_{1,2}}^2 + \varphi_{S\beta_{1,2}}^2} \quad (2.36)$$

$$\theta_{\varphi_{S_{1,2}}} = \arctg \left(\frac{\varphi_{S\beta_{1,2}}}{\varphi_{S\alpha_{1,2}}} \right). \quad (2.37)$$

Thus, the flux estimate only depends on the stator resistance as a parameter. Voltage and current are measurable quantities.

Torque

The electromagnetic torque can be estimated from the estimation of the stator flux and the measurement of the stator current.

$$\hat{C}_{em} = P (\varphi_{S\alpha_{1,2}} I_{S\beta_{1,2}} - \varphi_{S\beta_{1,2}} I_{S\alpha_{1,2}}) \quad (2.38)$$

Selection of the vectors

In view of the above and of Figure 2.7, note that we can define the sectors S_i of the flux vector according to its phase. The plan is divided into six sectors.

- To increase the amplitude of the stator flux, we can apply V_i , V_{i+1} or V_{i-1} .
- To reduce the amplitude of the stator flux, we can apply V_{i+2} , V_{i+3} or V_{i-2} .
- To increase the electromagnetic torque we can apply V_{i+1} or V_{i+2} .
- To decrease the electromagnetic torque we can apply V_{i-1} or V_{i-2} .

Applying a zero vector fixes the flux vector and therefore the torque decreases. By fixing the stator flux vector, the rotor flux vector tends to catch up with it, which decreases the angle γ , and therefore the torque decreases.

We can establish the classical DTC table of the sequences below proposed by Takahashi to control the stator flux and the electromagnetic torque in Table 2.1.

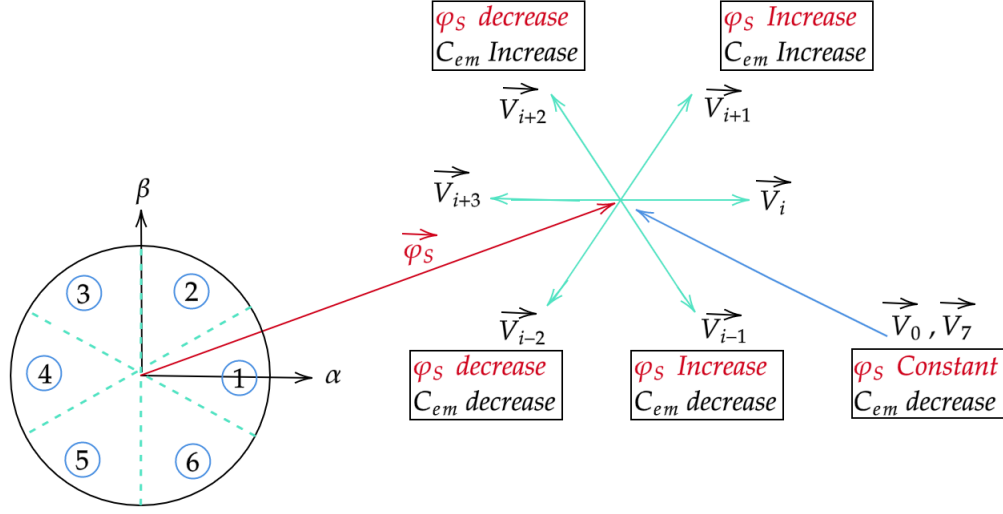


Figure 2.7: Hysteresis control: a) torque, b) flux

Table 2.1: Classical Table of DTC

$\Delta\psi_{S_{1,2}}$	ΔC_{em}	S_1	S_2	S_3	S_4	S_5	S_6
1	1	V_2	V_3	V_4	V_5	V_6	V_1
1	0	V_7	V_0	V_7	V_0	V_7	V_0
1	-1	V_6	V_1	V_2	V_3	V_4	V_5
0	1	V_3	V_4	V_5	V_6	V_1	V_2
0	0	V_0	V_7	V_0	V_7	V_0	V_7
0	-1	V_5	V_6	V_1	V_2	V_3	V_4

2.3.2 DTC of matrix converter

Direct matrix converter

“Casadei” is the first to introduce the use of MC with DTC in [90], various researches have been proposed in literature for improving its dynamic performances and to overcome its drawbacks.

The DTC of DMC is based on the table of classical DTC, and the control of input power factor for fixed that to one, a hysteresis comparator is added to confirm that, whatever is the sector which the input voltage vector is in, the MC takes any time two switching configurations with several directions for each VSI output vector selection by the classical DTC, this direction allows the possibility to control the input power factor by applied

one to increase the angle and the second to decrease. For example, if the vector chosen by the classical DTC is the vector V_1 , the configurations of the MC which correspond to it are $\pm 1, \pm 2, \pm 3$. These vectors cannot all be retained. We only use vectors that have the same direction and maximum amplitude [12].

It is proven that regardless of the sector of the input voltage vector, each vector chosen by the classical DTC gives the possibility of using two vectors of the MC (two configurations) [12,86]. If the sector of V_{in} is equal to 1, the vectors +1 and -3 can be used as explained in figure 2.8. This allows the possibility of controlling another major parameter which

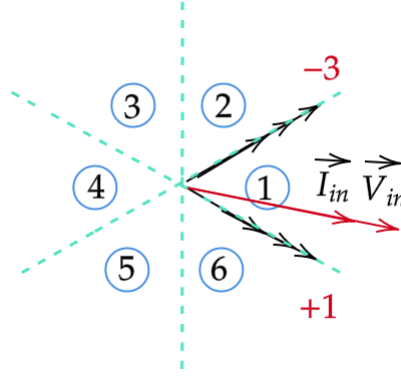


Figure 2.8: Input voltage, current sector and vectors +1, -3.

is the input power factor. Thereby, a third variable to control will be the mean value of $\prec \sin(\Psi_E) \succ$. If we want to function at unity input power factor, the term $\prec \sin(\Psi_E) \succ$ must be zero. The mean value of $\prec \sin(\Psi_E) \succ$ is obtained by applying a low pass filter to its instantaneous value. A third two-level hysteresis comparator is added as shown in figure 2.9.

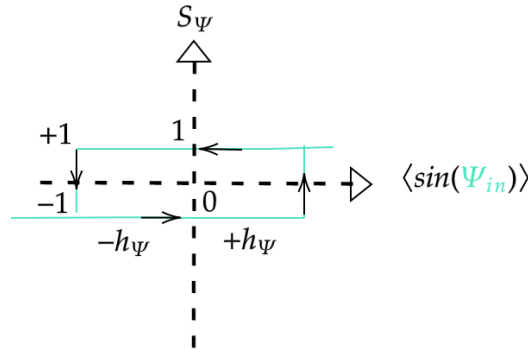


Figure 2.9: Hysteresis comparator of $\prec \sin(\Psi_{in}) \succ$.

Assume that we want to control the value of $\prec \sin(\Psi_E) \succ$ at 0, so $\prec \sin(\Psi_E) \succ = 0$. If the value of $\prec \sin(\Psi_E) \succ$ is positive $C_\Psi = 1$, therefore the input current vector lags behind the input voltage vector. The configuration -3 must be applied for it to catch up with the input voltage. On the other hand, if the estimated value of $\prec \sin(\Psi_E) \succ$ is negative therefore $C_\Psi = -1$, the input current vector is ahead of the input voltage vector. The configuration +1 must be applied to delay it. The all probability switching configuration of MC used in DTC is gives in the table 2.2.

Table 2.2: Table of DTC based on DMC

Sector of V_{in}	1		2		3		4		5		6	
Ψ_{in}	+1	-1	+1	-1	+1	-1	+1	-1	+1	-1	+1	-1
V_1	-3	1	2	-3	-1	2	3	-1	-2	3	1	-2
V_2	9	-7	-8	9	7	-8	-9	7	8	-9	-7	8
V_3	-6	4	5	-6	-4	5	6	-4	-5	6	4	-5
V_4	3	-1	-2	3	1	-2	-3	1	2	-3	-1	2
V_5	-9	7	8	-9	-7	8	9	-7	-8	9	7	-8
V_6	6	-4	-5	6	4	-5	-6	4	5	-6	-4	5

For the case of the unity power factor operation, the input current I_{in} and voltage V_{in} vectors have the same direction, The computation of its sector can be based on the phase of the input voltage. Figure 2.11 shows the overall diagram of the DTC for DMC.

classical table of DTC controls the inverter stage, which is described before for controlling the torque and the flux of DSM.

Control of rectifier

The purpose is to get the maximum output voltage V_{dc} of the virtual rectifier and to fix the input power factor to one, based on SVM, the reference vector can be synthesized with these two adjacent vectors I_γ , I_δ and a null vector I_0 . This is illustrated in figure 2.12 [10].

$$I_{in} = d_\gamma I_\gamma + d_\delta I_\delta \quad (2.39)$$

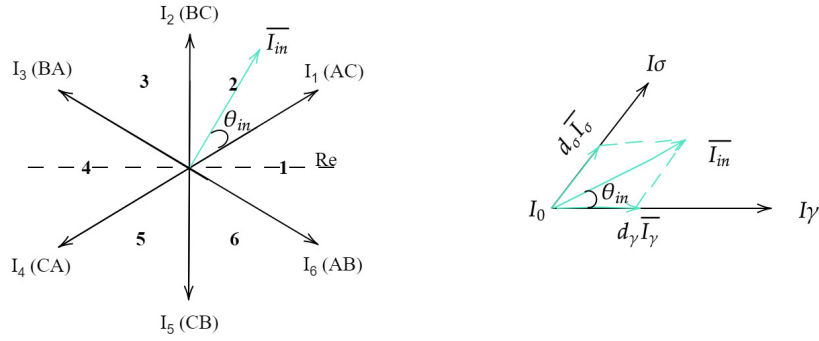


Figure 2.12: Input current modulation for rectification stage

The duty cycles of each sector are defined:

$$\begin{cases} d_\gamma = m_c \sin(\theta_{in}) \\ d_\delta = m_c \sin(\frac{\pi}{3} - \theta_{in}) \\ d_0 = 1 - d_\gamma - d_\delta \end{cases} \quad (2.40)$$

where $m_c = \frac{|I_{in}|}{I_{rec}^{rec}}$ and $0 < \theta_{in} < \frac{\pi}{3}$. The null vector I_0 is eliminated in order to maximize the voltage V_{dc} . The two cyclic ratios are normalized according to the following equation:

$$\begin{cases} d'_\gamma = \frac{d_\gamma}{d_\gamma + d_\delta} = \frac{\sin(\theta_{in})}{\cos(\frac{\pi}{6} - \theta_{in})} \\ d'_\delta = \frac{d_\delta}{d_\delta + d_\gamma} = \frac{\sin(\frac{\pi}{3} - \theta_{in})}{\cos(\frac{\pi}{6} - \theta_{in})} \\ d_0 = 0 \text{ and } d'_\gamma + d'_\delta = 1 \end{cases} \quad (2.41)$$

Control of inverter

For the inverter stage, the conventional DTC is applied. Therefore, according to the torque, flux, and sector, the conventional DTC chooses the inverter vector that makes it conceivable to guarantee the regard bands of the error bands (flux and torque) as explained previously. The following figure 2.13 explained this technique.

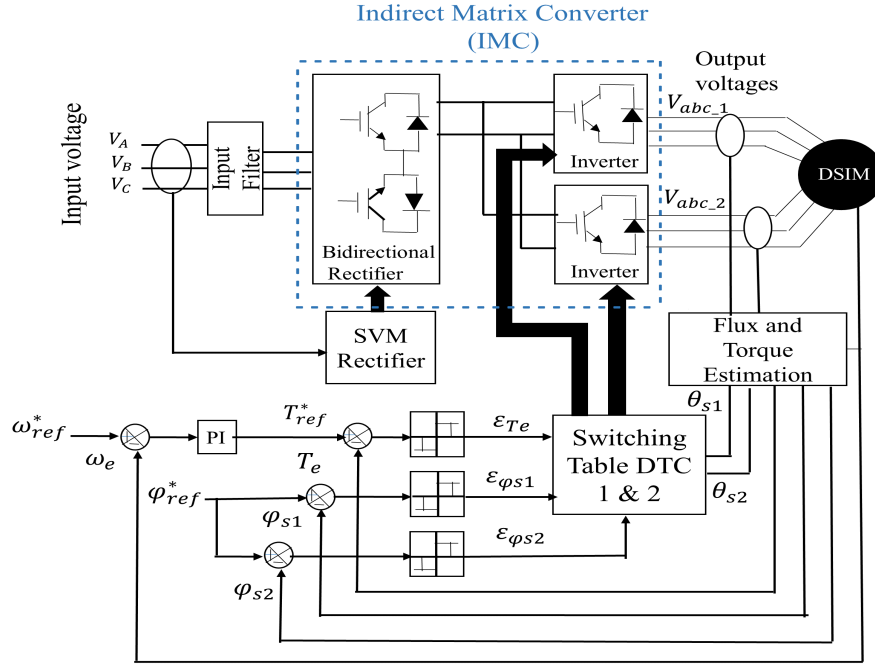


Figure 2.13: DTC based on IMC.

2.4 Simulation Results

In order to analyze the performance of the direct torque control of DSIM fed by direct and indirect matrix converter, several simulation tests have been carried out for different operating conditions with the following parameters.

Noting that the choice of hysteresis bands is related to the possible switching frequency of the matrix converter as well as the frequency of execution of the control program. The DTC for MC has been applied to DSIM, the parameters of the machine are given in the appendix A (Table 2.3). Speed control has been added using a PI-type corrector. The output of the regulator is the reference torque.

The DTC has been simulated with different references of speed and resistant torque to verify its static and dynamic performance.

Table 2.3: Simulation parameters

Input voltage	Hysteresis bands	Input filter	PI Controller
$\mathbb{V}4220/380\text{V } 50\text{Hz}$	Flux $h_\varphi=0.01 \text{ Wb}$	$C_f = 18\mu\text{f}$	$K_P=37.5$
2-2	Torque $h_{T_e}=0.2 \text{ N.m}$	$L_f = 3\text{mH}$	$K_I=0.35$
	$\prec \sin(\Psi_{in}) \succ \text{B}\Psi = 0.001$		

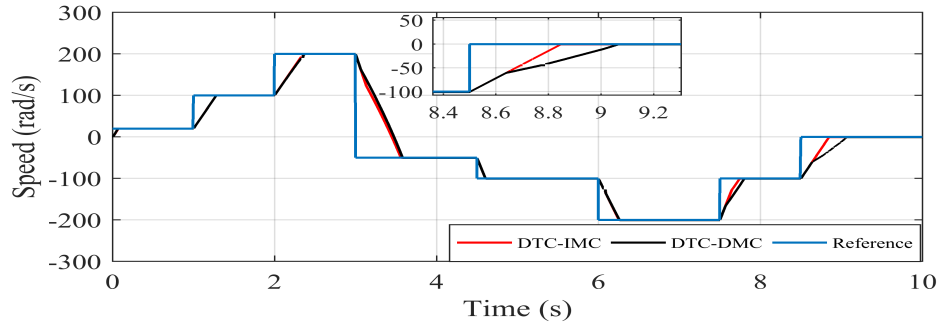


Figure 2.14: Speed.

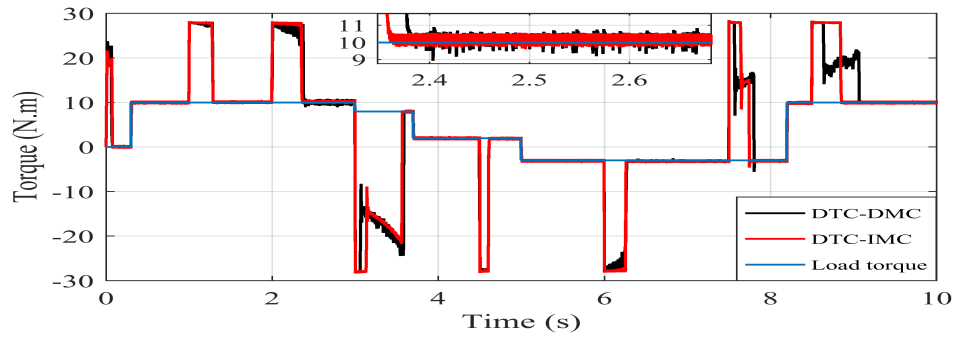


Figure 2.15: Electromagnetic torque.

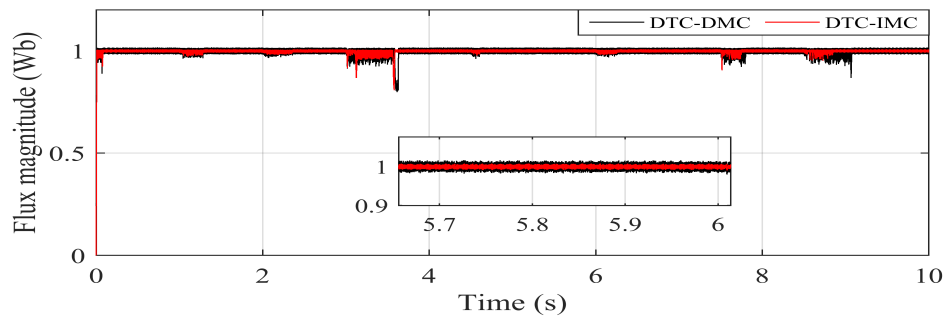


Figure 2.16: Stator flux magnitude.

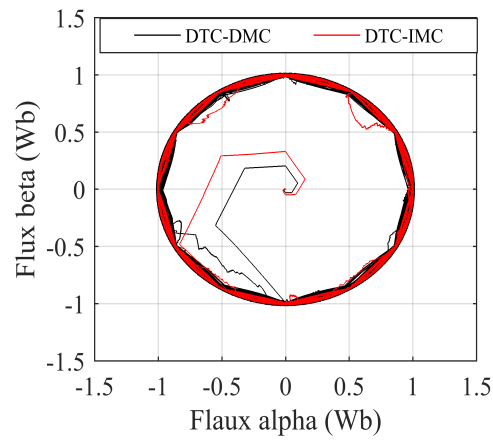


Figure 2.17: Stator flux circular trajectory.

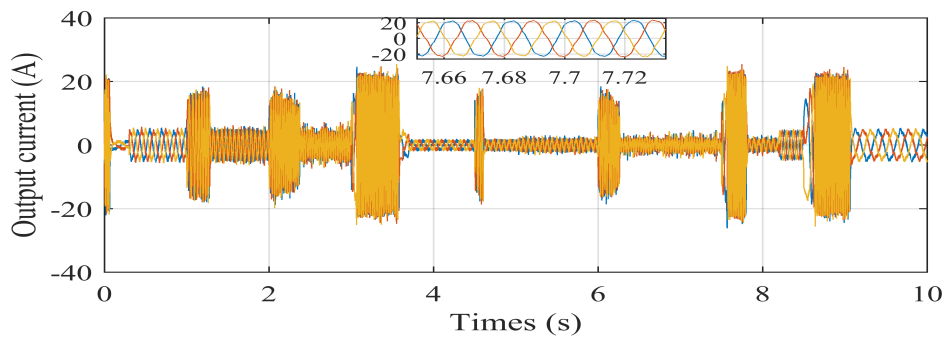


Figure 2.18: Output currents of DSIM with DMC.

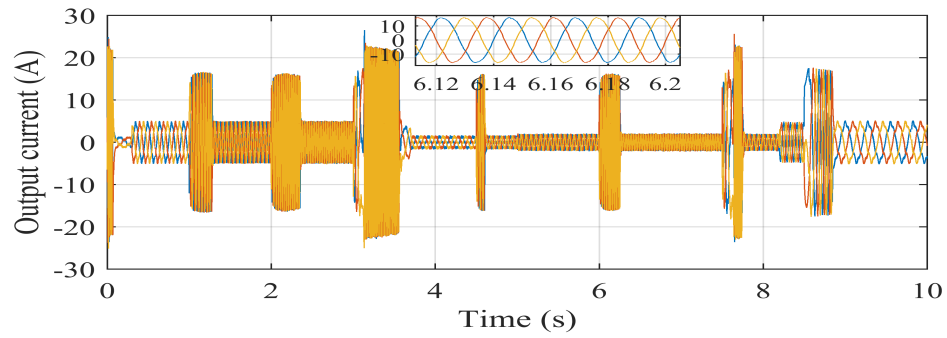


Figure 2.19: Output currents of DSIM with IMC.

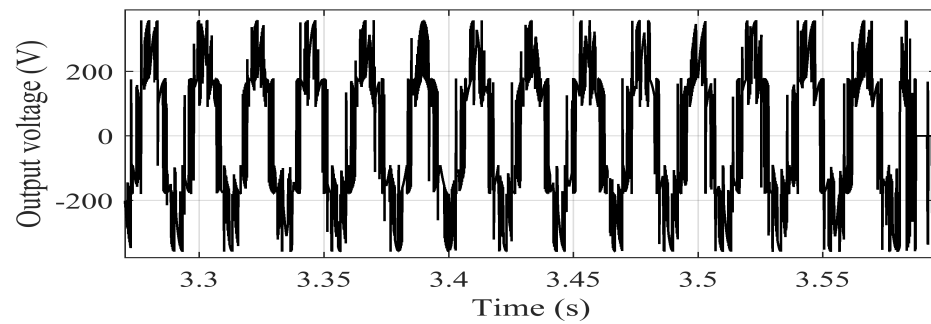


Figure 2.20: Output voltage of DMC.

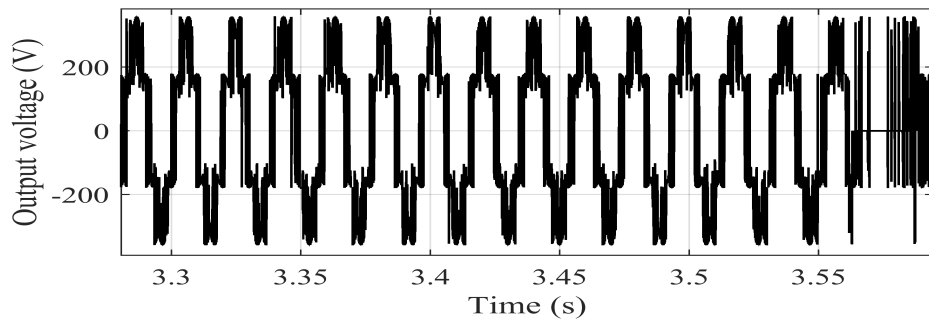


Figure 2.21: Output voltage of IMC.

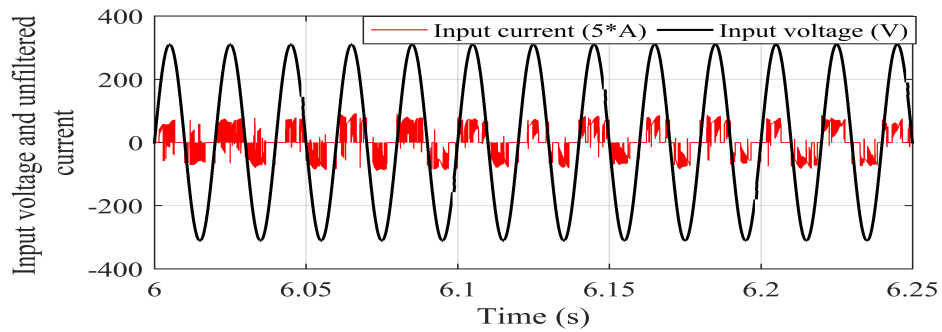


Figure 2.22: Input voltage and unfiltered current of DMC.

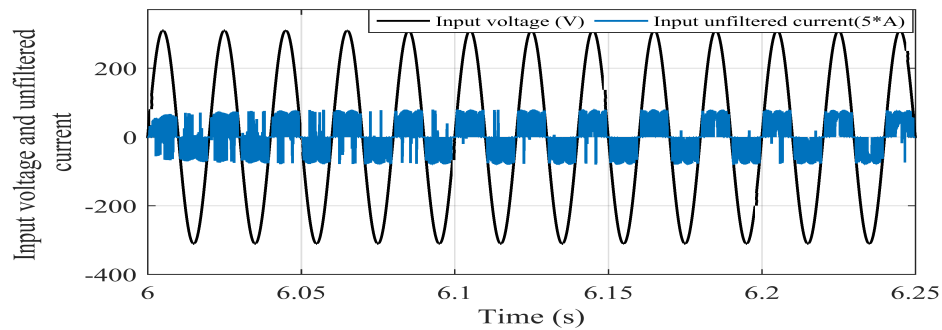


Figure 2.23: Input voltage and unfiltered current of IMC.

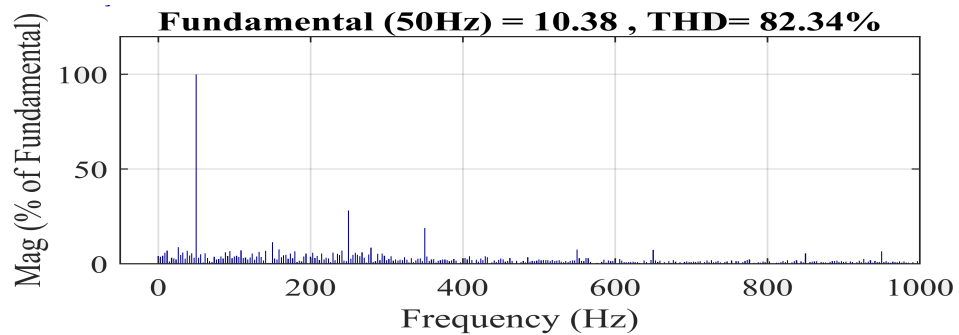


Figure 2.24: THD of input unfiltered current of DMC.

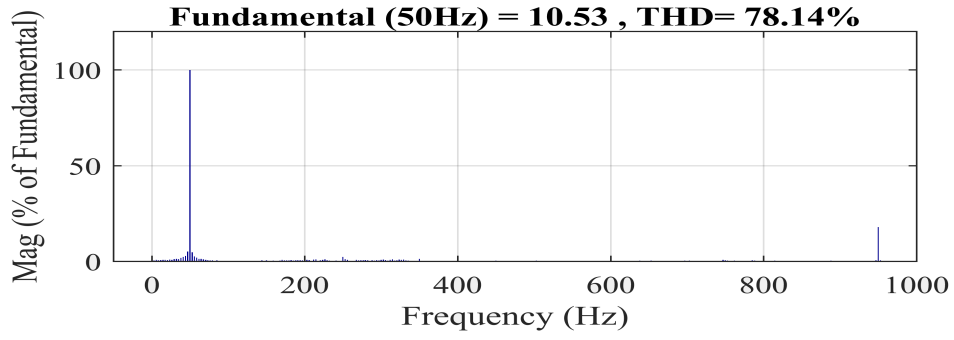


Figure 2.25: THD of input unfiltered current of IMC.

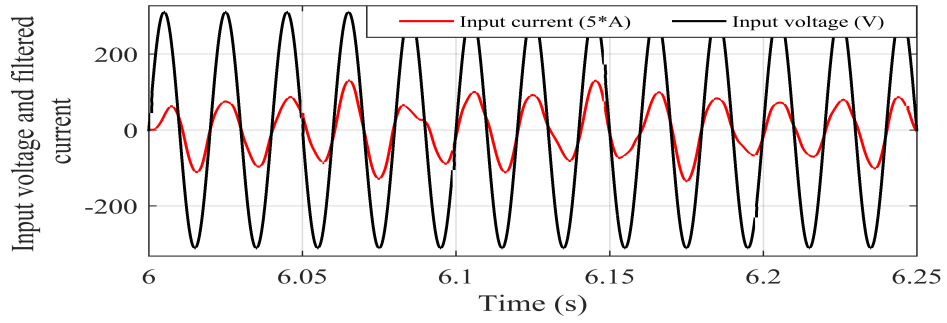


Figure 2.26: Input voltage and filtered current of DMC.

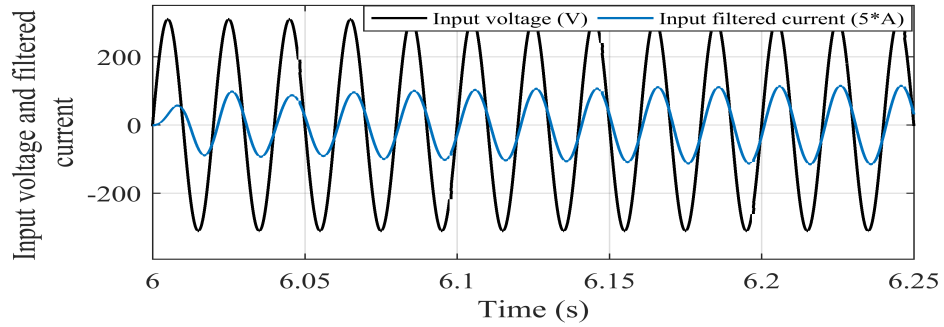


Figure 2.27: Input voltage and filtered current of IMC.

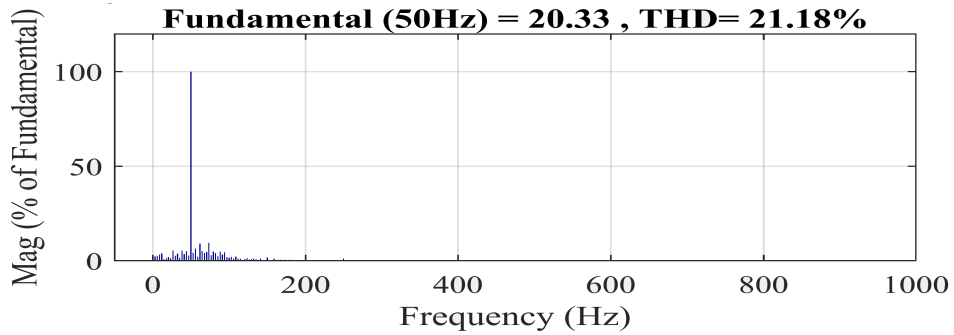


Figure 2.28: THD of input filtered current of DMC.

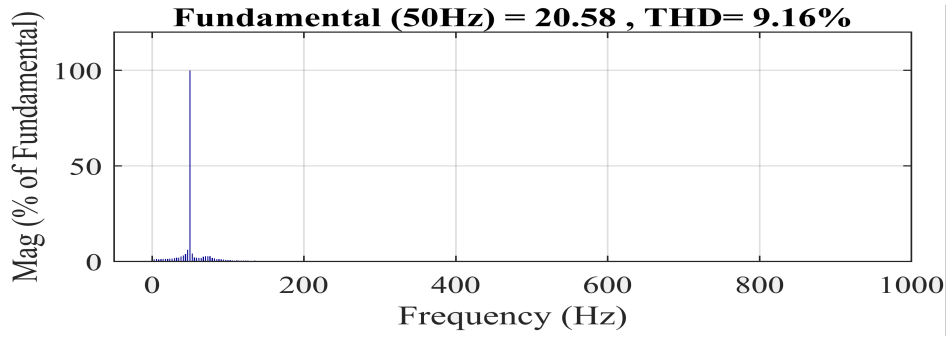


Figure 2.29: THD of input filtered current of IMC.

From the analysis of simulation results of the two DTC, we can comment the following points:

- The machine's speed pursues exactly its set point, also when the load torque is varied as shown in figure 2.15. for both techniques. The reference is chosen to include different scenarios (constant, low, average and high speed, inversion and zero speed), one can see that the DTC based on IMC has a faster response, very small tracking error, and high robustness against external disturbances in all areas compared to the DTC with DMC.
- Figure 2.16 presents the electromagnetic torque for two DTC, the couple is equal to the load torque in steady-state and respects its band for two DTC, but in the transient steady the DTC of DMC contains some overshoot of its band. Furthermore, it is remarked that the dynamic of the torque in DTC-IMC is faster and has much fewer ripples compared to the DTC-DMC.
- The flux as well follows its reference and respects its hysteresis band for both DTC figures (2.17, 2.18), there are overshoots in the case of low speeds and especially when changing the direction of rotation, although DTC of IMC is less undulating than DMC.
- Figures (2.19, 2.20) show the output current, which is sinusoidal forms, and changed their amplitude and frequency according to speed reference and load torque and we can see that the form of the current for IMC it's less polluted in harmonics and its temporal variation curve is very close to the sinusoid compared to DMC.
- The output voltage in both figures (2.21, 2.22) it is changed according to the frequency and the load torque. Then the DTC based on IMC gives the best form and low harmonics distortion.

- The input voltage with the unfiltered and filtered current Figure (2.23-2.30), is in phase, that is to say, the operation with unity of power factor for both techniques, and we can remark the highest quality and low harmonics in the DTC of IMC, because the fictive rectifier is switching by constant frequency (SVM).
- The results obtained with the IMC are in general, better than those obtained with DMC.

2.5 Conclusion

Direct torque control of dual stator induction motor fed by two types of matrix converter namely direct and indirect has been presented in this chapter. The simulation results obtained illustrate that the DTC with IMC is superior compared to the DMC, it yields good electrical and electromagnetic execution such as the stator flux, electromagnetic torque, and total harmonic distortion of the input current in all operating conditions. Furthermore, the use of the IMC simplifies significantly the control scheme and makes its implementation easier. However, Controlling the switching frequency is still a problem for this strategy (DTC with MC). The next chapter will be devoted to fuzzy logic control to improve the performance of the DTC control of the DSIM, either in terms of minimizing ripples in terms of reducing the switching frequency.

Chapter 3

Intelligent DTC of DSIM Powered by Two-Stage Matrix Converter Based on Stator Resistance Estimator

“ *Acquire wisdom from the story of those who have already passed.* ”

Uthman ibn Affan

3.1 Introduction

In this chapter, we will present a brief overview of fuzzy logic theory and its basic principles. Then, after describing the structure of a fuzzy controller, we apply it to the switching table to improve the dynamic performances of Direct Torque Control (DTC) for Dual Stator Induction Motor (DSIM) powered by a two-stage matrix converter, by using a fuzzy inference system instead of the hysteresis controllers (HC) and switching table. Moreover, a robust Fuzzy-PI (F-PI) speed controller is designed to enhance the performance of the speed control loop and robustness with respect to load disturbances. Then, in order to overcome the influence of the parameter variation influence(stator resistance), we have adopted an estimator to compensate its variation. The performance

and robustness of this method are shown by simulation results and compared to those of conventional DTC-IMC control.

3.2 History and principle of fuzzy logic

Fuzzy logic is a technique of artificial intelligence, just like neural networks and genetic algorithms, has been the subject of a significant amount of scientific research in recent years. In 1965, Professor L. Zadeh proposed for the first time, the theoretical basis of this logic in a celebrated article entitled (Fuzzy set) [91]. At that time, the theory of fuzzy logic was not taken seriously. Indeed, computers, with their exact functioning (1 or 0), was started to spread on a large scale. On the other hand, fuzzy logic made it possible to deal with non-exact variables whose value can vary between 1 and 0. Mamdani was the first to apply this new formalism. Initially, this theory was applied in non-technical fields, such as medicine and commerce [92]. Thereafter, the concept of fuzzy logic has been extended to many areas. One of the successful examples, fuzzy control, which has found many applications in the industry [93].

3.2.1 Fuzzy controller

A fuzzy system (FS) is a nonlinear relationship that allows you to take digital data (net inputs), pass them into a fuzzy domain, and then have a scalar output (net output).

A fuzzy controller is just a fuzzy system designed to control a process. Its general structure is shown schematically in figure 3.1 [83, 94].

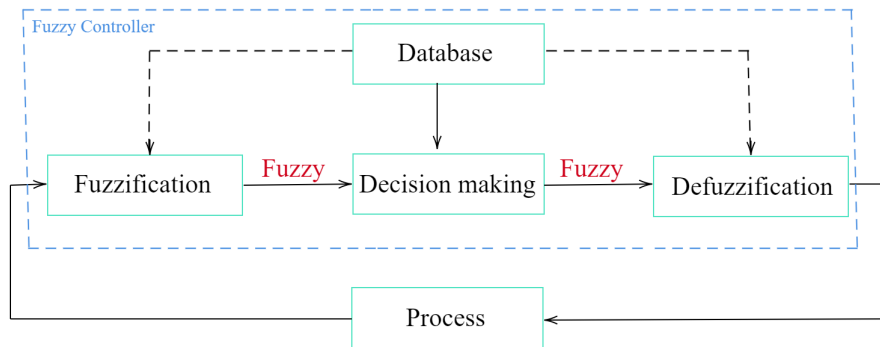


Figure 3.1: General block diagram of a fuzzy controller.

A fuzzy controller typically passes through the following steps:

- The choice of the fuzzification strategy,

- Establishing the basic rules,
- The choice of the inference method,
- The choice of the defuzzification strategy.

Fuzzification: It consists of transforming real quantities into linguistic variables that are associated with a database with sets characterizing them.

The basic rules: those are the rules that make it possible to link fuzzy input and output variables, they have the form: if-then.

The inference method: It calculates the fuzzy set associated with the command and is done by the fuzzy inference operations and the aggregation of the rules. Fuzzy inference relies on the use of a fuzzy implication operator for each rule to be analyzed. Two approaches are commonly used, Mamdani and Larsen.

The defuzzification: This stage consists of transforming the fuzzy information decided by the inference mechanism into a numerical or physical magnitude to realize the control law of the system.

3.3 Implementation of fuzzy logic to DTC of DSIM fed by IMC

In the classical DTC, the flux and torque errors are directly used to select the inverters switching state, without considering a large or small error. Moreover, the switching state selection for a big error that just exists in the variation of the reference or at start-up is the same as in natural operation. This clearly suggests a weak dynamic response. The system's dynamic performance can be ameliorated if the voltage vector is chosen as reported to the value of the error of the flux and torque and the placement of the flux. In this part, the DTC based on fuzzy logic control is proposed to improve the dynamic performances of the conventional DTC. This technique proposes to replace the hysteresis comparators, switching table and the PI controller by fuzzy controller.

The overall structure of the fuzzy controller is shown in figure 3.2.

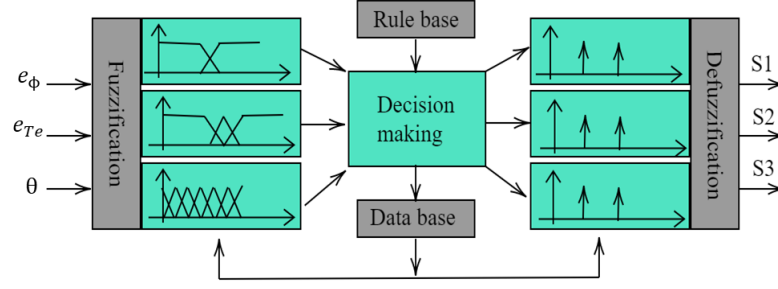


Figure 3.2: Fuzzy controller internal structure.

In the proposed system, for the inputs, triangular and trapezoidal Membership Functions (MFs) are used. While for outputs, singleton functions are used. The selection of each MF and its parameters were marked depending on the conventional DTC results and according to the effects of the possible variation of torque and flux error. The selection of MFs range was based on the hysteresis controllers'band-width', see figures (3.3 and 3.4), but not restricted to its minimum and maximum so all the possibilities can be estimated. Six sectors were used for the sector angle identification. any triangular MF of a sector i ($i = 1$ to 6) is fixed at $j \cdot 60^\circ$ ($j = 1$ to 6) as shown in figure 3.5. MFs have interfered because in their intersections the control action is the same.

3.3.1 Fuzzification

Flux error

We used two fuzzy sets that represent the state of the flux error (N for negative error, P for positive error). Their MFs are shown in figure 3.3.

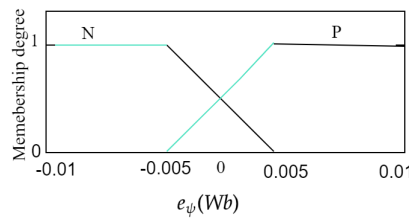


Figure 3.3: Fuzzy MFs of flux error

Torque error

For the torque error, we used three fuzzy sets (N for negative error, Z for zero error, P for positive error). Their MFs are shown in figure 3.4.

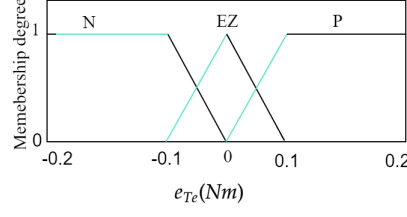
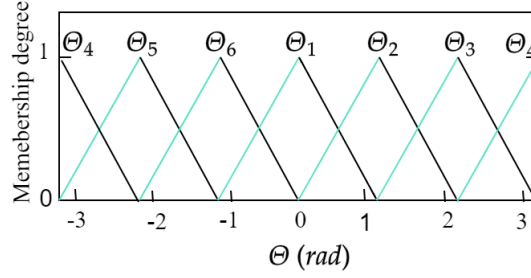


Figure 3.4: Fuzzy MFs of torque error

The sector angle

We selected six fuzzy sets presented in figure 3.5. This choice of six sectors has been selected to improve the precision and the performance of the control.

Figure 3.5: Fuzzy MFs of angle θ_s

The output of the FLC

The output of the fuzzy regulator is collected of three sub-outputs showing three switchings of the inverter (S1, S2, and S3), the being of the discourse of any output is split into two fuzzy sets (zero and one only) whose MFs are chosen in singleton forms as appeared in figure 3.6.

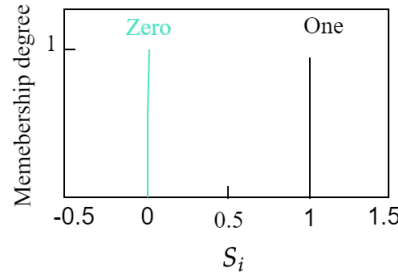


Figure 3.6: Fuzzy membership functions of output.

3.3.2 The inference method and the Fuzzy rules

The aim is to obtain a rapid response of the torque and a stator flux around a set value. Thus, the fuzzy rules should be designed according to the output and input variables based on the classical DTC switching table 2.1. These rules are collected in table 3.1. The inference method is based on the Mamdani procedure (Max-Min) chosen for its low complexity of the implementation and its good results given. This can be described as:

$$IF(e_\varphi \text{ is } A_{p,n}) \text{ AND } (e_{Te} \text{ is } B_{p,z,n}) \text{ AND } (\theta \text{ is } \theta_{1,6}) \text{ THEN} \\ (S_1 \text{ is } D_{0,1}) \text{ AND } (S_2 \text{ is } D_{0,1}) \text{ AND } (S_3 \text{ is } D_{0,1})$$

where $A_{p,n}$, $B_{p,z,n}$, $\theta_{1,6}$ and $D_{0,1}$ are the fuzzy sets used for the flux error, the torque error and the sector identification angle of the stator flux and the switching states of the inverter respectively. We can take an example for explaining this:

$$IF(e_\varphi \text{ is } P) \text{ AND } (e_{Te} \text{ is } P) \text{ AND } (\theta \text{ is } \theta_1)$$

Which means that the value of the reference of the stator flux is greater than the estimated value, AND the reference of the electromagnetic torque is greater compared to the estimated torque AND the stator flux is located in sector 1, therefore, we must choose an appropriate vector which increases the stator flux AND will increase the electromagnetic couple. Then the appropriate control vector in sector 1 for this case according to table 3.1 and table 2. 1 is V_2 . The complete rule will be,

$$IF(e_\varphi \text{ is } P) \text{ AND } (e_{Te} \text{ is } P) \text{ AND } (\theta \text{ is } \theta_1) \text{ THEN } (S_1 \text{ is } 1) \text{ AND } (S_2 \text{ is } 1) \text{ AND } (S_3 \text{ is } 0)$$

All these rules are grouped in the following inference table:

Table 3.1: The DTC based on the fuzzy rules

$\Delta\varphi_{s1,2}$	ΔC_{em}	θ_1	θ_2	θ_3	θ_4	θ_5	θ_6
P	P	110	010	011	001	101	100
P	Z	111	000	111	000	111	000
P	N	101	100	110	010	011	001
N	P	010	011	001	101	100	110
N	Z	000	111	000	111	000	111
N	N	001	101	100	110	010	011

3.3.3 Defuzzification

This stage aims to convert the output fuzzy results to non-fuzzy values, the Mean Of Maximum (MOM) method is used for the proposed method, so permitting an efficient control of the system, it's based on determining the output as existence the abscissa identical to the average value of the resulting maximum abscissa.

3.3.4 PI FLC

The Fuzzy-PI controller is the same as the PI controller, it tracks the variation of output torque. The input of the F-PI is the error and its derivative, for determining the direction of variation, while the output is the reference torque corresponding to the control. k_1 , k_2 , and k_3 are variable normalization gains that are used to adjust the inputs and output at appropriate values without changing the control structure. Figure 3.7 describes this controller.

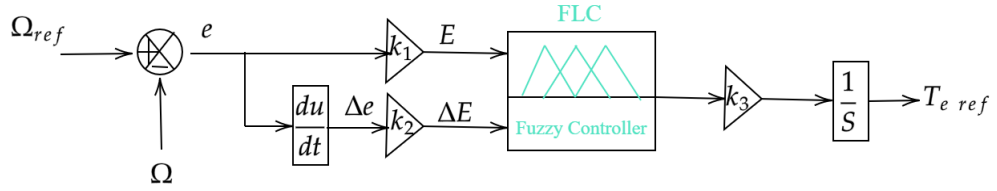


Figure 3.7: Structure of the F-PI.

The two inputs of the fuzzy regulator are the speed error noted e and its variation Δe , they are defined by the following equations:

$$e(k) = \Omega_{ref}(k) - \Omega(k) \quad (3.1)$$

$$\Delta e(k) = e(k) - e(k-1) \quad (3.2)$$

The output of the regulator corresponds to the reference of the electromagnetic couple noted $T_{e_{ref}}$.

The MFs into the input and output variables are trapezoidal and triangular MFs. The 7-segment MFs are: Negative Grand (NG), Negative Medium (NM), Negative Small (NS), Zero (Z), Positive Small (PS), Positive Medium (PM), and Positive Grand (PB), as shown in figures (3.8, 3.9).

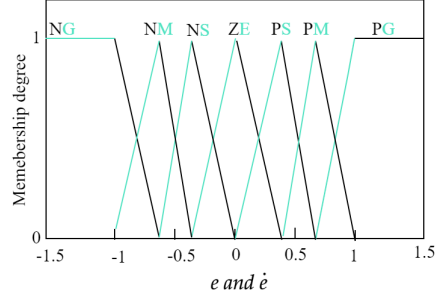


Figure 3.8: Input MFs.

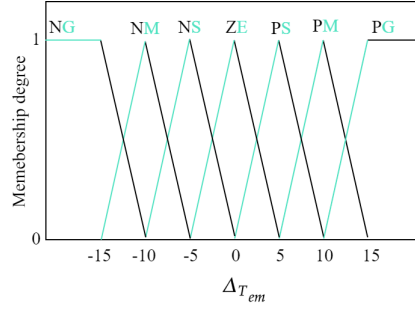


Figure 3.9: Output MFs.

When the inputs e (error) and Δe (error derivative) are transformed into linguistic variables, the FLC gives a suitable change in the output control signal for the DSIM, based on the rules shown in table 3.2.

Table 3.2: Rules generated for the PI-FLC

ΔT_e	$\dot{e}$						
	NG	NM	NS	ZE	PS	PM	PG
NG	NG	NG	NG	NG	NM	NS	ZE
NM	NG	NG	NG	NM	NS	ZE	PS
NS	NG	NG	NM	NS	ZE	PS	PM
ZE	NG	NM	NS	ZE	PS	PM	PG
PS	NM	NS	ZE	PS	PM	PG	PG
PM	NS	ZE	PS	PM	PG	PG	PG
PG	ZE	PS	PM	PG	PG	PG	PG

3.3.5 Estimator for compensation stator resistance variations

In this part, an estimator has been developed to estimate the machine's stator resistance variation, especially in low-speed operation. Each variation of R_s will be compensated by

a PI estimator. The scheme of this compensator is shown in figure 3.10. A low-pass filter is used to eliminate the high frequencies from the difference between the reference flux and the current flux. The output of the PI regulator is the variation of the stator resistance ΔR_s , this variation is constantly added to the stator resistance estimated previously, $\hat{R}_{s-1}$ to obtain the estimated stator resistance, $\hat{R}_s$.

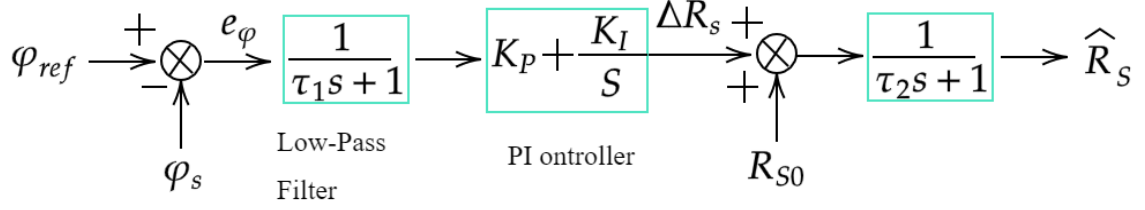


Figure 3.10: PI estimator for stator resistance

The proposed F-DTC control with stator resistance estimator aims to improve the dynamic performances (high ripples in the flux and torque and the variation of the stator resistance) while combining them with the benefits of the matrix converter that we mentioned previously. It is based on the use of a FIS instead of the hysteresis controllers and the switching table. In addition, the F-PI replaces the classical PI. The proposed scheme is shown in figure 3.11.

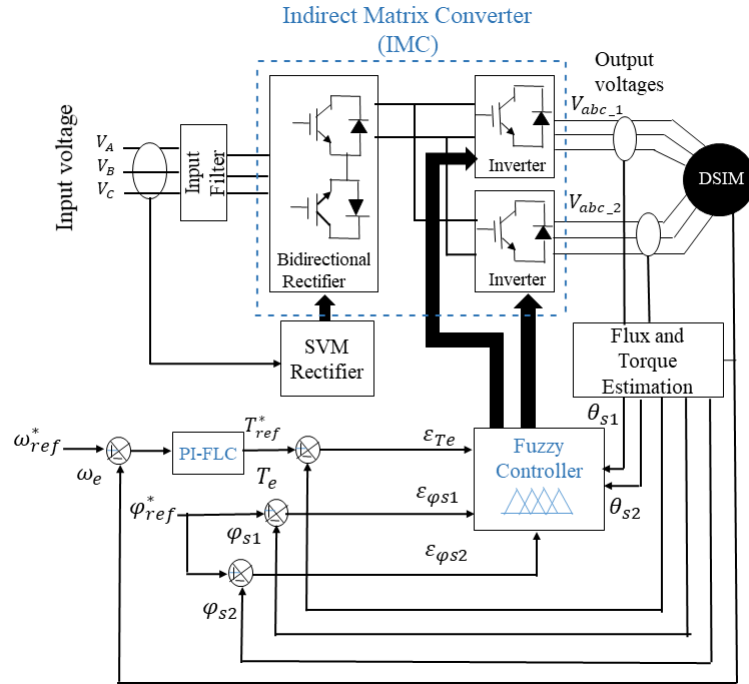


Figure 3.11: The proposed DTC based on a fuzzy inference system.

3.4 Numerical simulation

In order to demonstrate the advantages provided by the suggested intelligent DTC, several simulation tests have been carried out for several operating conditions.

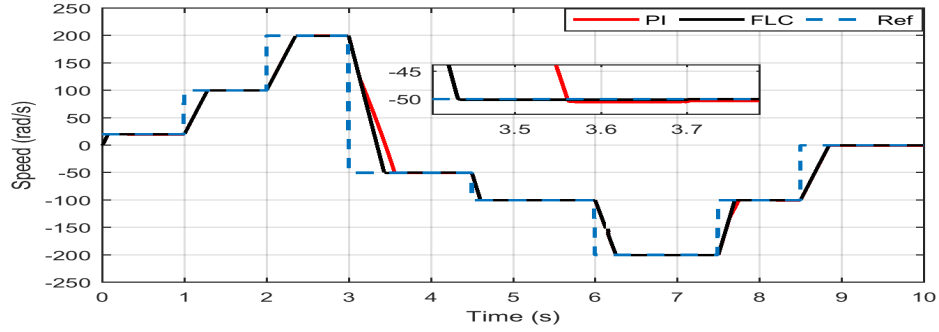


Figure 3.12: Speed of DSIM.

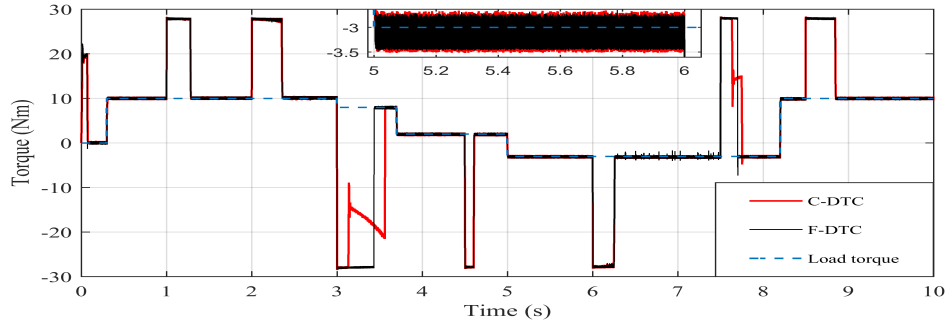


Figure 3.13: Electromagnetic torque.

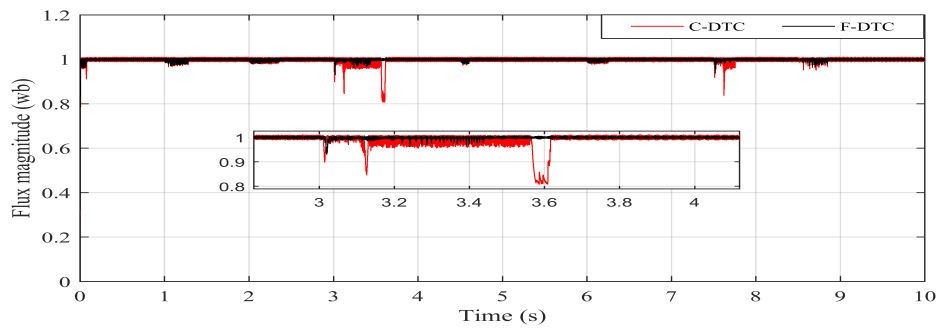


Figure 3.14: Stator flux magnitude.

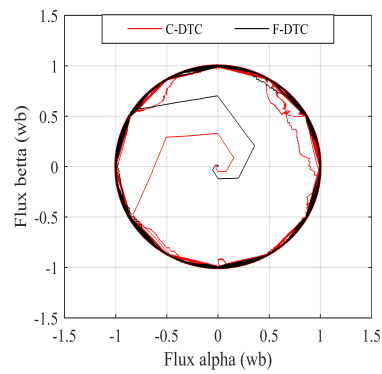


Figure 3.15: Stator flux circular trajectory .

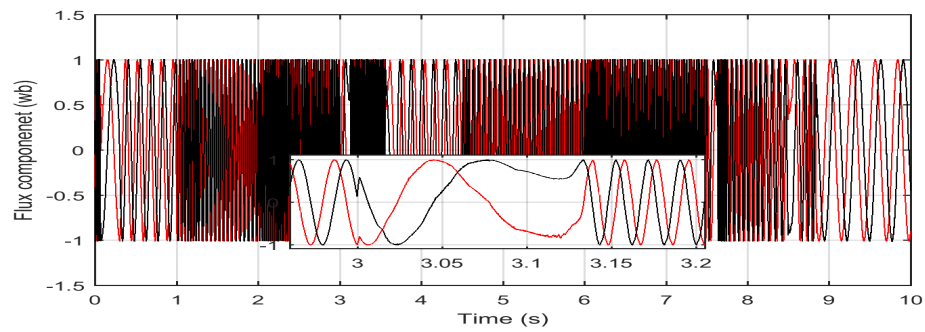


Figure 3.16: Stator flux axes components based on C-DTC.

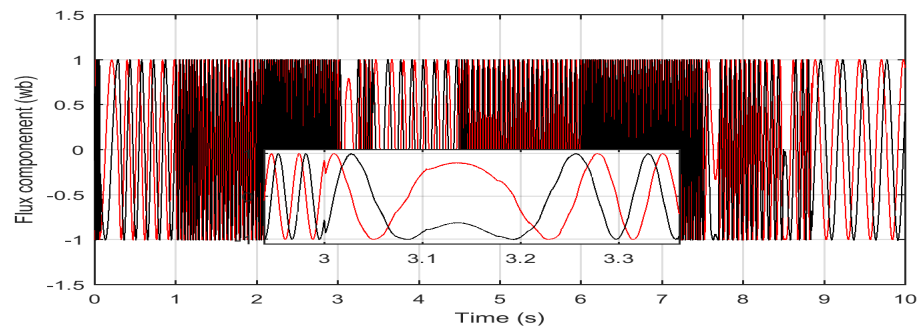


Figure 3.17: Stator flux axes components based on F-DTC.

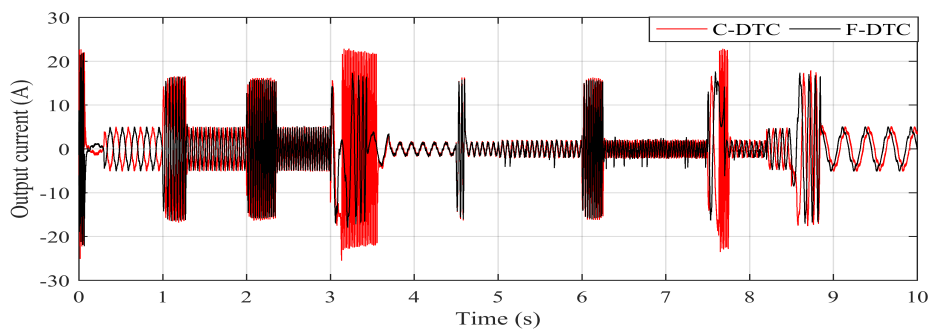


Figure 3.18: Output currents of DSIM with C-DTC and proposed F-DTC.

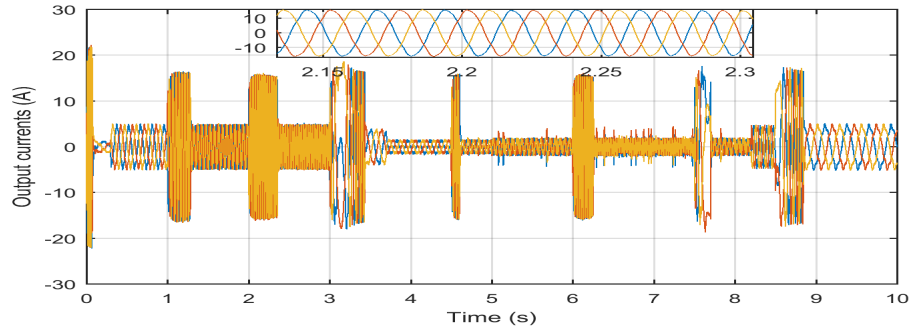


Figure 3.19: Output currents of DSIM based on F-DTC.

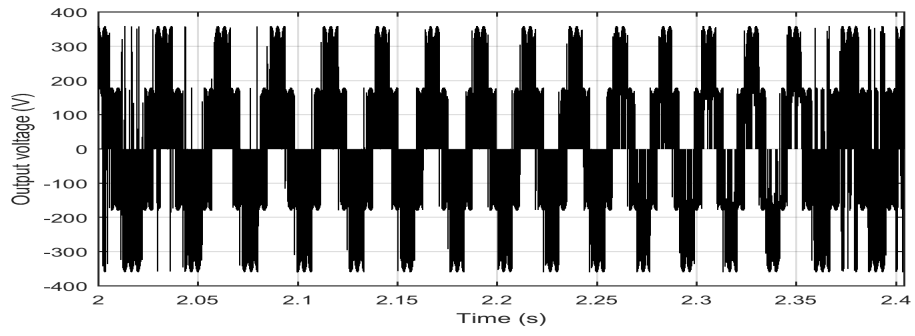


Figure 3.20: Output voltage of IMC with F-DTC.

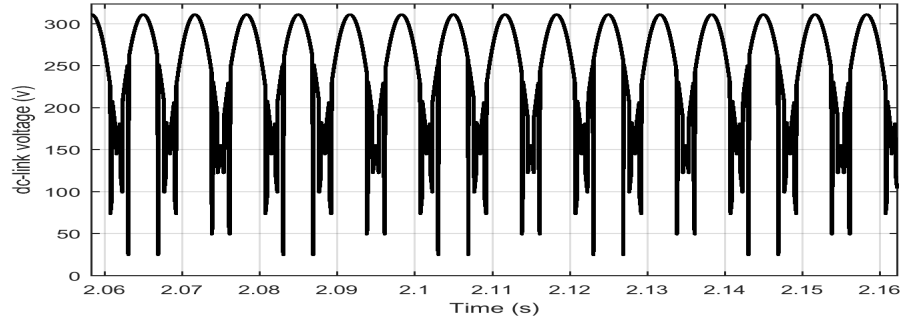
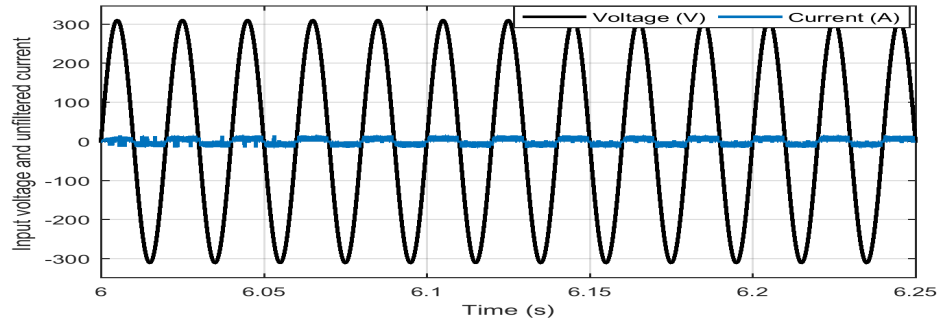
Figure 3.21: Output voltage of the rectifier V_{dc} with F-DTC.

Figure 3.22: Voltage and unfiltered current of the source with F-DTC.

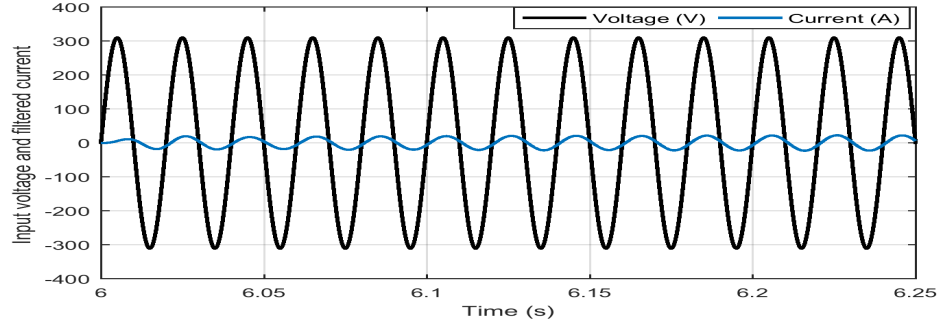


Figure 3.23: Voltage and filtered current of the source with F-DTC.

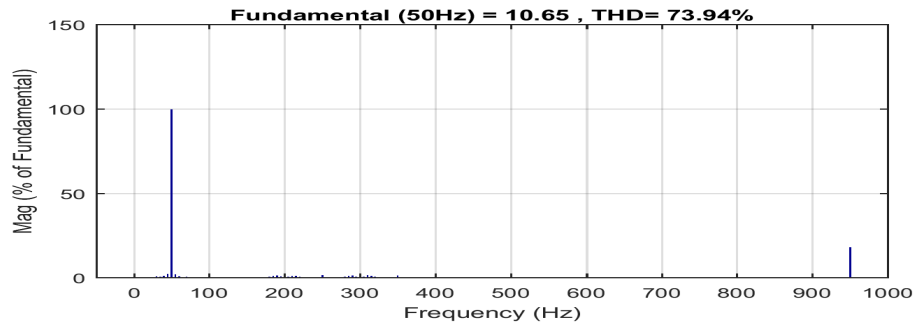


Figure 3.24: THD of input unfiltered current with F-DTC.

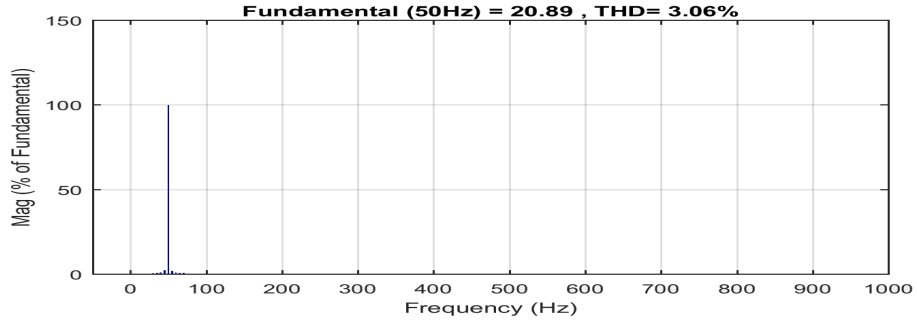


Figure 3.25: THD of input filtered current with F-DTC.

3.4.1 First Test: Response performance

To assess the robustness and the stability of the proposed controller in all possibilities of operation conditions of speed and torque, we can take reference values. The rotor speed tracks almost perfectly its reference (see figure 3.12), moreover, the reference is chosen to include many scenarios (constant reference, low, average and high speed, inversion and zero speed), and one can see that the fuzzy control has a faster response, very small tracking error and high robustness against external disturbances in all area compared

to the conventional PI. The torque follows well the variation of the load torque that is shown in figure 3.13, it also has a small tracking error for all the operating ranges studied. Furthermore, it is remarked that the dynamics of the torque in Fuzzy-DTC is faster and has much fewer ripples compared to the conventional DTC with hysteresis controller. Figures 3.14 to 3.17 respectively show the flux magnitude, the circular trajectory, and axes components, one can see that the use of fuzzy logic can greatly reduce the flux ripples and the flux follows its reference for the entire operating range. This reflects the good decoupling between the flux and torque axes that is achieved by the proposed control. The output currents are practically sinusoidal, they change their magnitude and frequency according to the speed reference and load torque. In addition, there is no distortion in the current waveform, its value is less important compared to the hysteresis-based DTC, this is remarkable especially when reversing the speed reference at 3 and 8.7 seconds (see Figure 3.18). Figure 3.19 shows the output currents of DSIM based on fuzzy-DTC. Figure 3.20 shows the output voltage of the IMC, it varies in amplitude and frequency according to the speed reference and the load torque variation. Figure 3.21 shows DC-link voltage (V_{dc}), generated by the modulation of the rectifier stage by the SVM technique. Due to the zero-current vector cancellation for maximizing the output V_{dc} , the DC-link voltage generated by the rectification stage does not consist of the zero voltage levels so the value of the DC-link voltage, is not constant, due to absence of the filtering capacity. The input voltage and both filtered and unfiltered currents are shown in figures 3.22-3.23. The input voltage and current are in phase, which means that the system operates under the unity power factor. This owes to the advantages of using the matrix converter. Figures 3.24, 3.25 show that the Fuzzy-DTC has low harmonics distortion (3.06%) compared to the hysteresis-based DTC shown in previous chapter (9.16%). The proposed control was able to reduce the ripples in the torque and stator flux and to minimize the harmonics in the input and output currents.

3.4.2 Robustness test: Parameters variation

This test has been made as simulation research to check the estimator's capacity with the variation of parameters (stator resistance). It depends on increasing the stator resistance R_S (50% of nominal value) when the machine works in the low-speed region (20, 30, and 40 rad/s). Figure 3.26 illustrates the estimation of the stator resistance, it tracks almost perfectly its reference, moreover, the reference is chosen to include many scenarios (increase, constant, and decrease). The speed follows well the references (figure 3.27) and

one can see that the fuzzy control has a faster response, very small tracking error, and high robustness against external disturbances in all area compared to the conventional PI during the large influence of the stator resistance. This shows the benefits of the proposed method against the variation of parameters. Figures 3.28-3.29, show the electromagnetic torque and flux magnitude respectively. It is observed that the dynamics of the torque in F- DTC is faster, stable, and has much fewer ripples, on the other hands the regulation of the stator flux for the C-DTC fails and the flux cannot be maintained to its rate of 1 Wb, and we can remark that the effect on flux magnitude in F-DTC was lower due to the optimization process.

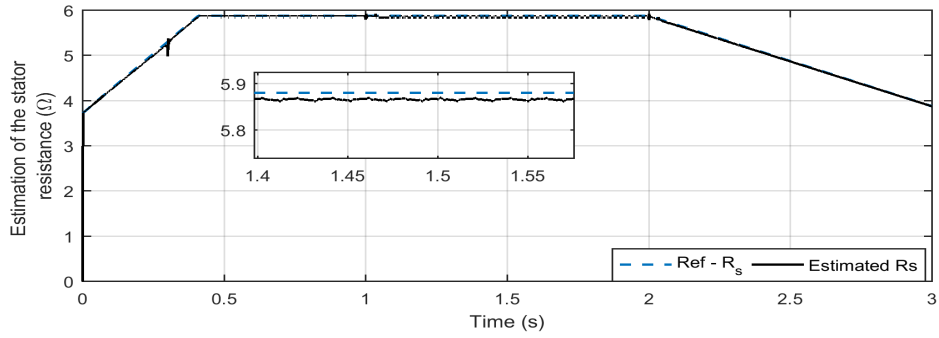


Figure 3.26: Estimated resistance.

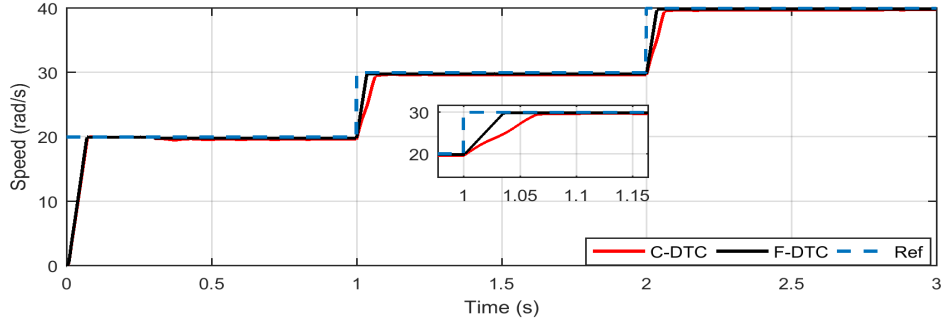


Figure 3.27: Speed of DSIM(During stator resistance variation).

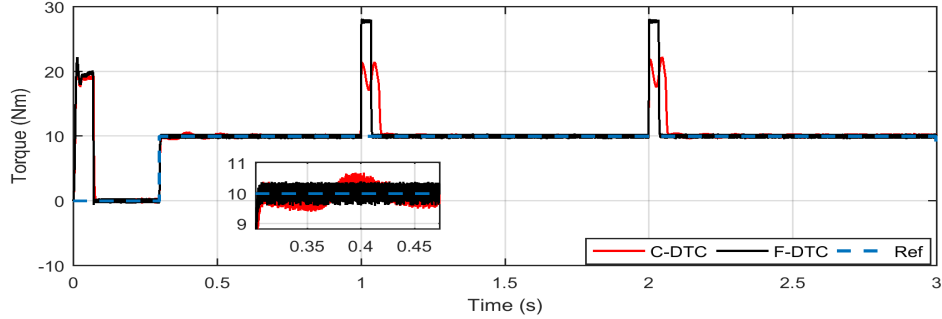


Figure 3.28: Electromagnetic torque(During stator resistance variation).

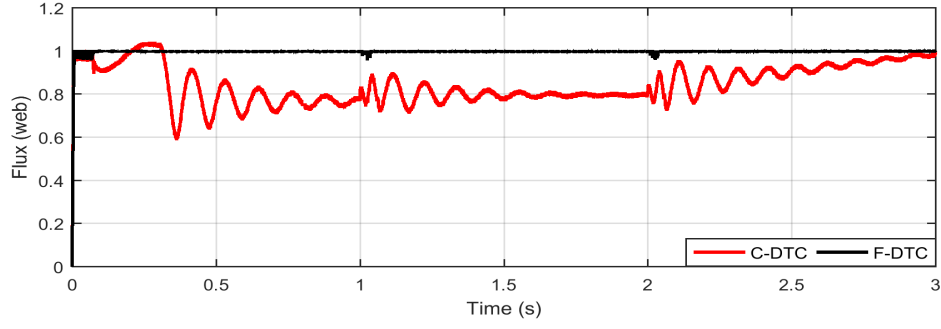


Figure 3.29: Stator flux magnitude.

3.5 Conclusion

In this chapter, we have proposed an improvement of DTC control based on a fuzzy inference system instead of the hysteresis controllers and switching table. To do this, we first have discussed the basics of fuzzy logic control. According to the results obtained, the main objectives have been achieved. It is clear that the proposed strategy gives good results compared to the classical control, as perfect tracking of references, reduction of torque and flux ripples, as well as an improvement in the quality of stator current signals. Indeed, the major interest of fuzzy logic in control is the capacity to control a system without depending on its parameters. Its effectiveness is proven when modeling a system is difficult, which is the case with our system. However, replacing the conventional hysteresis controllers and switching table by fuzzy controller can reduce the torque and flux ripples, but the ripples can be still large, and the low harmonic distortion appears in the input current caused by the lack of synchronization between the two stages of the IMC because the rectifier stage is controlled by SVM which leads to a constant switching frequency, where the inverter stage by a fuzzy logic controller which leads to a variable switching frequency, and on the other hand, it requires powerful microcontrollers in order to rapidly execute the control algorithms. In order to overcome this problem with improved dynamic performances (constant switching frequency, and reducing the low harmonic distortion in the input current), a new DTC based on Space Vector Modulation (SVM) with the synchronization between the two-stage of IMC will be proposed in the next chapter.

Chapter 4

Adaptive Optimized DTC-SVM using Metaheuristic Bat algorithm for DSIM fed by IMC Based on Robust Synergetic Speed Controller

“ A hopeless man sees difficulties in every chance, but a hopeful person
sees chances in every difficulty. ”

Ali ibn Abi-Talib

4.1 Introduction

After the study of direct torque control based on direct and indirect matrix converter and the fuzzy logic control in the previous chapters, we can remark that the ripples can be still large, and the low harmonic distortion appears in the input current caused by the lack of synchronization between the two stages of the IMC. In order to overcome this problem, a new DTC based on Space Vector Modulation (SVM) with the synchronization between the two stages of IMC is proposed. On the other hand, an adaptive PI controller based on a gradient descent algorithm using Metaheuristic Bat Algorithm (MBA) is presented to generate the desired stator voltages and the switching states of the inverter stage by the SVM. Furthermore, the control scheme performance is enhanced by inserting a robust

Synergetic Controller (SC) in the outer loop for speed regulation. The obtained simulation results under different operating conditions illustrate the benefit of the proposed control technique, also, it demonstrates the feasibility of the proposed control approach for real-world systems.

4.2 Proposed DTC-SVM

The proposed DTC-SVM technique is based on the use of adaptive regulators PI optimized by metaheuristic BA to adjust the adaptive rates for the flux and torque errors. The output of these regulators is the reference voltages which will be generated with SVM of the inverter stage, where also the rectifier based on the SVM and the synchronization between them has been done.

4.2.1 Rectifier control

For the rectifier stage, one applies the aforementioned SVM in order to maximize the output voltage V_{dc} , control the input power factor, and fixed the switching frequency as mentioned previously.

4.2.2 Inverter control

Like the rectifier stage, the inverter stage also is controlled by the SVM technic for controlling the output voltage and to ensure the synchronization between them (rectifier and inverter). The two-level inverter is a three-phase voltage inverter. It consists of three arms, each of which is made up of two bidirectional power switches operating in a complementary manner to avoid short-circuiting of the DC bus and ensuring continuity of the load current. The condition described by the equation (4.1) must be satisfied. Therefore, only one switch per arm must be passing.

$$S_{jp} + S_{jn} = 1 \quad (4.1)$$

where S_{jp} , S_{jn} are the connection functions of a single arm.

With respecting the relation (4.1), the inverter stage can have the eight states possible which are presented in the table 4.1. The input and output quantities are linked together by the following conversion matrix.

Figure 4.1 represents the reference vector V_{ref} (V_S) in the plane according to its ampli-

Table 4.1: Possible switching states for the inverter

Vectors	States	V_a	V_b	V_c	$ V_S $	θ_o	I_{rec}	
Active	V_1	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 1 \end{bmatrix}$	$\frac{2}{3}U_{rec}$	$-\frac{1}{3}U_{rec}$	$-\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	0	I_A
	V_2	$\begin{bmatrix} 1 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$	$\frac{1}{3}U_{rec}$	$\frac{1}{3}U_{rec}$	$-\frac{2}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$\frac{\pi}{3}$	$-I_C$
	I_3	$\begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 1 \end{bmatrix}$	$-\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$\frac{2\pi}{3}$	I_B
	I_4	$\begin{bmatrix} 0 & 1 & 1 \\ 1 & 0 & 0 \end{bmatrix}$	$-\frac{2}{3}U_{rec}$	$\frac{1}{3}U_{rec}$	$\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	π	$-I_A$
	I_5	$\begin{bmatrix} 0 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix}$	$-\frac{1}{3}U_{rec}$	$-\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$-\frac{2\pi}{3}$	I_C
	I_6	$\begin{bmatrix} 1 & 0 & 1 \\ 0 & 1 & 0 \end{bmatrix}$	$\frac{1}{3}U_{rec}$	$-\frac{2}{3}U_{rec}$	$\frac{1}{3}U_{rec}$	$\frac{2}{3}U_{rec}$	$-\frac{\pi}{3}$	$-I_B$
Zeros	I_0	$\begin{bmatrix} 0 & 0 & 0 \\ 1 & 1 & 1 \\ 1 & 1 & 1 \\ 0 & 0 & 0 \end{bmatrix}$	0	0	0	0	0	0

tude and phase, as well as the different variants that it can take according to the states of the inverter, moreover, its projection on the two adjacent vectors is presented.

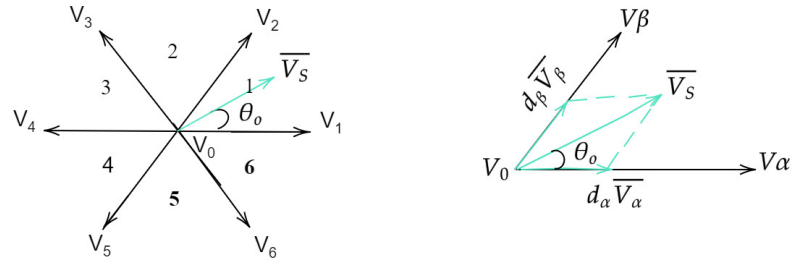


Figure 4.1: Output voltage modulation for inverter stage.

At each instant, V_S is located between two adjacent vectors V_α and V_β with an opening

equal to $\frac{\pi}{3}$, from which and from the zero vector located at the origin of the hexagon, the reference vector can be synthesized, based on the SVM, according to the following equation:

$$V_S = d_\alpha V_\alpha + d_\beta V_\beta + d_0 V_0 \quad (4.2)$$

where d_α, d_β and d_0 represent the duty cycles. They are defined as follows:

$$\begin{cases} d_\alpha = m_v \cdot \sin(\frac{\pi}{3} - \theta_0) \\ d_\beta = m_v \cdot \sin(\theta_0) \\ d_0 = 1 - d_\alpha - d_\beta \end{cases} \quad (4.3)$$

where m_v is the voltage ratio of the inverter and θ is the angle of V_s compared to the sector.

As known, the U_{pn} is not constant because of the absence of the filtering capacity and to avoid the problem of low frequency harmonics in the input currents, the variation of U_{pn} must be taken into account in the calculation of the ratio of the inverter, so m_v is given by the following equation:

$$m_v = \frac{2}{\sqrt{3}} r \frac{\cos(\frac{\pi}{6} - \theta_{in})}{\cos(\Psi_E)} \quad (4.4)$$

where r is the ratio of modulation is written by: $r = \frac{|V_S|}{|V_{in}|}$, Ψ_E is the input phase displacement and θ_{in} is the angle of the input current vector relative to the sector.

4.2.3 Synchronization between the two stages of IMC

In order to obtain a correct order of the input currents and output voltages, and in order not to have low frequency harmonics, the synchronization between the two stages is essential. The modulation combines the switching states of the rectification stage and that of the inversion uniformly, during each sampling period T_s .

Figure 4.2 shows how to choose this combination which represents the synchronization between the two stages which is generally done in the inverter stage. In this example, the current and voltage reference vectors are located in sector 1 for each of the two stages. The redundant switching states for V_0 are used to reduce the number of switching operations per sampling period. Then, in order to minimize the switching frequency, the inverter vectors are used with two-way switching, the two halves are unequal, their durations represent the application durations of the vectors I_γ, I_δ .

The vector application times are given by the following expressions:

$$\begin{cases} t_{r1} = d_\gamma T_S \\ t_{i1} = \frac{1}{2}d_0 d_\gamma T_S \\ t_{i2} = d_\gamma d_\beta T_S \\ t_{i3} = d_\gamma d_\alpha T_S \end{cases} \begin{cases} t_{i4} = \frac{1}{2}d_0 d_\delta T_S \\ t_{i5} = d_\delta d_\alpha T_S \\ t_{i6} = d_\delta d_\beta T_S \end{cases} \quad (4.5)$$

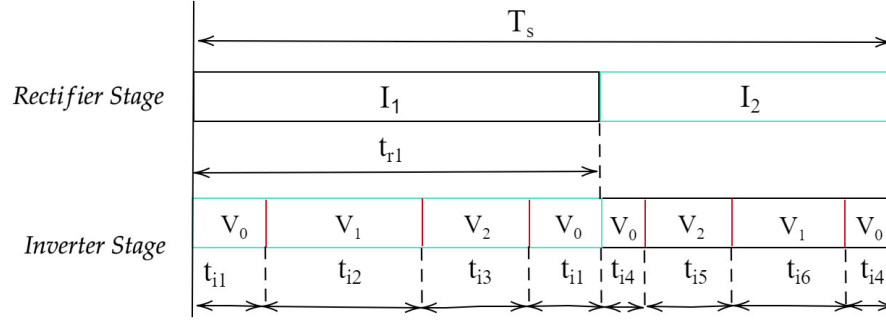


Figure 4.2: The synchronization between the two stages of IMC.

4.2.4 Switching states

Two-level inverter switching states

The switching states of the upper half arms of the two-level inverter are shown below (the states of the lower half arms are complementary).

	T_s							
	$T_s/2$				$T_s/2$			
	$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$	$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$
S_{11}								
S_{12}								
S_{13}								

Figure 4.3: Inverter switching states for sector 1

T_s							
$T_s/2$				$T_s/2$			
$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$	$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$
S_{11}							
S_{12}							
S_{13}							

Figure 4.4: Inverter switching states for sector 2

T_s							
$T_s/2$				$T_s/2$			
$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$	$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$
S_{11}							
S_{12}							
S_{13}							

Figure 4.5: Inverter switching states for sector 3

T_s							
$T_s/2$				$T_s/2$			
$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$	$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$
S_{11}							
S_{12}							
S_{13}							

Figure 4.6: Inverter switching states for sector 4

T_s							
$T_s/2$				$T_s/2$			
$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$	$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$
S_{11}							
S_{12}							
S_{13}							

Figure 4.7: Inverter switching states for sector 5

	T_s							
	$T_s/2$				$T_s/2$			
	$T_0/4$	$T_\alpha/2$	$T_\beta/2$	$T_0/4$	$T_0/4$	$T_\beta/2$	$T_\alpha/2$	$T_0/4$
S_{11}								
S_{12}								
S_{13}								

Figure 4.8: Inverter switching states for sector 6

Note that at each switching, only two switches are involved, moreover, each switch of the inverter turns on and off only once per sampling period.

Switching states of IMC

Figure 4.9 illustrates an example of switching the switches in sector 1 of the inverter while respecting the synchronization between the two stages.

	T_s							
	d_γ				d_σ			
	t_{i1}	t_{i2}	t_{i3}	t_{i1}	t_{i4}	t_{i5}	t_{i6}	t_{i4}
S_{11}								
S_{12}								
S_{13}								
S_{21}								
S_{22}								
S_{23}								

Figure 4.9: Switch states for sector 1

4.3 Design of adaptive controllers

In this part, the steps of the design of adaptive PI controller based on the gradient descent algorithm has been developed to control the torque and flux of the machine during operation. The output of these regulators is the reference voltages ($V_{d_{s1,2}}, V_{qs}$) which will be generated by SVM of the inverter stage.

The general expression of the discret PI controller that is commonly in the literature is

written as [95, 96]:

$$u(k) = K_P e(k) + K_I \sum_{j=1}^k e(j) \quad (4.6)$$

The digital incremental PI algorithm is shown as follow:

$$u(k) = u(k-1) + K_P(e(k) - e(k-1)) + K_I e(k) \quad (4.7)$$

where K_P , K_I , $e(k)$, $r(k)$ and $y(k)$ are the proportional and integral differential coefficients, the system error which is equal to $e(k) = r(k) - y(k)$, desired value (T_{ref}, φ_{ref}) and the estimated value (T_e, φ_e) respectively.

The general structure of the proposed adaptive PI controller optimized by BA is shown in figure 4.10.

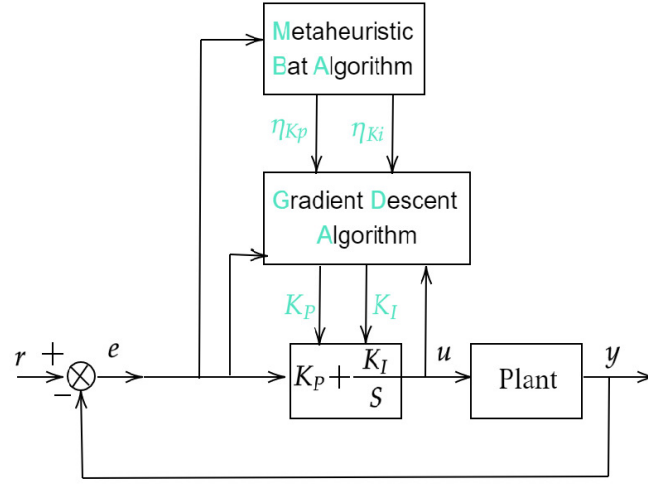


Figure 4.10: Proposed adaptive PI control system.

Then, to describe the adapting parameters (K_P, K_I) , the objective function is defined as:

$$E_k = \frac{1}{2}(r(k) - y(k))^2 = \frac{1}{2}e(k)^2 \quad (4.8)$$

The updated parameters of the PI regulator based on the gradient descent algorithm are given by the following equations:

$$K_P(k+1) = K_P(k) - \eta \frac{\partial E}{\partial K_P} \quad (4.9)$$

$$K_I(k+1) = K_I(k) - \eta \frac{\partial E}{\partial K_I} \quad (4.10)$$

where η is the adaptive rate.

The derivation of the premise part in (4.9)–(4.10) can be expressed by the following:

$$\Delta K_P(k) = \eta \frac{\partial E(k)}{\partial K_P(k)} = \eta \frac{\partial E(k)}{\partial y(k)} \frac{\partial y(k)}{\partial u(k)} \frac{\partial u(k)}{\partial K_P(k)} \quad (4.11)$$

Furthermore, the other derivative terms are determined as:

$$\frac{\partial E(k)}{\partial y(k)} = -(r(k) - y(k)) = -e(k) \quad (4.12)$$

Then

$$\frac{\partial y(k)}{\partial u(k)} \approx \frac{\Delta y(k)}{\Delta u(k)} = \frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \quad (4.13)$$

$$\frac{\partial u(k)}{\partial K_P(k)} = (e(k) - e(k-1)) \quad (4.14)$$

Thus, when replacing the equations (4.12 - 4.14) in (4.11), the variable parameter of K_P is written as:

$$\Delta K_P(k) = \eta \frac{\partial E(k)}{\partial K_P(k)} = -\eta e(k) \left[\frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \right] (e(k) - e(k-1)) \quad (4.15)$$

For the variable parameter K_I :

$$\Delta K_I(k) = \eta \frac{\partial E(k)}{\partial K_I(k)} = \eta \frac{\partial E(k)}{\partial y(k)} \frac{\partial y(k)}{\partial u(k)} \frac{\partial u(k)}{\partial K_I(k)} \quad (4.16)$$

Also from (4.6, 4.7), the latest derivative terms are expressed by :

$$\frac{\partial E(k)}{\partial y(k)} = -e(k) \quad (4.17)$$

$$\frac{\partial y(k)}{\partial u(k)} \approx \frac{\Delta y(k)}{\Delta u(k)} = \frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \quad (4.18)$$

$$\frac{\partial u(k)}{\partial K_I(k)} = e(k) \quad (4.19)$$

Thereafter, when replacing the equations (4.17 - 4.19) in (4.16), the variable parameter of K_I is expressed as:

$$\Delta K_I(k) = \eta \frac{\partial E(k)}{\partial K_I(k)} = -\eta e(k) \left[\frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \right] e(k) \quad (4.20)$$

Finally, the adaptive parameters of the PI regulator by the gradient descent algorithm are written by the following expressions:

$$K_P(k+1) = K_P(k) - \eta e(k) \left[\frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \right] (e(k) - e(k-1)) \quad (4.21)$$

$$K_I(k+1) = K_I(k) - \eta e(k) \left[\frac{y(k-1) - y(k-2)}{u(k-1) - u(k-2)} \right] e(k) \quad (4.22)$$

In order to optimize the adaptive rates η , a metaheuristic Bat algorithm is proposed to adjust it, this method is illustrated in the next section.

4.3.1 Metaheuristic Bat Algorithm

A metaheuristic algorithm, called the Bat Algorithm (BA), was developed by Xin She Yang [97, 98], and it is inspired by the echolocation behavior of microbats, with varying beat rates of emission and loudness.

One can idealize the following rules for the echolocation characteristics of microbats :

- 1- All bats use echolocation to sense distance, moreover, they distinguish between prey/food and foundation obstructions in a couple of supernatural manners.
- 2- While searching for prey, bats fly randomly with a velocity ν_j at a certain position x_j with an unchangeable frequency f_{min} , varying loudness A_0 and wavelength λ . They can routinely tune the rate of beat emission $r \in [0, 1]$ and alter the wavelength of their radiated beats, depending on the proximity of their goal.
- 3- one supposes that the loudness varies from a higher positive constant value A_0 to a lower constant value A_{min} .

The new solutions velocities ν_j^t and x_j^t at a time step t , after identification of the velocity ν_j and position x_j that corresponds to each (j) , is expressed as follows:

$$f_j = f_{min} + (f_{max} - f_{min})\alpha \quad (4.23)$$

$$\nu_j^t = \nu_j^{t-1} + (x_j^{t-1} - x^*) f_j \quad (4.24)$$

$$x_j^t = x_j^{t-1} + \nu_j^t \quad (4.25)$$

where α is in the range of $[0, 1]$, and x^* stands for the current global best location. As the value of $f_i \cdot \lambda_i$ increases, one can use f_i (or λ_i) to tune the variation of the velocity. For the implementation, each bat is randomly assigned a randomly drawn frequency from (f_{min}, f_{max}) . The local search is a random walk around the current best solutions, and the new solution for each bat can be generated locally by:

$$x_{new} = x_{old} + \varepsilon A^t \quad (4.26)$$

where $\varepsilon \in [0, 1]$ is a random number, while $A^t = \prec A_j^t \succ$ represents the average loudness of all the bats at this time step. As the loudness generally reduces once a bat has found its prey, the rate of pulse emission increases. The loudness is typically chosen from $[A^0, A_{min}] = [1, 0]$. Supposing that the value of A_{min} explains that a bat has just found the prey, and the moment stop emitting any noise. The rate of beat emission and the loudness are given by:

$$r_j^{t+1} = r_j^0 [1 - \exp(-\gamma t)], \quad A_j^{t+1} = \beta A_j^t \quad (4.27)$$

where β and γ are constant. β is like the cooling factor of a cooling schedule. For any $\gamma \succ 0$ and $0 \prec \beta \prec 1$

$$A_j^t \rightarrow 0, r_j^t \rightarrow r_j^0; \text{ast} \rightarrow \infty \quad (4.28)$$

where in general states can be selected $\beta = \gamma = [0.9 - 0.975]$.

The pseudo code of Bat algorithm summarized in figure 4.11. The parameters used for this simulation are listed in Table 4.2.

Table 4.2: Parameters of BA

Algorithm	Parameters
BA	$\beta = \gamma = 0.9;$
	$f_{min} = 0, f_{max} = 1;$
	$A_0 \in [0, 1];$
	$r \in [0, 1];$
	$N=18, Lb=0.1, Ub=50.$

4.4 Synergetic Control

The synergetic control is a control technique quite close to Sliding Mode Control (SMC) in the sense that it forces the system to evolve with a dynamic pre-chosen by the designer. This approach explicitly uses a nonlinear model for the synthesis of the command [99–101].

Suppose that the system to be ordered is described by a set of nonlinear equations of the following form:

$$\dot{x} = f(x, u, t) \quad (4.29)$$

where

$x = (x_1, x_2, \dots, x_n)$ is the state vector, $u = (u_1, u_2, \dots, u_m)$ the vector of input control, f a nonlinear function and continuous in time.

The first step in designing a synergetic control is the formation of macro variables defined in terms of system state variables. The macro variable is defined by:

$$\varphi(\Psi) = \Psi \quad (4.30)$$

The command will make the system to operate on the manifold $\Psi = 0$, the dynamic evolution of the macro-variables according to the equation is defined by:

$$T\dot{\Psi} + \Psi = 0 \quad T \succ 0 \quad (4.31)$$

Where the function φ must satisfy, the following conditions to ensure the stability of this functional equation: $\Psi(0) = 0$, $\varphi(\Psi)\Psi \succ 0$ for all $\Psi \neq 0$

The solution of the Eq. (4.31) is given by:

$$\Psi(t) = \Psi_0.e^{\frac{t}{T}} \quad (4.32)$$

The parameter T defines the speed of convergence of macro-variables to the intersection of manifolds $\Psi = 0$.

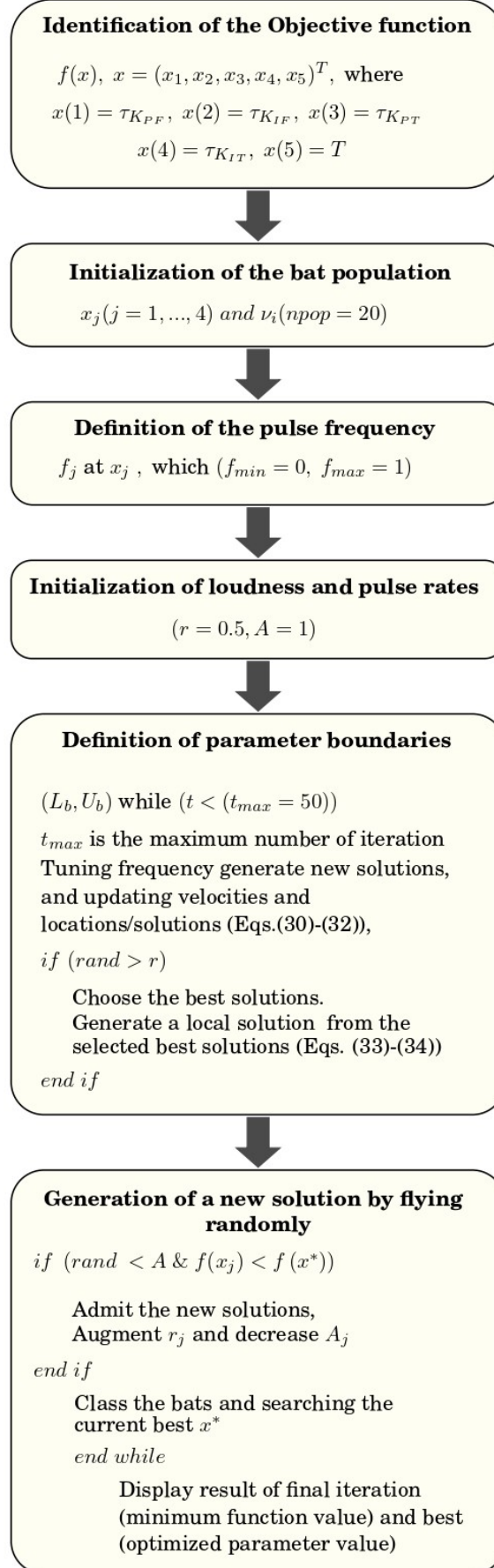


Figure 4.11: Flow chart of Bat algorithm.

4.4.1 Synergetic speed controller design

In DTC is well known used a classical PI controller in the external speed to make the reference torque. In this work, we proposed a new robust controller design basing on the synergetic control. The speed controller generates the reference torque T_{ref} . The speed error is defined by:

$$e(t) = \Omega_{ref} - \Omega_r = 0 \quad (4.33)$$

Consider the macro-variable given by:

$$\Psi = K_P e(t) + K_I \int e(t) \quad (4.34)$$

where K_p and K_I are the proportional and integral parameters of the speed macro-variable

$$\dot{\Psi} = K_P \dot{e}(t) + K_I e(t) \quad (4.35)$$

The dynamic equation of DSIM is given as:

$$\dot{\Omega} = \frac{1}{J} (T_r - T_{em} - f_r \Omega) \quad (4.36)$$

where, J : inertia moment, f_r : coefficient of friction, C_r : Load torque. By substituting Eq. (4.36) in the equation (4.31), and basing on the synergetic theory, we can write:

$$T(K_P(\dot{\omega}_{ref} - \dot{\omega}_r) + K_I(\omega_{ref} - \omega)) + K_P(\omega_{ref} - \omega_r) + K_I(\omega_{ref} - \omega_r) = 0 \quad (4.37)$$

$$T_{em}^* = T_r - f_r \Omega - J \left(\dot{\Omega}_{ref} + \frac{1}{K_P} \left(\frac{\Psi}{T} + K_I(\Omega_{ref} - \Omega) \right) \right) \quad (4.38)$$

The synergetic control law must ensure the stability of the closed loop speed control. One can use the following Lyapunov function:

$$V = \frac{1}{2} \Psi(e)^2 \quad (4.39)$$

after differentiation one gets:

$$\dot{V} = \Psi(e) \dot{\Psi}(e) \quad (4.40)$$

$$\dot{V} = -\frac{1}{T} \Psi^2(e) \leq 0 \quad (4.41)$$

Thus, the inequality (4.41) will ensure the stability of closed speed control loop.

The proposed adaptive DTC-SVM control based on the robust SC aims to improve the performances such as the high ripples in the flux and torque by fixed the switching frequency and to minimizes the THD in the input current via the synchronization between the two stages of the IMC. Figure 4.12. shows the overall diagram of the proposed method.

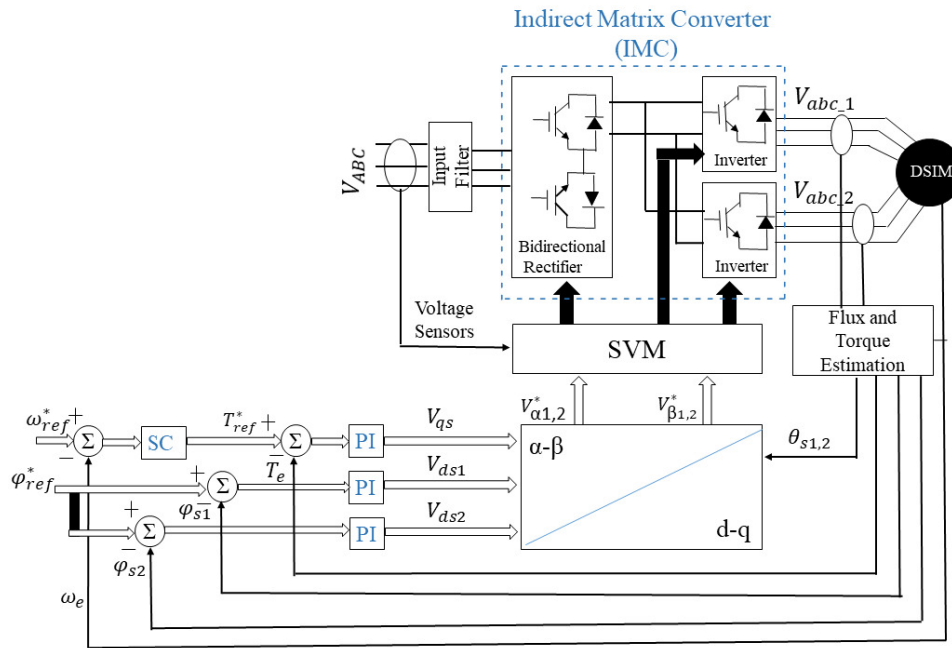


Figure 4.12: Proposed adaptive DTC-SVM based on synergetic speed controller.

4.5 Simulation results

To demonstrate the benefits provided by the proposed control, several simulation tests have been carried out for different operating conditions. The simulation is achieved using Matlab/Simulink software. The parameters utilized in this simulation are listed in the table 6.

To assess the robustness and the stability of the proposed controller in all possibilities of operation conditions of speed and torque, one can take different reference scenario. The rotor speed tracks almost perfectly its reference (see figure 4.13). The reference is chosen to include different scenarios (constant reference, low, average and high speed, inversion and zero speed), one can see that the SC has a faster response, very small tracking error and high robustness against external disturbances in all area compared to the conventional PI controller. The torque follows well the variation of the load torque (figure 4.14), it also has a small tracking error for all the operating ranges. Furthermore, it is remarked that the dynamics of the torque in proposed DTC is faster and has much fewer ripples compared to the conventional DTC with hysteresis controller. The aforementioned performances provided by the proposed controller demonstrate the applicability of the P-DTC to real-world applications. Figures 4.15 - 4.17 respectively show the circular trajectory and the flux magnitude. One can see that the use of P-DTC can greatly reduce flux ripples. The

flux follows its reference for the entire operating range. This reflects the good decoupling between the flux and torque axes that is achieved by the proposed control, this is very important when moving from simulation toward real application. The output currents are practically sinusoidal, with variable magnitude and frequency depend on the load torque and speed reference. In addition, there is no distortion in the current waveform, its value is less important compared to the C-DTC, this is remarkable especially when reversing the speed reference at 3 and 7.8 seconds (see figure 4.18). Figure 4.19 shows the output currents of DSIM based on P-DTC. Figures 4.20 shows the output voltage of the IMC, it varies in amplitude and frequency according to the variation of load torque and speed reference. The input voltage and both the unfiltered and the filtered currents for the P-DTC and the C-DTC are shown in figures (4.21, 4.23, 4.25 and 4.27). The input voltage and current are in phase, which means that the system operates under the unity power factor, This owes to the benefits of using the matrix converter. Figures (4.22, 4.24, 4.26 and 4.28) show the THD in the input current with the P-DTC, one can see that the P-DTC has low THD compared to the C-DTC. Figure 4.29 shows DC-link voltage (V_{dc}), generated by the modulation of the rectifier stage by the SVM technique. Due to the zero-current vector cancellation for maximizing the output V_{dc} , the DC-link voltage generated by the rectification stage does not consist of the zero voltage levels and the absence of the filtering capacity, so the value of the DC-link voltage is not constant. Figure 4.30 shows the input power factor of the source, one can see that the system is operating under the unity power factor in all the possibilities of operation conditions, this is the benefits of using the IMC. The history of the adaptation gains K_P and K_I by the gradient descent algorithm based on the metaheuristic bat algorithm for the flux and torque is shown in figures 4.31-4.32 respectively. The proposed control was able to reduce the ripples in the torque and flux and to minimize the harmonics distortion in the input and output currents, again this is very important when applying the P-DTC to a real-world system. The main drawback of the variable switching frequency in the C-DTC is solved by the proposed P-DTC. The optimal optimized adaptive rates based on the metaheuristic BA are given in table 4.2. Where the comparative study between the P-DTC and the C-DTC is presented in table 4.3.

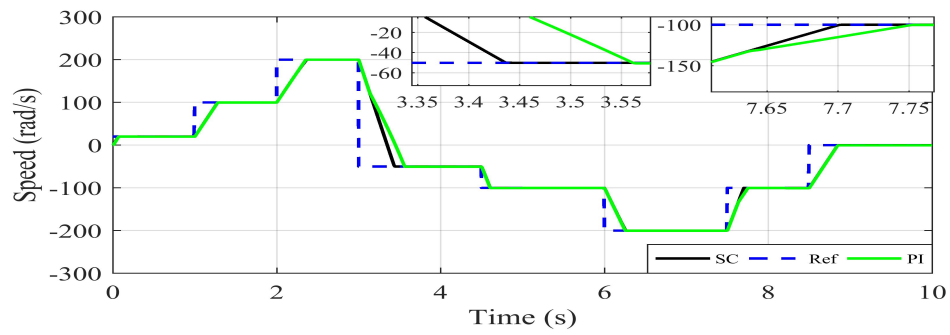


Figure 4.13: Speed of DSIM.

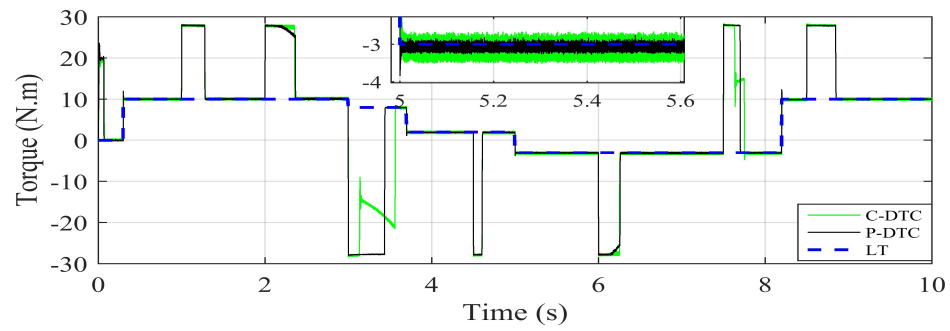


Figure 4.14: Electromagnetic torque.

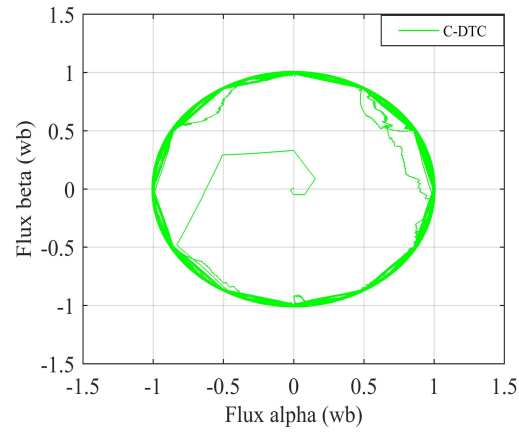


Figure 4.15: Stator flux circular trajectory with C-DTC

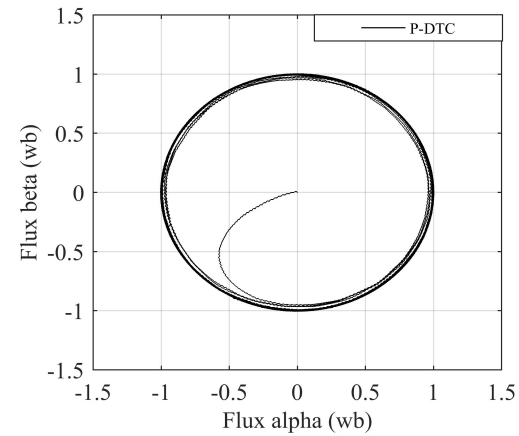


Figure 4.16: Stator flux circular trajectory with P-DTC

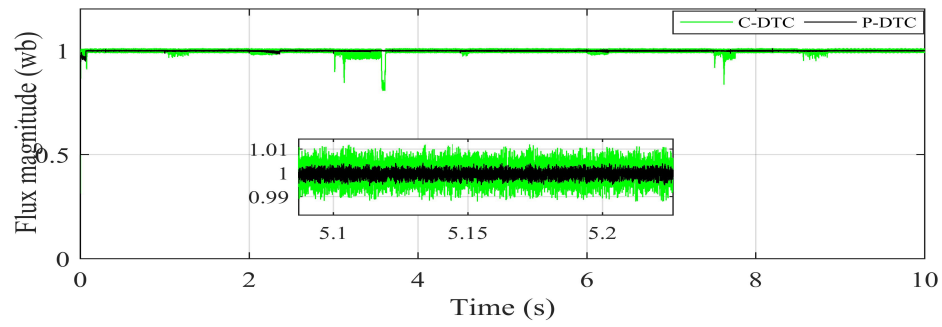


Figure 4.17: Stator flux magnitude.

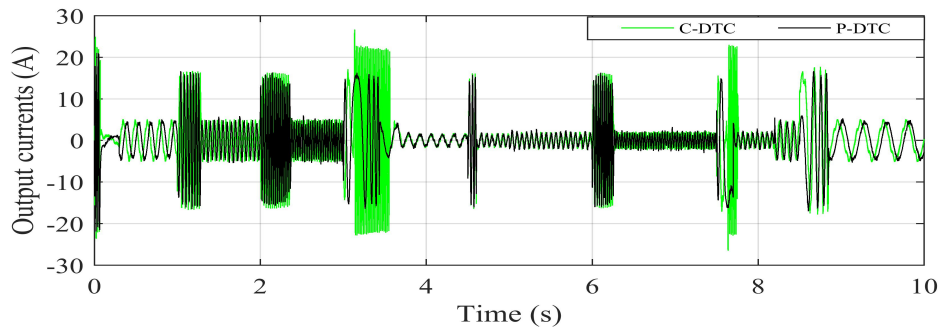


Figure 4.18: Output currents of DSIM with C-DTC and P-DTC.

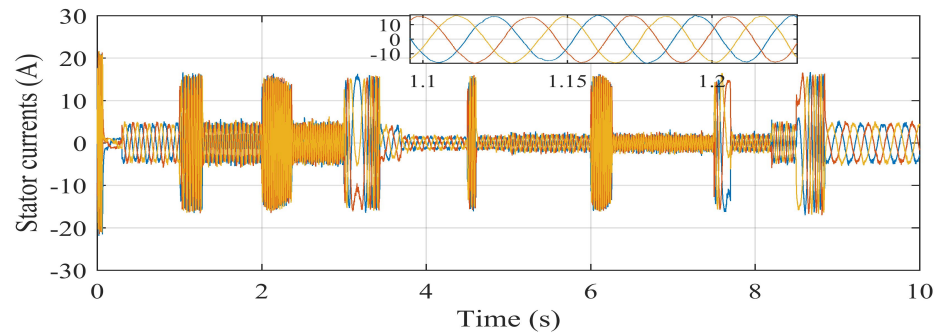


Figure 4.19: Output currents of DSIM with P-DTC.

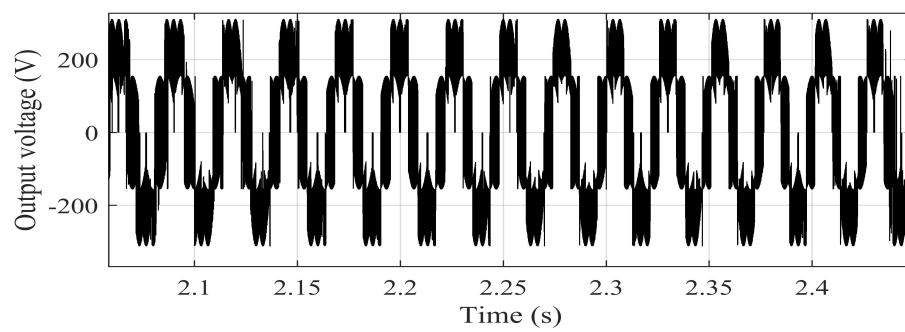


Figure 4.20: Output voltage of IMC.

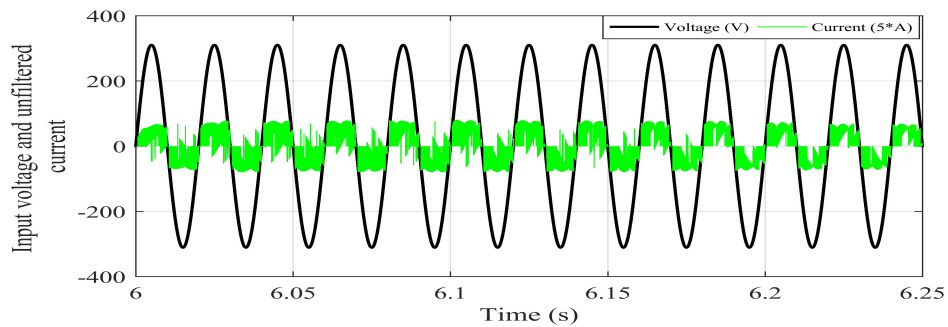


Figure 4.21: Voltage and unfiltered current of the source with P-DTC.

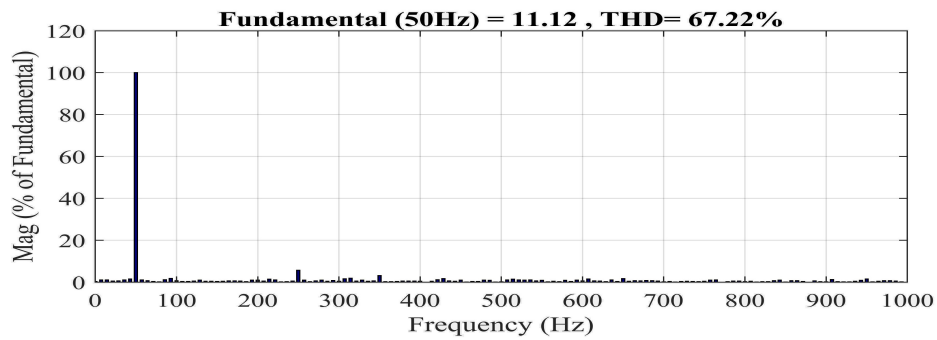


Figure 4.22: THD of input unfiltered current with P-DTC.

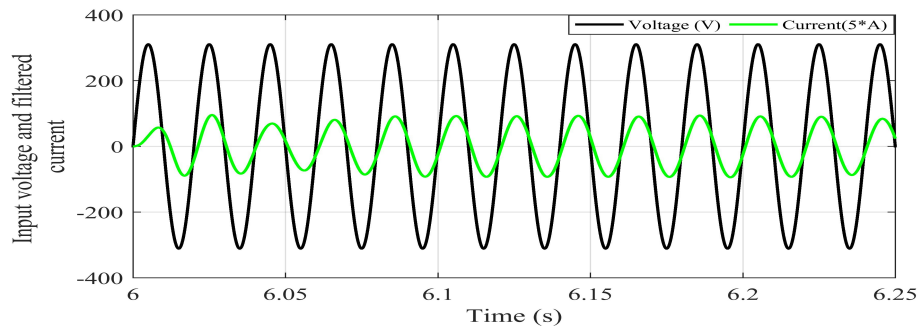


Figure 4.23: Voltage and filtered current of the source with P-DTC.

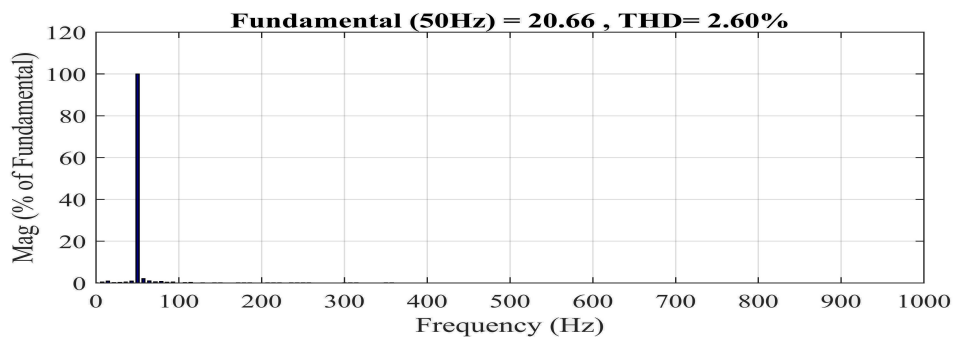


Figure 4.24: THD of input filtered current with P-DTC.

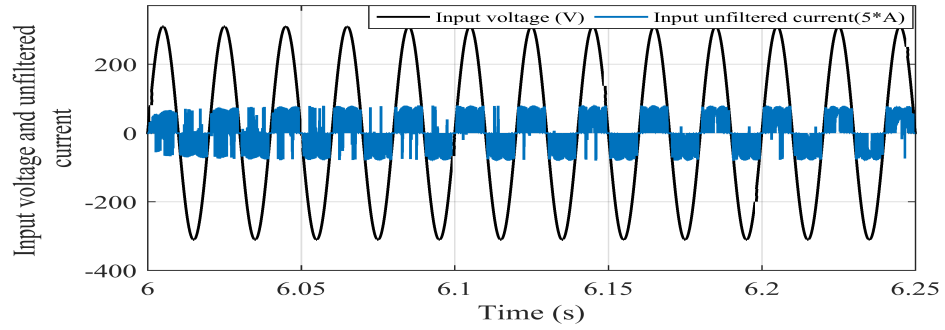


Figure 4.25: Voltage and unfiltered current of the source with C-DTC.

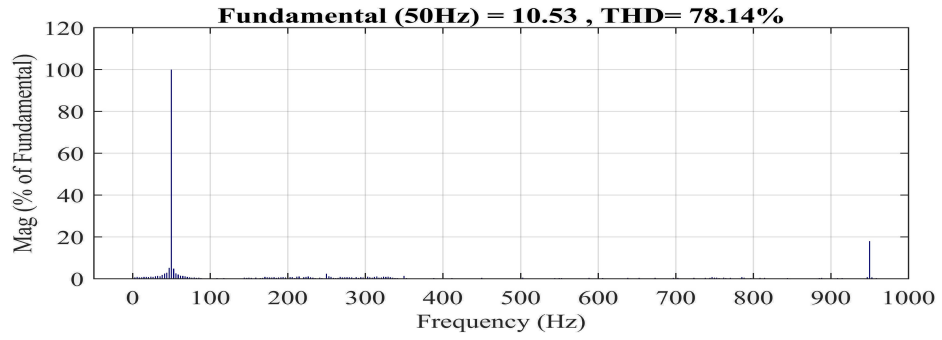


Figure 4.26: THD of input unfiltered current with C-DTC.

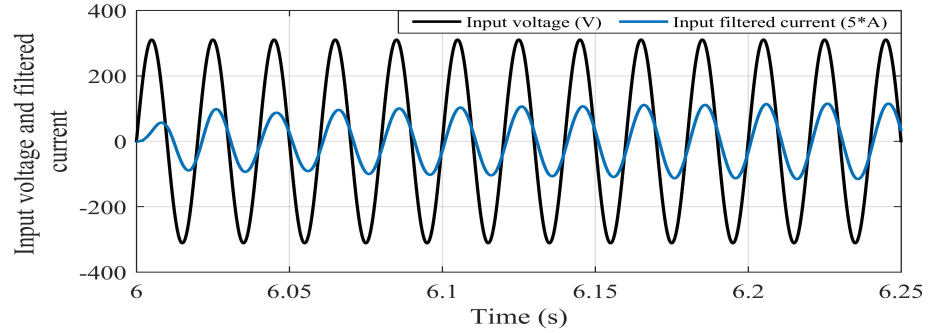


Figure 4.27: Voltage and filtered current of the source with C-DTC.

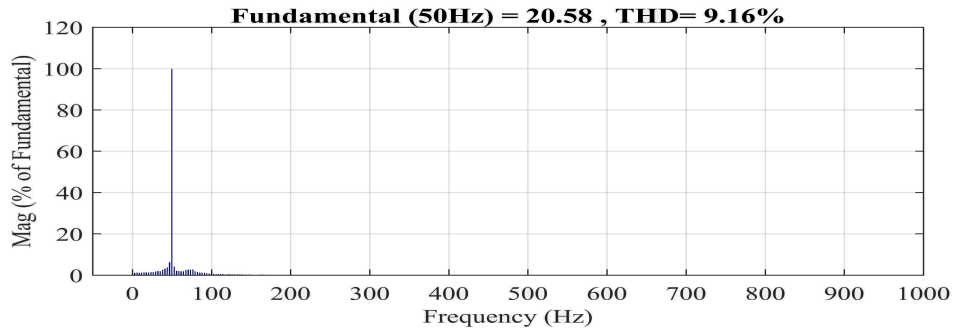


Figure 4.28: THD of input filtered current with C-DTC.

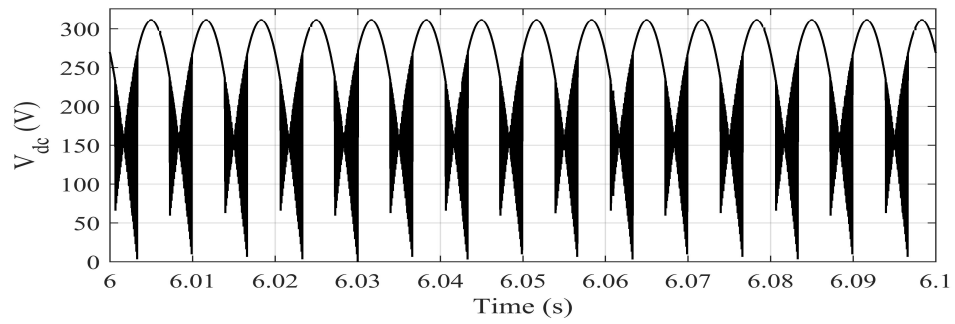
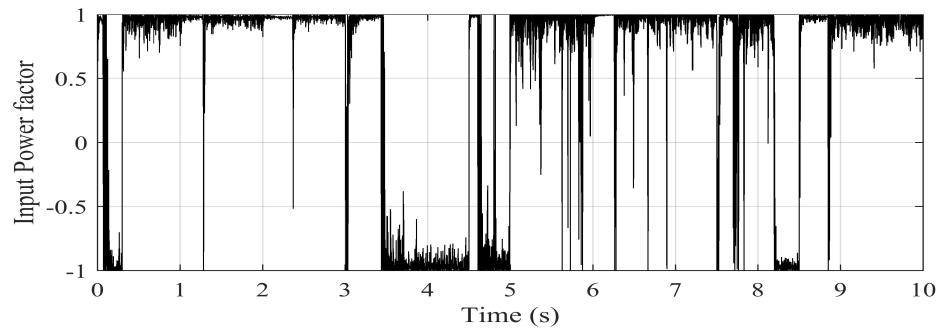
Figure 4.29: Output voltage of the rectifier V_{dc} .

Figure 4.30: Input power factor.

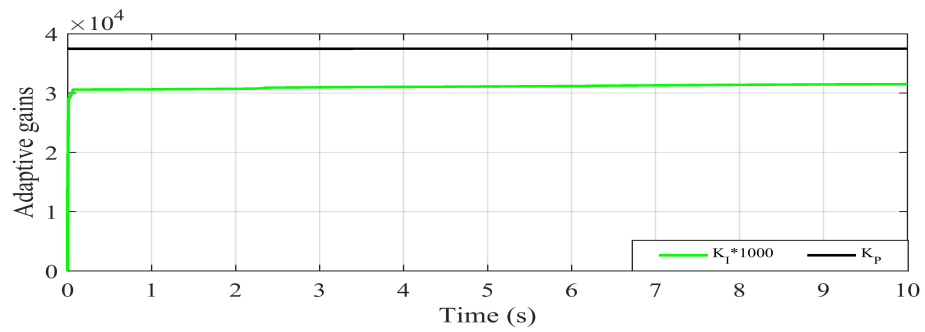


Figure 4.31: Adaptation gains of regulator PI Flux.

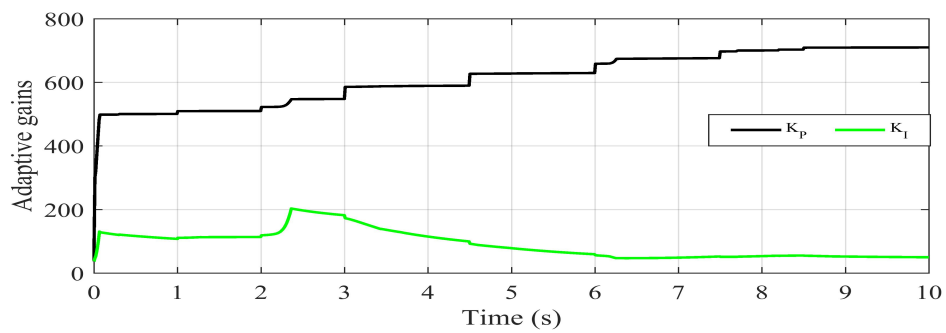


Figure 4.32: Adaptation gains of regulator PI Torque.

Table 4.3: Optimal adaptive rates optimized by BA

Aaptive rates	Flux	Torque	SC
η_{K_P}	500	0.0012	-
η_{K_I}	0.0142	0.0098	-
T	-	-	1.3273e-3

Table 4.4: Comparative study between the results obtained with different control studies

Criteria	DTC-IMC	FDTC-IMC	DTC-SVM-IMC
Following speed reference	Yes	Yes	Yes
Response time for speed	Very slow (0.13 rad/s)	Slow (0.1 rad/s)	Fast (0.065 rad/s)
Flux error	0.018 Wb	0.01 Wb	0.001 Wb
Torque error	Medium there are peaks (0.26 N.m)	Medium there are peaks (0.2 N.m)	Small (0.1 N.m)
Output currents	Sinusoidal with distortion	Sinusoidal there are distortion	Pure sine wave
Output current peak when reversing the rotation direction	25 A	22 A	16 A
THD of unfiltered input current	82.34 %	78.14 %	67.22 %
THD of filtered input current	21.18 %	9.16 %	2.6 %
Switching frequency	Variable	Variable	Fixed
Low-frequency harmonics in the input current	Yes	Yes	No

4.6 Conclusion

In this chapter, we have tackled the problem of the variable switching frequency and the synchronization between the two stages of IMC in the classical DTC by a new Adaptive Optimized DTC-SVM. The use of BA for a DSIM is suggested to improve the dynamic performances such as reducing the ripples of the torque and flux, minimized the low harmonic distortion in the input current, working with close-to-unity power factor under all operating conditions, and hence, a better stator current waveform was achieved. The proposed control scheme performance is enhanced by inserting a robust SC in the outer loop for speed regulation, this controller gives a faster response, very small tracking error, and rises robustness with external disorders. Simulation results with a comparative analysis have been presented in order to assess the effectiveness of the proposed method. These special merits of the suggested scheme make it competitive to the other existing control techniques in the literature.

Part II

Control of Wind Generator Connected to The Electrical Grid by Multilevel Matrix Converter based on the DSIG

Chapter 5

Improving the execution and competency control of wind power generation systems with DSIG coupled to the grid via a TLMC

“ *The less attachment to the world. The easier your life.* ”

Umar ibn Al-Khattab

5.1 Introduction

Today, the capabilities of wind turbines are rising quickly. There are now wind turbines with a capacity of 1.5 MW exist and are nevertheless constantly growing. In order to accommodate the added power, the generation voltage has been increased. Consequently, the switching voltage is reduced to a manageable level using a Multilevel Converter (MLC). The power converters can produce electrical power over a wide range of wind speeds. Thus, these converters must be well controlled in order to be able to transmit all the power generated by the machine into the grid, because they can generate significant harmonic disturbances. This chapter briefly introduces the modeling and control of the variable-speed wind turbine firstly. Then, a DTC-SVM based on a fractional PID controller to generate the desired stator voltages and the switching states of the inverter

stage of the ITLMC by the SVM is proposed in this chapter. To transform the maximum mechanical power captured by a wind turbine into electrical energy, a novel and robust Fractional Adaptive MPPT Synergetic Controller (FASC) is used to optimize this conversion. This technique improves the Double Star Inducing Generator (DSIG) performances (in terms of torque, current, flux, speed, active power, reactive power during startup, and changes in wind speed). Additionally, compared to the conventional hysteresis DTC with MC and ITLMC converter, total harmonic distortion (THD) is reduced without requiring the use of complicated controllers.

5.2 Brief description of the system

The studied system is connected to the network, considered powerful, and consists of a wind generator based on the dual stator induction machine(DSIG), matrix converter(MC), input filter, and transformer. The overall diagram of the system is represented in figure 5.1.

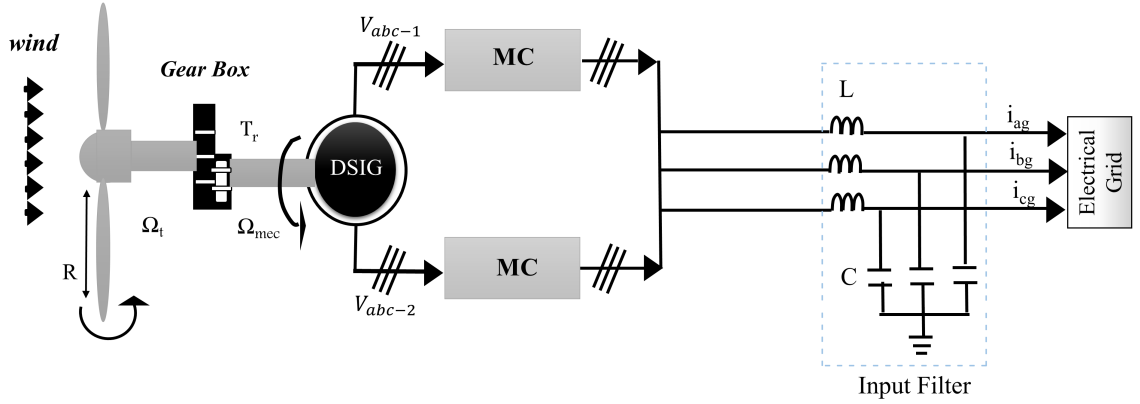


Figure 5.1: Schematic representation of the studied system.

5.3 Modeling of wind system

5.3.1 Wind energy conversion

A wind generator, more commonly known as a wind turbine, is a device that transforms part of the kinetic energy of the wind into mechanical energy on a transmission shaft, then into electrical energy via a generator [79, 102, 103].

Wind turbines are divided into three categories according to their nominal power and the diameter of their blade as shown in the following table 5.1:

Table 5.1: Classification of wind turbines according to power.

Wind turbine	length of the blades	Power
Small	less than 12 m	less than 40 kW
Medium	12 m to 45 m	from 45 kW to 1 MW
Grand	46 m and over	1 MW and above

5.3.2 Wind turbine types

Wind turbines are classified according to the geometric arrangement of the shaft on which the blades are mounted. There are mainly two main types: a vertical axis and a horizontal axis.

Vertical axis

Vertical axis wind turbines have the advantage of not requiring blade orientation systems, they have a mechanical part at ground level, thus facilitating the maintenance interventions. turbines [78, 81].



Figure 5.2: Vertical axis wind turbine.

In addition, some of these wind turbines must be trained at start-up, which are often very heavy and undergo strong mechanical stresses, thus pushing manufacturers to abandon

these aerogenerators in favor of horizontal axis wind turbines. Figure (5.2) shows an example of vertical axis wind.

Horizontal axis

Horizontal axis wind turbines are more widely used even if they often require blade orientation mechanisms. They have a higher aerodynamic efficiency than those provided by vertical axis wind turbines and start autonomously. In addition to the visual aspect of wind turbines, their impact on the environment is reduced [102,103].

A wind turbine only covers a very small percentage of the total area of the site on which it is installed, thus allowing most sites to maintain their industrial or agricultural activities. Their noise pollution is increasingly reduced. Figure (5.3) shows an example of horizontal axis wind turbines.



Figure 5.3: Horizontal axis wind turbine.

In our case, the wind system consists mainly:

- a horizontal axis turbine comprising three orientable blades and possessing specific coefficients of inertia, elasticity and friction to the air and with respect the support of the turbine;
- gearbox (G);
- a generator (DSIM) having inertia and a coefficient of friction.

5.3.3 Construction of wind turbine

A wind turbine generally consists of three main elements [102](figure 5.4):

- **Tower** : Usually made of steel tubing. Its height essentially depends on the dimensions of the blades. (for example, the radius of a wind turbine with a height of 100m is 45m).
- **Nacelle** : Collecting all the mechanical elements used to connect the wind rotor to the electric generator. The Nacelle consists of slow and fast shaft, speed multiplier, control system, cooling system, a wind vane indicating the direction of the wind, the anemometer measuring the wind speed and informing the control system when it blows quite hard and at the end of a disc brake (different from the aerodynamic brake) allowing to stop the system in the event of an overload. The generator is generally composed of a synchronous or asynchronous machine, hydraulic systems or electrical orientation of the blades (aerodynamic brake).
- **The rotor** : the rotor formed by the blades assembled in their hub.

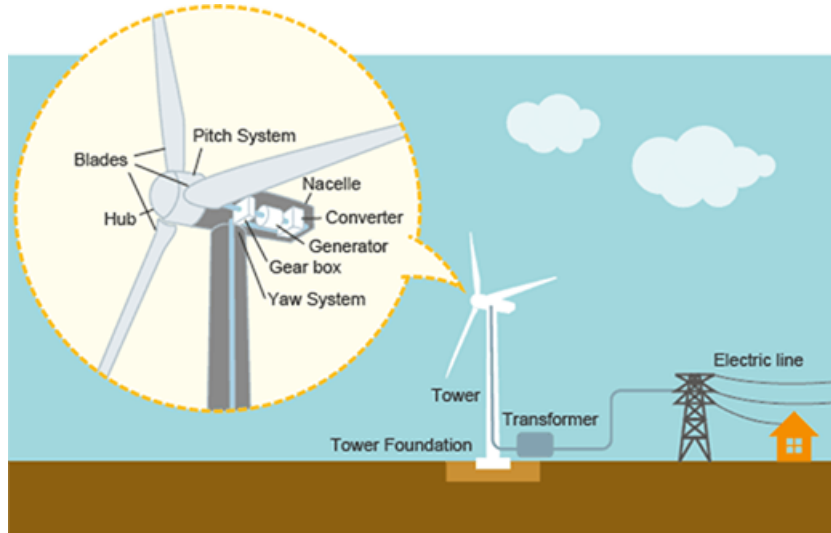


Figure 5.4: The main elements of a wind turbine.

5.3.4 Wind system working areas

The maximum power that can be produced by a wind turbine is its nominal power. The wind speed for which the nominal power of the wind turbine is reached is called nominal wind speed (V_{rated}). Above this, the power production remains constant regardless of wind speed until a maximum operating speed is reached at which the wind turbine is designed to shut down to protect its parts mechanics from the adverse effects of high wind speeds (V_{out}). The lowest wind speed at which a wind turbine will operate is called

the start-up speed (V_{in}). Between these two speeds, power output varies with wind speed, as shown in figure 5.5 [93, 104, 105].

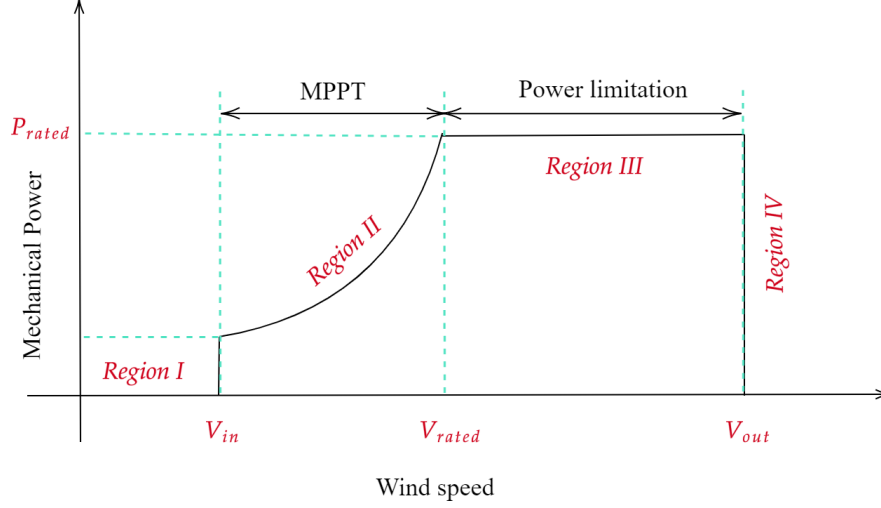


Figure 5.5: Wind system operating areas.

- Zone I: $\nu_{wind} < \nu_{in}$ where $P=0$: waiting phase (the turbine does not work);
- Zone II: $\nu_{in} < \nu_{wind} < \nu_{rated}$: this is the zone of maximum wind power extraction (MPPT);
- Zone III: $\nu_{rated} < \nu_{wind} < \nu_{out}$: the rotation speed is kept constant (power regulation at nominal value);
- Zone IV: $\nu_{wind} > \nu_{out}$: stopping the wind turbine.

5.3.5 Simplifying assumptions for the mechanical modeling of the turbine

The simplifying assumptions for the mechanical modeling of the turbine are given as [78, 105]:

- The wind speed is assumed to be uniformly distributed over all the blades, which makes it possible to consider all the blades as one and the same mechanical system;
- The coefficient of friction of the blades with respect to the air is very low and can be ignored;

- The friction losses of the turbine rotor are negligible compared to the friction losses on the generator side.

5.3.6 Wind turbine and gearbox models

In order to control the system, knowledge of the mechanical model of the turbine is necessary, the kinetic power of the wind is given by the following relation [79, 106, 107]:

$$P_{\text{wind}} = \frac{1}{2} m V_{\text{wind}}^2 \quad (5.1)$$

where

$$m = \rho S V_{\text{wind}} \quad (5.2)$$

We replace (5.2) in (5.1) we obtain:

$$P_{\text{wind}} = \frac{1}{2} \rho S V_{\text{wind}}^3 \quad (5.3)$$

and $S = \pi R^2$

Which give:

$$P_{\text{wind}} = \frac{1}{2} \rho \pi R^2 V_{\text{wind}}^3 \quad (5.4)$$

where:

- ρ : air density (which is 1.25 kg/m^3 in a normal atmosphere);
- S : circular surface swept by the turbine;
- R : the length of the blade;
- m : mass of air passing through the surface in one second;
- V_{wind} : wind speed (m/s).

The rotation of the blades by the wind creates a mechanical power on the shaft of the turbine expressed by:

$$P_t = \frac{1}{2} C_p(\lambda, \beta) \rho S V_{\text{wind}}^3 \quad (5.5)$$

where

C_p : power coefficient.

Numerical approximations have been developed in the literature to calculate the $C_p(\lambda, \beta)$ [102]. For the type of turbine studied here, the expression can be written by:

$$C_p(\lambda, \beta) = \left[0.73 \left(\frac{151}{\lambda'} \right) - 0.002 \beta - 13.2 \right] \exp \left(\frac{-18.4}{\lambda'} \right) \quad (5.6)$$

where

$$\lambda' = [1/(\lambda + 0.08\beta) - 0.035/(\lambda^3 + 1)]^{-1} \quad (5.7)$$

with

- β : the blades angle of orientation;
- λ : the speed ratio defined as the ratio between the speed linear of the blades, and the wind speed.

$$\lambda = \frac{\Omega_t R}{V} \quad (5.8)$$

The graphical representation of the power coefficient $C_p(\lambda, \beta)$ obtained with the expression (5.6), is illustrated in figure 5.6

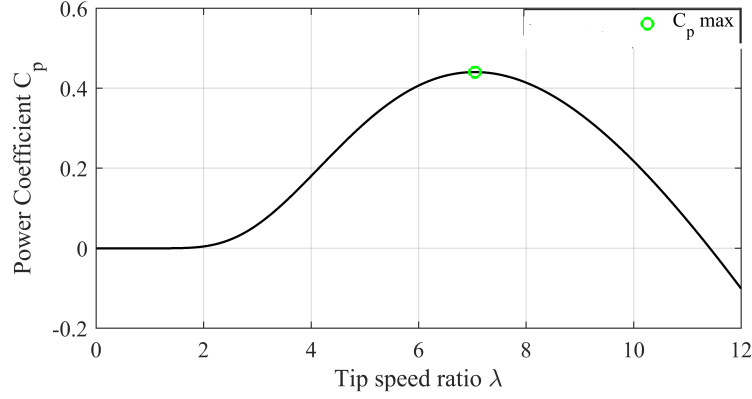


Figure 5.6: Power coefficient for the wind turbine model.

From the expression of the power (Eq 5.5) and knowing the speed of rotation of the turbine, the mechanical torque available on the shaft of the turbine is expressed as follows:

$$T_t = \frac{P_t}{\Omega_t} = C_p(\lambda, \beta) \frac{\rho}{2} S V_{wind}^3 \frac{1}{\Omega_t} \quad (5.9)$$

The multiplicator transforms the mechanical speed of the turbine into the speed of the generator and the aerodynamic torque of the turbine into the drive torque of the generator according to the following expressions:

$$\Omega_{mec} = G \cdot \Omega_{turb} \quad (5.10)$$

$$T_{mec} = \frac{T_{turb}}{G} \quad (5.11)$$

where

- Ω_{mec} : generator speed [rad/s];

- G : multiplicator gain;
- Ω_{turb} : angular speed of the turbine $[rad/s]$;
- T_{mec} : generator torque $[Nm]$;
- T_{turb} : turbine torque $[Nm]$.

The fundamental equation of the dynamics makes it possible to determine the evolution of the mechanical speed from the mechanical torque exerted on the rotor shaft of the wind turbine and the electromagnetic torque as follow [78, 81, 108]:

$$J \frac{d\Omega_{mec}}{dt} = T_{mec} - T_{em} - f\Omega_{mec} \quad (5.12)$$

where

- J : global moment of inertia ($J = \frac{J_{tur}}{G^2} + J_{gen}$) $[kg \cdot m^2]$;
- T_{em} : electromagnetic torque $[Nm]$;
- f : mechanical coefficient of friction $[N \cdot m \cdot s/rad]$.

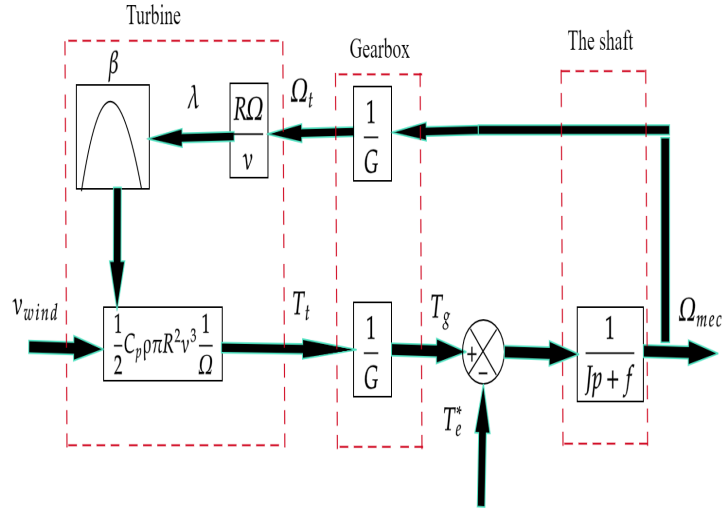


Figure 5.7: Block diagram of the wind converter model.

From the previous equations, the block diagram corresponding to the modeling of the turbine can be deduced as presented in figure 5.7.

We can remark that the turbine speed is controlled through the action on the electromagnetic torque of the DSIG.

5.4 Optimization of energy conversion of wind turbine

The development and utilization of wind energy have become an important part of energy development around the world because the wind is a sustainable and environmentally friendly source of energy. However, due to the erratic nature of this energy, the design, and control of the turbine modes must be optimized for extraction of the maximum possible power from wind energy. Different algorithms have been developed to obtain the optimum operating points of maximum power transfer. In general, they determine the operation of the turbine through its internal characteristics. Maximum Power Point Tracking (MPPT) is a set of wind energy conversion techniques. Various studies are being conducted to achieve better and faster MPPT techniques for any Wind Energy Conversion system (WECS) [81, 103, 105].

There are two types structure of the MPPT control:

- MPPT based on the mechanical speed control;
- MPPT without mechanical speed control.

5.4.1 MPPT based on mechanical speed control

The principle of the Maximum Power Point Tracking (MPPT) strategy consists, for a given blade angle, in adjusting, depending on the wind speed V_{wind} , the rotational speed of the turbine to a reference value Ω_{mec}^* to maintain an optimal relative speed in order to get a maximum power coefficient C_{p_max} . From the relation (5.8), we can rewrite the expression (5.10) as follows:

$$\Omega_{mec}^* = \left(\frac{v_{wind} \cdot \lambda_{opt}}{R} \right) \cdot G \quad (5.13)$$

The action on the electromagnetic torque (and therefore on the power converted by the generator) that will make it possible to obtain Ω_{mec}^* and thus maximize the power extracted with the C_{p_max} . Figure 5.8 illustrates the basic principle of this technique.

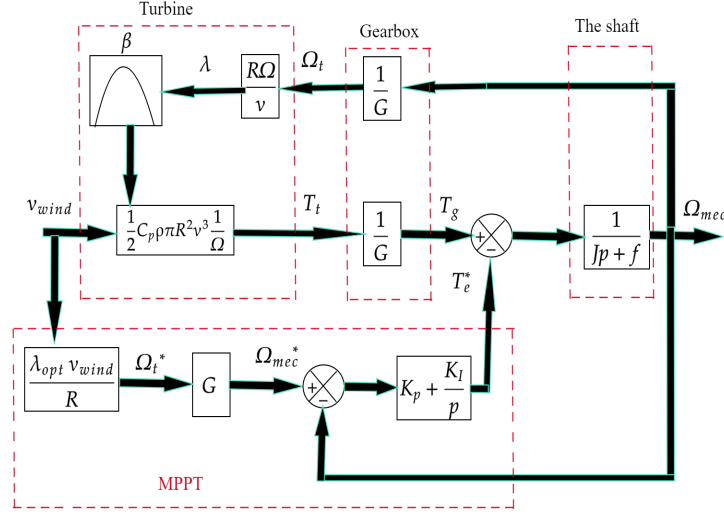


Figure 5.8: MPPT based on mechanical speed control.

5.4.2 MPPT without mechanical speed control

This control mode is based on the assumption that the wind speed varies very little in steady state, from the dynamic equation of the turbine, we can obtain the equation describing the steady state of the aerogenerator:

$$J \frac{d\Omega_{mec}}{dt} = T_{mec} - T_{em} - f\Omega_{mec} = 0 \quad (5.14)$$

If we neglect the effect of the viscous torque ($f\Omega_{mec}$) we obtain:

$$T_{mec} = T_{em} \quad (5.15)$$

So, the reference of electromagnetic torque is estimated as a function of the aerogenerator torque:

$$T_{eref} = \frac{T_t}{G} \quad (5.16)$$

The aerogenerator torque is itself estimated as a function of the wind speed and the turbine speed (equation 5.17):

$$T_{test} = C_p(\lambda, \beta) \frac{\rho}{2} S V_{wind-es}^3 \frac{1}{\Omega_t} \quad (5.17)$$

The estimated turbine speed is calculated from the measurement of the mechanical speed:

$$\Omega_{test} = \frac{\Omega_{mec}}{G} \quad (5.18)$$

The measurement of the wind speed being delicate, an estimate of its value can be obtained from the equation:

$$V_{wind_{est}} = \frac{\Omega_{test} \cdot R}{\lambda} \quad (5.19)$$

Thus, we obtain a global command relation:

$$T_{eref} = \frac{1}{2} C_P(\lambda, \beta) \rho \pi R^5 \frac{\Omega_{mec}^2}{G^3 \lambda^3} \quad (5.20)$$

To extract the maximum power generated, it is necessary to fix the speed ratio λ_{opt} which corresponds to the maximum of the power coefficient $C_{P_{opt}}$.

The electromagnetic torque reference must then be fixed to the following value:

$$T_{eref} = \frac{1}{2} C_{P_{opt}}(\lambda, \beta) \rho \pi R^5 \frac{\Omega_{mec}^2}{G^3 \lambda_{opt}^3} \quad (5.21)$$

The following figure 5.9 presents the principle of the MPPT control of the wind turbine without speed control.

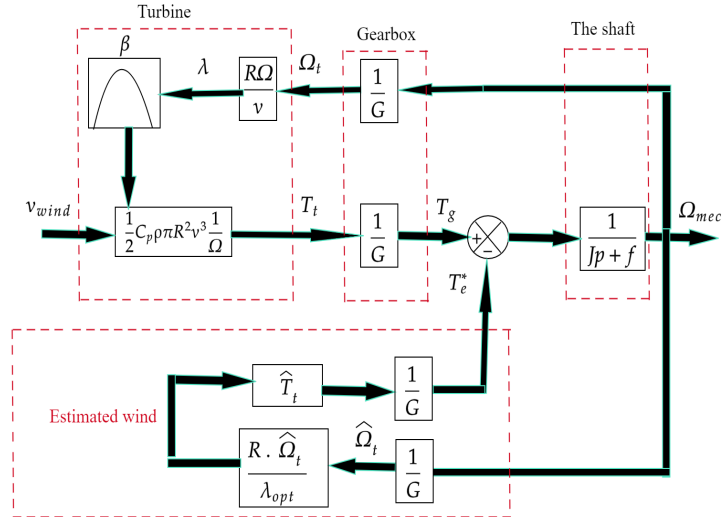


Figure 5.9: MPPT without speed control.

5.5 The proposed Fractional DTC-SVM based on MPPT-FASC

The proposed fractional DTC-SVM of the Double Star Induction Generator (DSIG) integrated into the electrical grid by an Indirect Three Level Matrix Converter (ITLMC) based on MPPT fractional adaptive synergetic controller (FASC) design procedures is shown in the following developments.

5.5.1 DTC-SVM of Indirect Three Level Matrix Converter (ITLMC)

The evolution of multilevel inverters has been a motive for the development of the multi-level matrix converter and in particular that of the three-level matrix converter [46, 109, 110].

The use of these inverters makes it possible to improve the quality of the output voltage by synthesizing them in three levels and also makes it possible to reduce the maximum voltage of the switches during dimensioning [21]. Combining this with the advantages of the matrix converter, the indirect three level matrix converter (ITLMC) can generate better voltages than the IMC. This converter uses the DC bus voltage U_{pm} and the neutral point of the input filter to synthesize the output voltages [46, 109].

Figure 5.10 shows the circuit of the ITLMC converter. It consists of two stages: a rectifier stage like the IMC and an inverter stage formed by a three-level inverter structure NPC. They are connected in three points (p, n, 0): the points p and n are the outputs of the rectifier and the point “0” is the neutral point of the input filter.

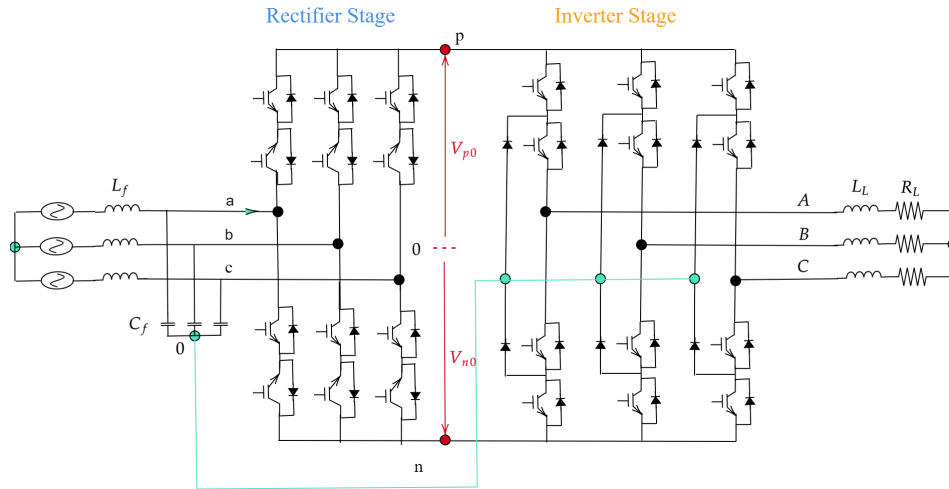


Figure 5.10: Topology of ITLMC.

Indeed, each arm can have three possible states which are illustrated in Figure 5.11.

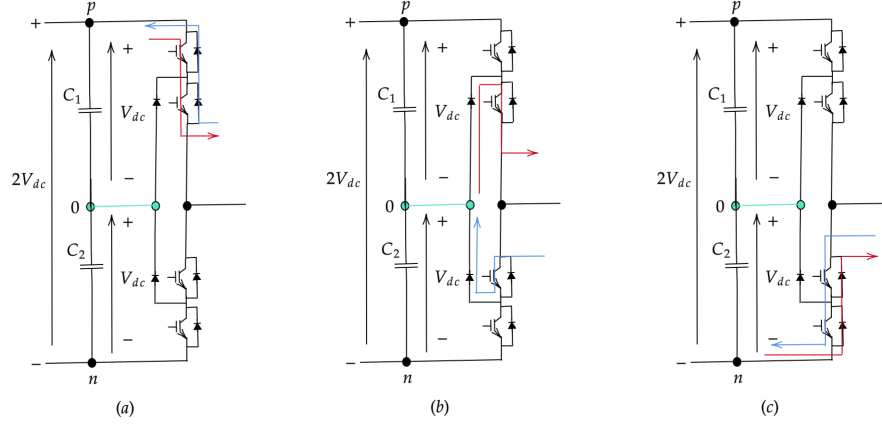


Figure 5.11: Possible states of an arm of the NPC inverter: (a) state p (b) state 0 (c) state n.

The latter also shows that the switches (S_{1x} , S_{3x}) and (S_{2x} , S_{4x}), where x represents one of the three arms, are complementary two by two [46, 109].

- **State p** (Figure 5.11.a): the two switches (S_{1X} , S_{2X}) are closed while the other two are open.
- **State o** (Figure 5.11.b): the two switches (S_{2X} , S_{3X}) are closed while the other two are open.
- **State n** (Figure 5.11.c): this is the inverse of state p.

Table 5.2 summarizes these three states of the arm, as well as the value of its voltage with respect to the midpoint 0, while table B.2 (Appendix B) illustrates all 27 possible configurations of the inverter. Based on this table, we can define the voltages of the arms

Table 5.2: possible states of an arm

S_{1x}	S_{2x}	S_{3x}	S_{4x}	V_{0x}	State of the arm
1	1	0	0	V_{dc}	p
0	0	1	1	$-V_{dc}$	n
0	1	1	0	0	0

with respect to the midpoint 0 as being a connection matrix multiplied by the voltages of the DC bus. This can be translated by the following equation:

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = [M] \times \begin{bmatrix} V_{p0} \\ V_{n0} \end{bmatrix} \text{ and } \begin{bmatrix} I_p \\ -I_p \end{bmatrix} = [M]^t \times \begin{bmatrix} I_A \\ I_B \\ I_C \end{bmatrix} \quad (5.22)$$

- little vectors: LV ($V_1, V_4, V_7, V_{10}, V_{13}, V_{16}$)
- Zero vectors: ZV (V_0)

For category 3 (LV), each vector has two redundant configurations, giving us more choices for balancing the DC bus voltages. The null vector has three redundant configurations. And the other two categories of vectors have only one configuration.

BV divide space into 6 sectors as shown in figure 5.12. Each sector can be divided into 4 regions (see figure 5.13).

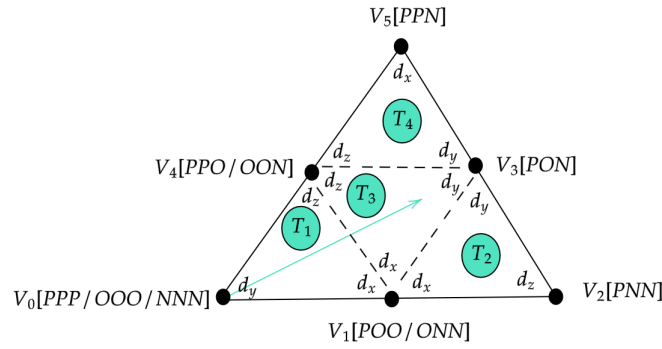


Figure 5.13: regions of a sector.

These regions are triangles whose heads are the ends of the vectors. SVM modulation in this case uses three vectors to synthesize the reference output voltage vector. These three vectors are the heads of the region where the reference voltage vector is located. For example in the case of figure 5.13 the vectors used are the vectors V_1 , V_3 and V_4 . So

$$\overline{V_S} = d_x v_x + d_y v_y + d_z v_z \quad (5.26)$$

where d_x , d_y and d_z are the duty cycles of the vectors V_x , V_y and V_z respectively. They are defined for each region as shown in table 5.3. m_v is the inverter voltage ratio defined by:

$$m_v = \frac{\sqrt{3} \cdot |V_S|}{U_{pn}} \quad (5.27)$$

The control of the ITLMC is generally done in the same way as the IMC. The rectifier is controlled to have a maximum U_{pn} voltage and a desired input phase shift by SVM. Also, the inverter is controlled with SVM to synthesize the output voltages from two levels as for the IMC, plus an additional level created from the midpoint of the input filter capacitances. This allows the outputs to take the 0 voltage of this level like the two voltage extremes. This makes it possible to reduce the reverse voltage of the switches by

Table 5.3: Duty cycles for the different regions

region	d_x	d_y	d_z
T_1	$2m_v \sin(\frac{\pi}{3} - \theta_o)$	$1 - 2m_v \sin(\frac{\pi}{3} + \theta_o)$	$1 - d_x - d_y$
T_2	$1 - 2m_v \sin(\theta_o)$	$2m_v \sin(\frac{\pi}{3} + \theta_o) - 1$	$1 - d_x - d_y$
T_3	$2 - 2m_v \sin(\frac{\pi}{3} + \theta_o)$	$2m_v \sin(\theta_o)$	$1 - d_x - d_y$
T_4	$2m_v \sin(\theta_o) - 1$	$2m_v \sin(\frac{\pi}{3} - \theta_o)$	$1 - d_x - d_y$

two compared to the IMC and makes it possible to reduce the harmonics of the output voltages with the SVM to have the desired output voltages. To synthesize the output voltages and avoid the problem of low-frequency harmonics, synchronization between the two stages is essential.

The synchronization between the rectifier stage and inverter stage

Synchronization between the two stages is essential due to the absence of the dc link capacitor and to guarantee that the output multilevel voltage is capable of being at the right level. Figure 5.14 shown it.

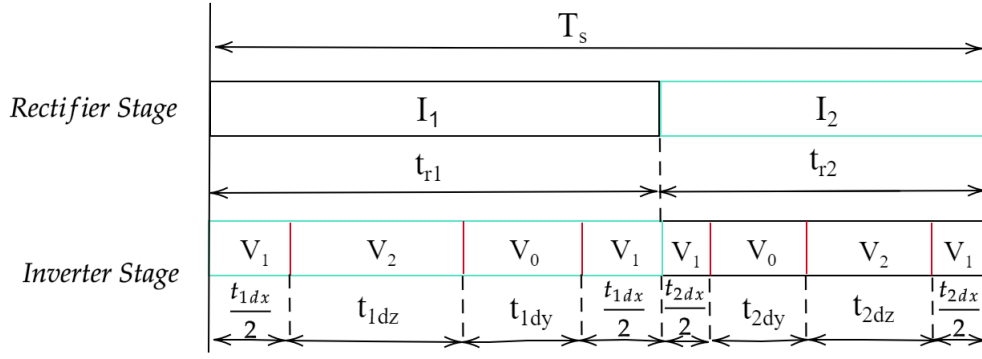


Figure 5.14: The synchronization between the two stages of ITLMC.

5.5.2 Design of fractional PID controllers

The fractional order control of a plant is defined by fractional order differential equations. Accordingly, fractional calculus authorizes the derivatives and integrals to be of arbitrary

order [111]. The calculus operator is described as:

$${}_a D_t^\alpha = \begin{cases} \frac{d^\alpha}{dt^\alpha} & \alpha > 0 \\ 1 & \alpha = 0 \\ \int_a^t (d\tau)^{-\alpha} & \alpha < 0 \end{cases} \quad (5.28)$$

where α is the fractional order, a and t are operational limits.

Podlubny in [112] first suggested the structure of a FOPID controller, which can be expressed in s domain as:

$$U_{FOPID}(s) = (K_p + K_i/s^\lambda + K_d s^\mu) E(s) \quad (5.29)$$

Profoundly, a FOPID controller has five parameters to make the design adaptable. Where, proportional, integral and derivative constants are correspondingly the gains K_p , K_i and K_d while λ and μ are fractional order operations. The extension effect of a classical PID controller based on fractional calculus gives the FOPID controller. Plainly, one takes λ and μ equal 1 as a result we get a classical PID controller whilst PI and PD controllers can be attained by selecting $(\lambda = 1, \mu = 0)$ and $(\lambda = 0, \mu = 1)$ respectively as presented in figure 5.15. Thus, the closed-loop system characteristics controlled by means of a fractional order PID cautiously conquer more performance for mitigating the low-frequency oscillations in a consistent manner.

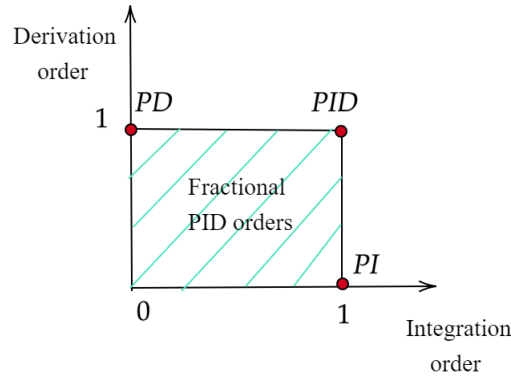


Figure 5.15: A fractional order PID form.

Figure 5.16 depicts the fundamental architecture of the fractional PID controller that optimized by BA to regulate the flux and torque. The output of these controllers is the reference voltages $(V_{ds1,2}, V_{qs})$, that the inverter stage's SVM will produce.

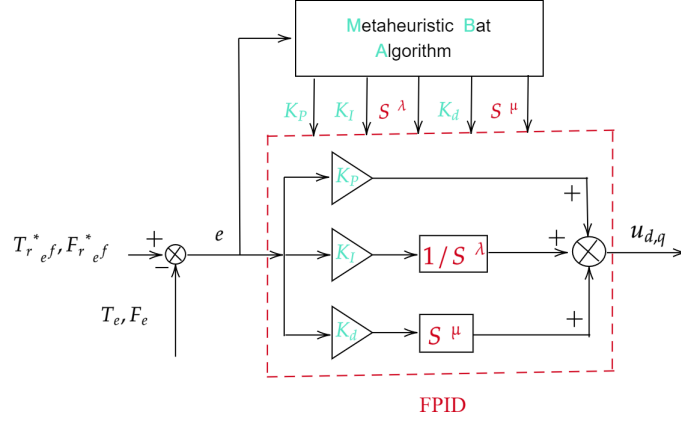


Figure 5.16: A fractional order PID control system optimized by BA.

5.5.3 Fractional Adaptive Synergetic MPPT Controller

To operate the system on the manifold where the macro-variable is equal to zero is the goal of the synergetic controller. According to the synergetic control theories, the speed controller generates the reference torque T_{ref} .

The speed error is defined by:

$$e(t) = \Omega_{ref} - \Omega_r = 0 \quad (5.30)$$

Consider the fractional macro-variable given by:

$$\Psi = K_P e(t) + K_I^\lambda \int e(t) \quad (5.31)$$

where K_p and K_I^λ are the proportional and fractional integral parameters of the speed macro-variable.

$$\dot{\Psi} = K_P \dot{e}(t) + K_I^\lambda e(t) \quad (5.32)$$

The dynamic equation of DSIG is given as:

$$\dot{\Omega} = \frac{1}{J} (T_r - T_{em} - f_r \Omega) \quad (5.33)$$

where, J : inertia moment, f_r : coefficient of friction, C_r : Load torque.

By substituting Eq. (5.33) in the equation (4.31), and basing on the synergetic theory, we can write:

$$T (K_P (\dot{\omega}_{ref} - \dot{\omega}_r) + K_I^\lambda (\omega_{ref} - \omega)) + K_P (\omega_{ref} - \omega_r) + K_I^\lambda (\omega_{ref} - \omega_r) = 0 \quad (5.34)$$

$$T_{em}^* = T_r - f_r \Omega - J \left(\dot{\Omega}_{ref} + \frac{1}{K_P} \left(\frac{\Psi}{T} + K_I^\lambda (\Omega_{ref} - \Omega) \right) \right) \quad (5.35)$$

Indeed, the K_P and K_I are given by the adaptive gradient descent algorithm. A Bat Algorithm optimization is demonstrated in the previous chapter (chapter 4) to get efficient parameters. The general structure of the proposed Fractional Adaptive Synergetic Controller (FASC) is shown in figure 5.17.

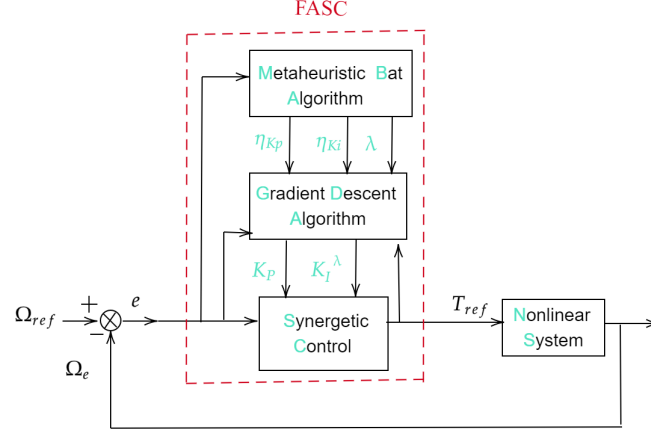


Figure 5.17: Fractional adaptive optimized MPPT synergetic controller.

The whole basic structure of the proposed control method is shown in Fig. 5.18.

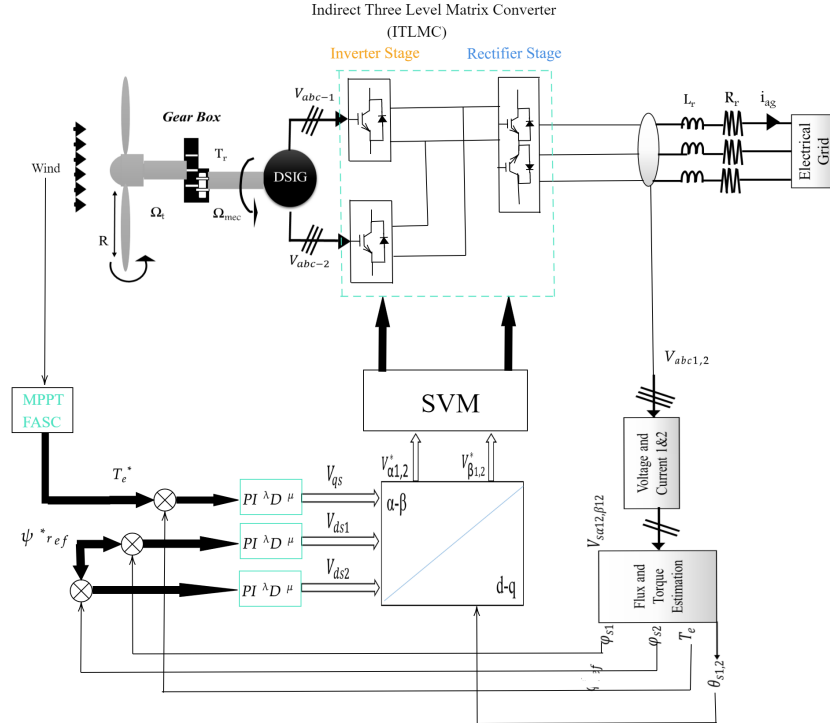


Figure 5.18: Block diagram of fractional DTC-SVM scheme for ITLMC based on FASC.

5.6 Numerical simulation

Several simulation experiments have been conducted for various operating circumstances to illustrate the advantages offered by the suggested control. Software from Matlab and Simulink is used to simulate the process. In order to see its performance and compare it with the DTC-ITLMC and DTC - MC. Table A.2 (Appendix A) is a list of the parameters used in this simulation.

The simulation was run with a variable wind speed profile (Fig.5.19). The MPPT control provides an electromagnetic torque reference to ensure maximum power control in any wind area. Fig 5.20 illustrates the allure of speed. One can observe that the rotational speed perfectly matches its reference, which varies with the imposed wind profile. As can be shown, the FASC performs better than the traditional PI and SC controller in all areas, including response time, tracking inaccuracy, and robustness to external disturbances. it implies that the MPPT is properly ensured. Figs. 5.21–5.22 demonstrate that the proposed FASC controller converges more quickly than its counterparts (PI and SC), approaching the ideal values of $C_p = 0.44$ and $\lambda = 7.05$. The electromagnetic torque closely tracks the variations in the references caused by the wind and variations in rotating shaft speed (see Figures. 5.23-5.25). Additionally, it is noted that the suggested DTC-SVM based on ITLMC has faster and significantly fewer ripples in its torque dynamics compared to the traditional DTC with MC and ITLMC. Figures 5.25-5.31 illustrate the flux magnitude and circular trajectory , accordingly. It is clear that applying the suggested adjustment can significantly reduce flux ripples. Throughout the whole operational range, the flux follows its reference. This highlights the effective decoupling between the flux and torque axes that the suggested control achieves. The output voltage is a multilevel voltage (Figure 5.32) and low harmonic distortion compared to others (Figures 5.33 and 5.34), it varies in amplitude and frequency according to the speed reference variation. The voltage of a phase connected to the grid as well as the current are illustrated in Figures(5.35, 5.37 and 5.39). The frequency of the electrical network is 50Hz. It has been shown that the voltage is imposed by the grid and the amplitude of the current injected into the grid is imposed by the command. We also note that the voltage wave of the network and that of the current of the latter are in phase opposition, which means that the system operates under the unity power factor, This owes to the benefits of using the matrix converter. This justifies that the flow of power is always from the wind turbine to the electrical network. Figures (5.36, 5.38 and 5.40) show the THD in the input current, one can see that the P-DTC has lower THD compared to the others. Figures 5.41 and

5.42 show the active and reactive powers supplied to the electrical network. Figure 5.43 shows the input power factor of the grid, one can see that the system is operating under the unity power factor in all the possibilities of operation conditions. The neutral current is variable according to the modulation index (Figure 5.44), but its average value is zero, which avoids distortions in the voltages of the capacitors (input voltage of the converter).

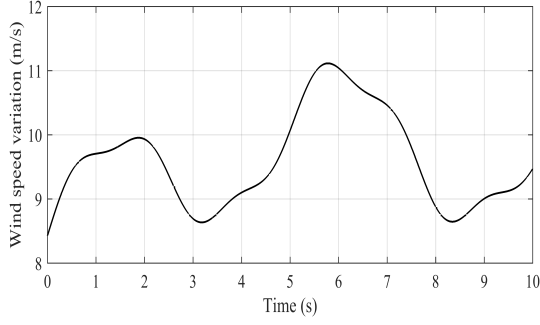


Figure 5.19: Wind speed.

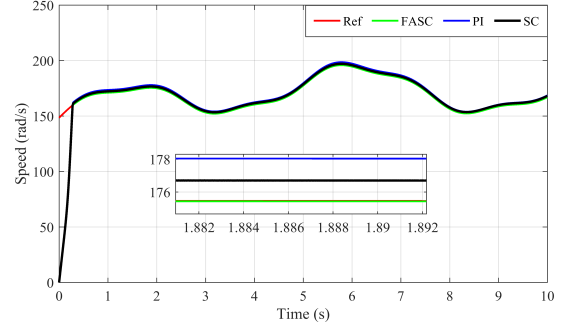


Figure 5.20: Speed of DSIG.

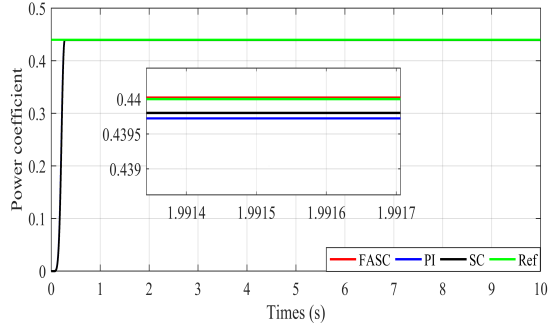


Figure 5.21: Power coefficient.

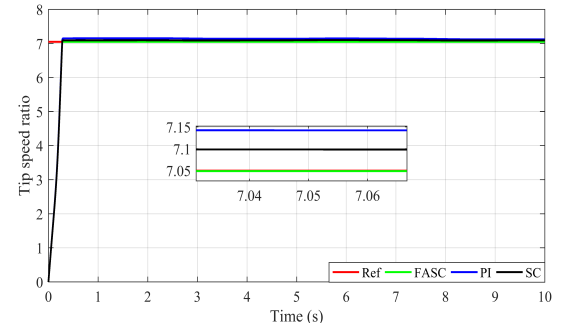


Figure 5.22: Tips speed ratio.

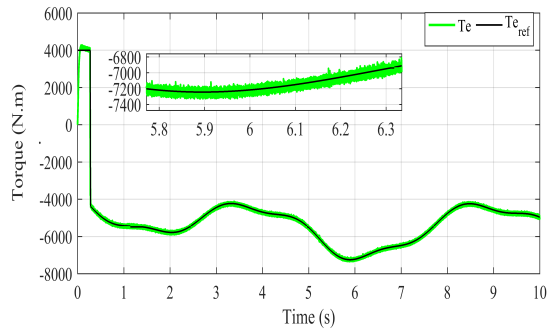


Figure 5.23: Electromagnetic torque with PDTC.

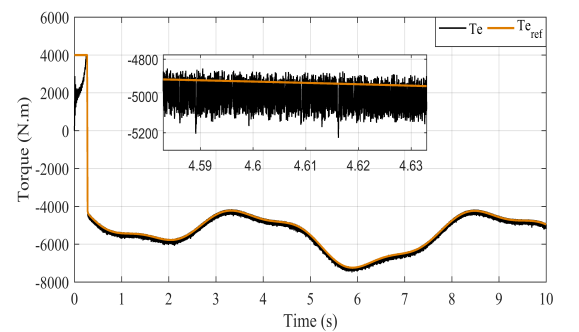


Figure 5.24: Electromagnetic torque with DTC-ITLMC.

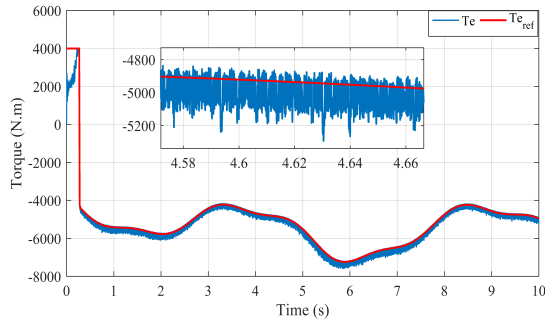


Figure 5.25: Electromagnetic torque with DTC-MC.

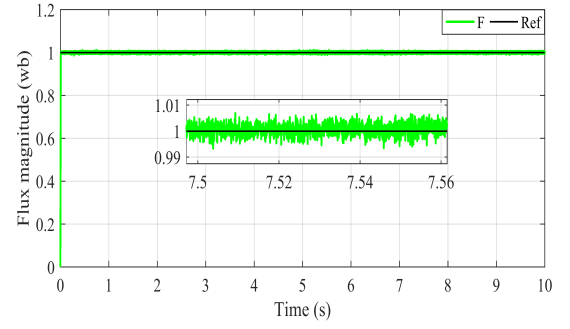


Figure 5.26: Stator flux magnitude with PDTC.

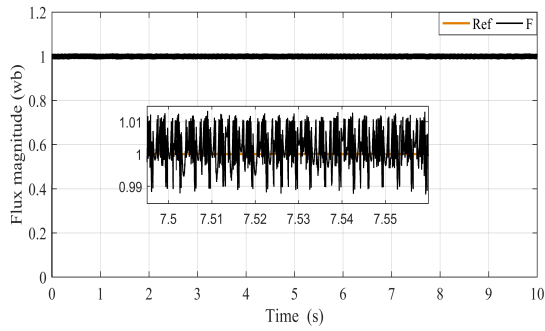


Figure 5.27: Stator flux magnitude with DTC-ITLMC.

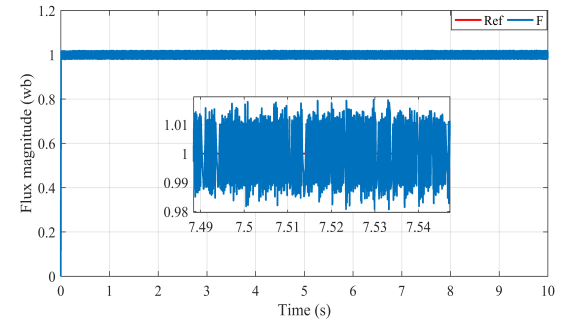


Figure 5.28: Stator flux magnitude with DTC-MC.

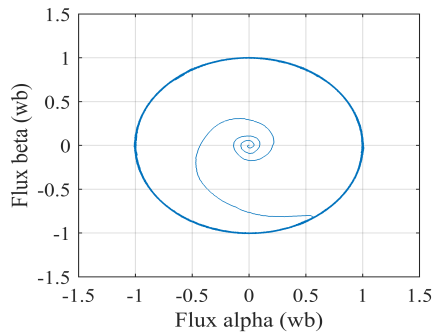


Figure 5.29: Stator flux circular trajectory with PDTC.

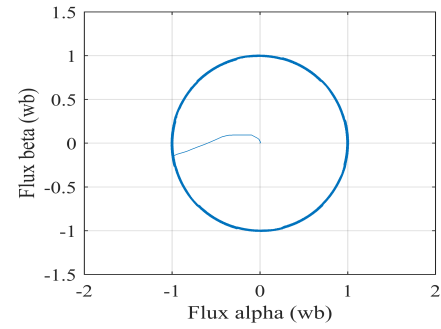


Figure 5.30: Stator flux circular trajectory with DTC-ITLMC.

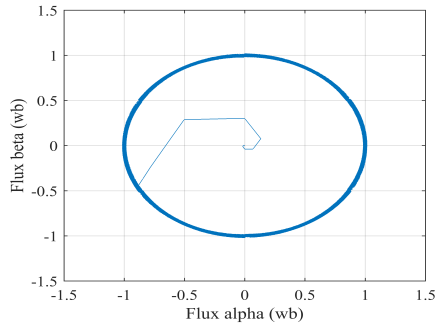


Figure 5.31: Stator flux circular trajectory with DTC-MC.

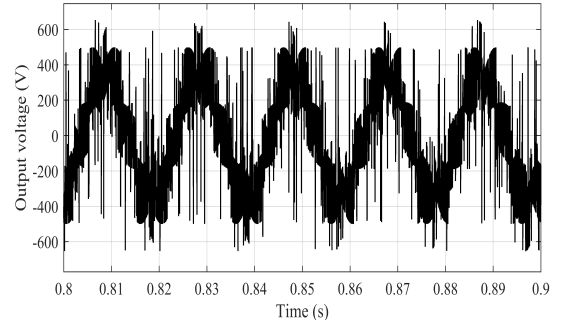


Figure 5.32: Output voltage of ITLMC with PDTC.

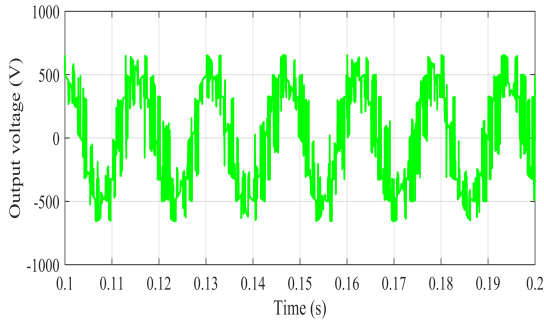


Figure 5.33: Output voltage of ITLMC with DTC.

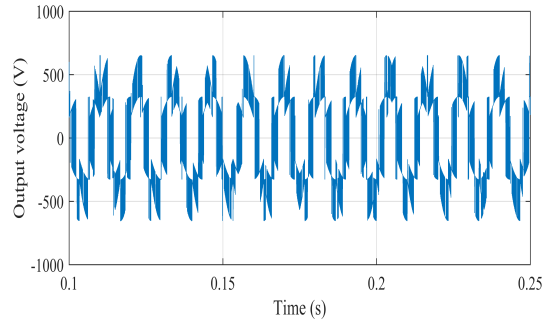


Figure 5.34: Output voltage of MC with DTC.

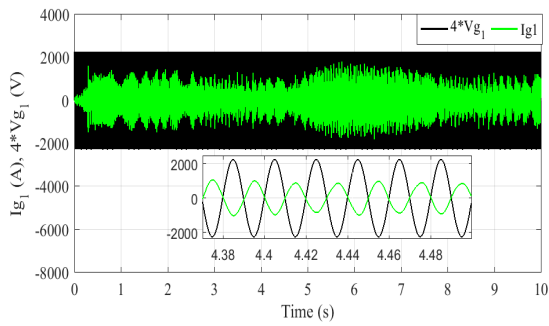


Figure 5.35: Grid voltage and current with PDTC.

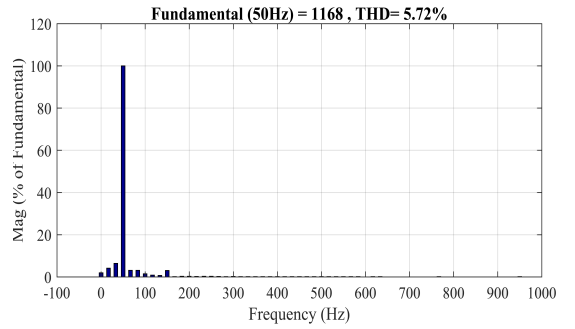


Figure 5.36: THD of grid current with PDTC.

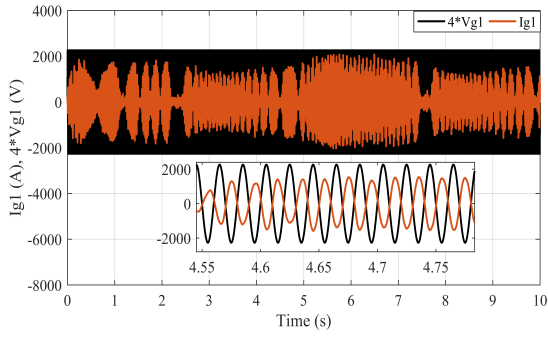


Figure 5.37: Grid voltage and current with DTC-ITLMC.

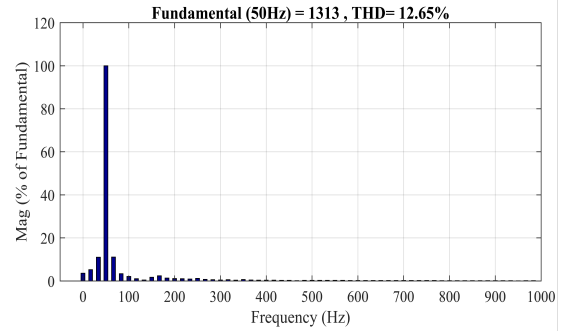


Figure 5.38: THD of grid current with DTC-ITLMC.

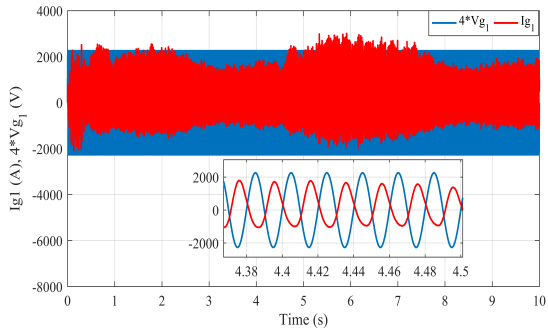


Figure 5.39: Grid voltage and current with DTC-MC.

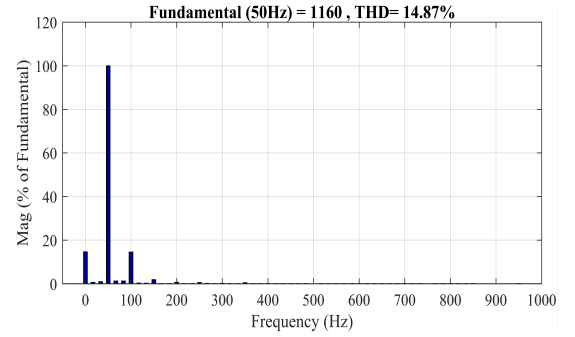


Figure 5.40: THD of grid current with DTC-MC.

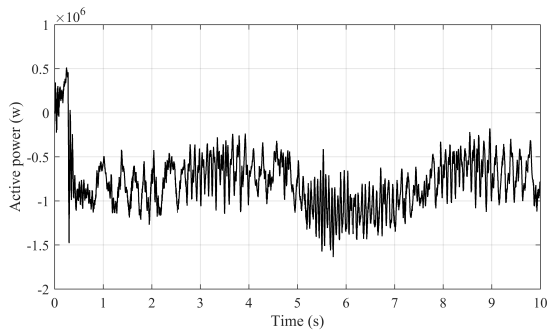


Figure 5.41: Active power.

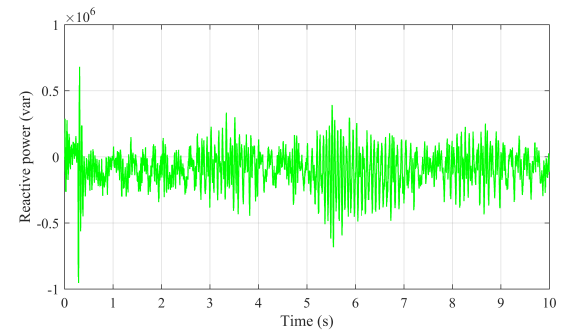


Figure 5.42: Reactive power.

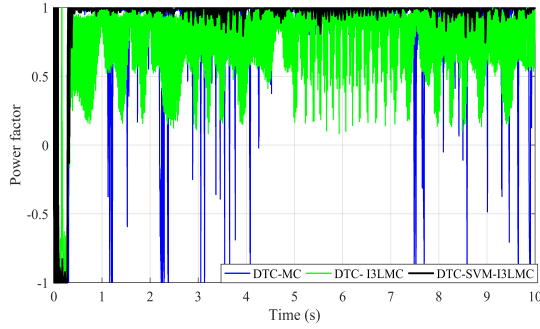


Figure 5.43: Power factor.

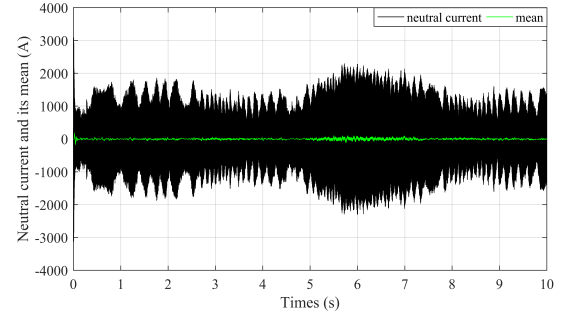


Figure 5.44: Neutral current and its average value

5.7 Conclusion

This chapter studies and describes a 1.5 MW wind turbine integrated into the grid by an indirect three-level matrix converter (ITLMC). A DSIG and the ITLMC are used in the analysis, modeling, and simulation of a variable-speed wind turbine. Two factors made the ITLMC necessary: the existence of high-power systems that required high-power converters and the high caliber of the output waveforms that were supplied. The stable operation of the turbine was achieved by applying a fractional DTC-SVM control. In addition, the proposed control scheme performance is enhanced by inserting a robust FASC controller to achieve maximum power. The suggested algorithm is put to the test using Matlab/Simulink in order to confirm the theoretical results. The findings attained attest to the effectiveness and reliability of the suggested control.

General Conclusion

The study and control of matrix converters (MC) and their usage in powering a dual stator induction machine, were the research topic of the research covered by this thesis, where it gave some contributions. This study is carried out by simulation under Matlab/Simulink. The first chapter of this work was devoted to the general study and the art state of this type of converter, its different topologies, and the modulations proposed in the literature. The bibliographical study showed that this research theme is topical and promising, attracting the attention of researchers around the world. Then, the instantaneous modeling was treated as well as the average modeling with the connection generating functions and its various components (power circuit, input filter, protection). Finally, the industrial implementation of the matrix converter for the variation of speed and/or frequency was presented. This last modest task remains somewhat limited and is dominated by Japanese companies, in particular the Yaskawa company.

We have dedicated the second chapter to the DTC of DSIM fed by two types of matrix converters namely direct and indirect. The simulation results obtained illustrate that the DTC with IMC is superior in all operating conditions, it yields good electrical and electromagnetic execution such as the stator flux, electromagnetic torque, and total harmonic distortion (THD) in the input current. Furthermore, the control method is also much simplified and easier to implement when the IMC is used.

In the third chapter, we have proposed an improvement of DTC control based on a fuzzy inference system instead of the hysteresis controllers and switching table. According to the results obtained, the main objectives have been achieved. The speed perfectly follows its reference value even with the variation of the resistive torque, the average torque follows the steady state of load torque and the same for the flux.. Note that with DTC, the ripple is fixed by the band of the hysteresis comparator. Fuzzy-DTC reduces torque and flux ripples. It also improves tracking of the setpoint, especially when reversing the speed. Indeed, the major interest of fuzzy logic in control is the capacity to control system with high uncertainty level. Its effectiveness is proven when modeling a system is difficult,

which is the case with our system.

The fourth chapter aimed at an adaptive DTC-SVM optimized by Metaheuristic Bat Algorithm (MBA). The latter was proposed as a solution for the problem of the variable switching frequency and the synchronization between the two stages of IMC in the classical DTC. With this proposed method, dynamic performances such as reducing the ripples of torque and flux, minimizing the low harmonic distortion in the input current, working with close-to-unity power factor under all operating conditions, and hence, a better stator current waveform were achieved. By including a robust synergetic controller (SC) in the outer loop for speed regulation, the performance of the suggested control method is improved.

After presenting an overview of the mathematical model of variable speed turbine and associated components, the power of the captured wind, and the converted electrical power. and control of the variable speed wind turbine, a fractional order DTC-SVM for a 1.5 MW wind turbine using a multilevel matrix converter (ITLMC) based on an adaptive synergetic MPPT controller was proposed in the final chapter 5. The findings attained attest to the effectiveness and reliability of the suggested control.

Finally, we can conclude that the matrix converter is an important alternative for the AC/AC conversion given its advantages and its multitudes of applications, and given the immensity of the research work on this type of converter, it is expected that it will soon replace conventional AC-AC converters in several applications including wind applications, due to the advantages it will offer over other options.

In the future perspectives, this work can be continued as follows:

- In order to validate the developed control algorithms, it is necessary to create an experimental test bench.
- Studying grid-DSIM interaction in the presence of grid instability. Look at other matrix converter structures, particularly multi-level and polyphase structures.
- Use of matrix converters in renewable energy systems for improving the quality of electrical energy (active filtering).
- Analysis of the switching issue in matrix converters and the implementation of smooth switching.

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Appendix A

Parameters

A.0.1 The parameters of the DSIM

Table A.1: The parameters of DSIM and turbine

Parameters	Units	Value
$R_{s1} = R_{s2}$	Ω	3.72
R_r	Ω	2.12
$L_{s1} = L_{s2}$	H	0.022
L_r	H	0.006
L_m	H	0.3672
J	kg.m^2	0.0625
K_f	N.m.s/rad	0.001
P_n	kW	4.5
P	-	1

A.0.2 The parameters of the DSIG

Table A.2: Parameters of DSIG

DSIG parameters	
P_n	1.5 MW
V_n	400 V
$R_{s1} = R_{s2}$	$0.008 \, \Omega$
$L_{s1} = L_{s2}$	$1.34 \, 10 \times 10^{-4}$
L_m	$0.0045 \, H$
R_r	$0.007 \, \Omega$
L_r	$6.7 \, 10 \times 10^{-5}$
p	2
F_s	50 Hz
J	$30 \, \text{kg.m}^2$
K_f	$2.5 \, \text{Nm s/rd}$
Turbine parameters	
R	36 m
Number of blades	3
Hub height	85 m
Gearbox G	90

Appendix B

All possible configurations of MC and a three-level NPC inverter are presented:

Table B.1: Possible configurations for MC

Configuration		State	$ V_S $	α_S	I_I	β_I
abb	+1	active	$\frac{2}{\sqrt{3}}V_{AB}$	$\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_A$	$-\frac{\pi}{6}$
baa	-1	active	$\frac{2}{\sqrt{3}}V_{AB}$	$-\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_A$	$\frac{5\pi}{6}$
bcc	+2	active	$\frac{2}{\sqrt{3}}V_{BC}$	$\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_A$	$\frac{\pi}{2}$
cbb	-2	active	$\frac{2}{\sqrt{3}}V_{BC}$	$-\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_A$	$-\frac{\pi}{2}$
caa	+3	active	$\frac{2}{\sqrt{3}}V_{CA}$	$\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_A$	$-\frac{5\pi}{6}$
acc	-3	active	$\frac{2}{\sqrt{3}}V_{CA}$	$-\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_A$	$\frac{\pi}{6}$
bab	+4	active	$\frac{2}{\sqrt{3}}V_{AB}$	$\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_B$	$-\frac{\pi}{6}$
aba	-4	active	$\frac{2}{\sqrt{3}}V_{AB}$	$-\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_B$	$\frac{5\pi}{6}$
cbc	+5	active	$\frac{2}{\sqrt{3}}V_{BC}$	$\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_B$	$\frac{\pi}{2}$
bc b	-5	active	$\frac{2}{\sqrt{3}}V_{BC}$	$-\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_B$	$-\frac{\pi}{2}$
aca	+6	active	$\frac{2}{\sqrt{3}}V_{CA}$	$\frac{5\pi}{2}$	$\frac{2}{\sqrt{3}}I_B$	$-\frac{5\pi}{6}$
cac	-6	active	$\frac{2}{\sqrt{3}}V_{CA}$	$-\frac{\pi}{6}$	$\frac{2}{\sqrt{3}}I_B$	$\frac{\pi}{6}$
bba	+7	active	$\frac{2}{\sqrt{3}}V_{AB}$	$-\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$-\frac{\pi}{6}$
aab	-7	active	$\frac{2}{\sqrt{3}}V_{AB}$	$\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$\frac{5\pi}{6}$
ccb	+8	active	$\frac{2}{\sqrt{3}}V_{BC}$	$-\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$\frac{\pi}{2}$
bbc	-8	active	$\frac{2}{\sqrt{3}}V_{BC}$	$\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$-\frac{\pi}{2}$
aac	+9	active	$\frac{2}{\sqrt{3}}V_{CA}$	$-\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$-\frac{5\pi}{6}$
cca	-9	active	$\frac{2}{\sqrt{3}}V_{CA}$	$\frac{\pi}{2}$	$\frac{2}{\sqrt{3}}I_C$	$\frac{\pi}{6}$
aaa	0_1	zero	0	-	0	-
bbb	0_2	zero	0	-	0	-
ccc	0_3	zero	0	-	0	-
abc		rotating	V_I	α_i	I_S	β_S
acb		rotating	V_I	$-\alpha_i + \frac{\pi}{3}$	I_S	$-\beta_S$
bac		rotating	V_I	$-\alpha_i - \pi$	I_S	$-\beta_S + \frac{2\pi}{3}$
bca		rotating	V_I	$\alpha_i + \frac{4\pi}{3}$	I_S	$\beta_S + \frac{2\pi}{3}$
cab		rotating	V_I	$\alpha_i + \frac{2\pi}{3}$	I_S	$\beta_S + \frac{4\pi}{3}$
cba		rotating	V_I	$-\alpha_i - \frac{\pi}{3}$	I_S	$-\beta_S + \frac{4\pi}{3}$

Table B.2: Possible states of the NPC inverter and its spaces vector

state of arm			Output voltage vector		
A	B	C	Vector	Amplitude	phase
P	P	P	V_0	0	0
O	O	O			
N	N	N			
P	O	O	V_1	$\frac{2}{3}V_{dc}$	0
O	N	N			
P	N	N	V_2	$\frac{4}{3}V_{dc}$	0
P	O	N	V_3	$\frac{2}{\sqrt{3}}V_{dc}$	$\frac{\pi}{6}$
P	P	O	V_4	$\frac{2}{3}V_{dc}$	$\frac{\pi}{3}$
O	O	N			
P	P	N	V_5	$\frac{4}{3}V_{dc}$	$\frac{\pi}{3}$
O	P	N	V_6	$\frac{2}{\sqrt{3}}V_{dc}$	$\frac{\pi}{2}$
O	P	O	V_7	$\frac{2}{3}V_{dc}$	$\frac{2\pi}{3}$
N	O	N			
N	P	N	V_8	$\frac{4}{3}V_{dc}$	$\frac{2\pi}{3}$
N	P	O	V_9	$\frac{2}{\sqrt{3}}V_{dc}$	$\frac{5\pi}{6}$
O	P	P	V_{10}	$\frac{2}{3}V_{dc}$	π
N	O	O			
N	P	P	V_{11}	$\frac{4}{3}V_{dc}$	π
N	O	P	V_{12}	$\frac{2}{\sqrt{3}}V_{dc}$	$-\frac{5\pi}{6}$
O	O	P	V_{13}	$\frac{2}{3}V_{dc}$	$-\frac{2\pi}{3}$
N	N	O			
N	N	P	V_{14}	$\frac{4}{3}V_{dc}$	$-\frac{2\pi}{3}$
O	N	P	V_{15}	$\frac{2}{\sqrt{3}}V_{dc}$	$-\frac{\pi}{2}$
P	O	P	V_{16}	$\frac{2}{3}V_{dc}$	$-\frac{\pi}{3}$
O	N	O			
P	N	P	V_{17}	$\frac{4}{3}V_{dc}$	$-\frac{\pi}{3}$
P	N	O	V_{18}	$\frac{2}{\sqrt{3}}V_{dc}$	$-\frac{\pi}{6}$

Appendix C

Definition of THD

$$THD = \frac{\sqrt{V_2^2 + V_3^2 + V_4^2 + \dots + V_n^2}}{V_1}$$