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بِسْمِ اللَّهِ الرَّحْمَنِ الرَّحِيمِ

وَأَن لَّيْسَ لِلْإِنسَانِ

أَلَّا مَّا سَعَىٰ

[النجم]

Dedication

I dedicate this thesis to my younger self for believing in herself and choosing to follow the major that she likes, although the hard work and efforts it took and how she was able to go through all of that.

To my parents Amel and Zine El abidine Ksantini who were always there for me every step of the way. Their financial and emotional support ensured that I studied in the best conditions. I wouldn't have arrived here without their support.

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Lydia

Dedication

I dedicate this thesis to my father, TEHAMI Mohammed, whose financial support made this journey possible. His sacrifice and dedication ensured that I had everything I needed to focus on my studies and reach this milestone.

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List of Abbreviations

Abbreviation	Signification
DSSG	Dual-Star Synchronous Generator
PWM	Pulse Width Modulation
THD	Total Harmonic Distortion
MLI	Multilevel Inverter
NPC	Neutral Point Clamped Inverter
FC	Flying Capacitor Inverter
MMC	Modular Multilevel Converter
FFT	Fast Fourier Transform

General Introduction

Electrical energy generation from renewable energy sources is widely adopted nowadays due to the increase in electricity consumption and the worldwide active investment in cleaner energy resources. The integration of renewable energy resources with the grid plays an important role in energy utilization. However, it is difficult to utilize electricity from renewable energy sources directly for grid injection due to their intermittency. Hence, the system requires power electronics converters as an interface between the renewable energy sources and the grid.

With economic development and population growth, great opportunities for development in the twenty-first century are opening for technologies, especially in electrical energy production and conversion. Several disciplines that shape the core elements of energy generation and conversion have attracted increasing researcher attention. We particularly mention dual-star configurations in synchronous machines. The six-phase machine in generator mode reveals its full potential, offering major improvements over conventional three-phase generators in terms of high power handling capabilities, improved fault tolerance, and power density.

To fully explore the power delivered by the DSSG, polyphase rectifiers are crucial, especially commanded ones (PWM). These represent a major improvement over conventional diode-based rectifiers, offering instantaneous control of the current waveform drawn from the AC source and regulation of reactive power flow. The inverters are crucial components responsible for converting the regulated DC voltage back into three-phase AC voltage suitable for grid synchronization and injection. Multilevel inverters have become widely used in modern power electronics due to their capability to generate high-quality voltage waveforms with reduced harmonic distortion. The multilevel structure provides smooth output voltage and lower voltage stress, with reduced dv/dt stress on loads and lower electromagnetic interference compared to conventional topologies.

This thesis represents, in the first place, a continuity of the work done by our fellow colleagues in studying the DSSG-PWM rectifier feeding a DC bus. Our work was focused on fully exploring the use of the DSSG-PWM rectifier in an energy conversion chain up to the DC-AC conversion stage. We investigate different multilevel inverter topologies, comparing them based on several criteria including redundancy, ease of development, ability to add levels, control strategy complexity, and THD, in order to evaluate the overall performance and suitability of each topology within the conversion chain.

The first chapter of this thesis presents a description of the energy conversion chain for renewable energy applications, covering each stage from mechanical rotational energy conversion up to the DC-AC conversion stage, with a general overview of each stage and its role within the conversion chain.

Chapter two is dedicated to the modeling of the DSSG-PWM rectifier conversion system. A deep understanding of the dynamic modeling of these two components is essential, as it not only allows us to fully understand the work done by our fellow colleagues, but also to fully explore the advantages of this system within the energy conversion chain.

Chapter three investigates three distinct multilevel inverter families: neutral point clamped multilevel inverter, flying capacitor multilevel inverter, and modular multilevel inverter (MMC). For the first two topologies, we investigate 3-level and 5-level configurations. For the MMC, we investigate 3-level, 5-level, 9-level, and 19-level configurations. For each topology, two modulation strategies are evaluated: level-shifted PWM and carrier phase-shifted PWM. The different topologies are compared based on several criteria including control complexity, ease of adding levels, and THD.

Chapter four covers the integration of the three multilevel inverter topologies with a dual-star synchronous generator through a rectifier stage, forming a complete energy conversion chain. The DC-link voltage is supplied by the rectified output of the generator, allowing a more realistic evaluation of each topology's behavior under dynamic supply conditions. The results are then compared to draw a final conclusion on the performance and suitability of each topology within the conversion chain.

Chapter 1: Renewable Energy Conversion Chain

1.1 Introduction

Electrical energy generation from renewable energy sources is widely adopted due to the increase in electricity consumption. The integration of renewable energy sources plays an important role in energy utilization. Due to their intermittent nature, renewable energy sources require power electronics converters to process and condition the generated power effectively. These converters must produce pure sinusoidal waves of voltage and current with minimal harmonics. [1]

This chapter presents the energy conversion chain for renewable energy applications, covering the stages from mechanical rotational energy conversion up to the DC-to-AC conversion stage. A general overview of the conversion chain is provided, including the system architecture and the role of each component. The conversion chain ends at the DC-to-AC conversion stage, as it represents the most critical and performance-defining step in the chain, where the quality of the output power is ultimately determined. For this reason, this study focuses mainly on the analysis and comparison of three multilevel inverter topologies, based on several criteria.

1.2 System Architecture

The proposed energy conversion chain follows a sequential power transformation process:

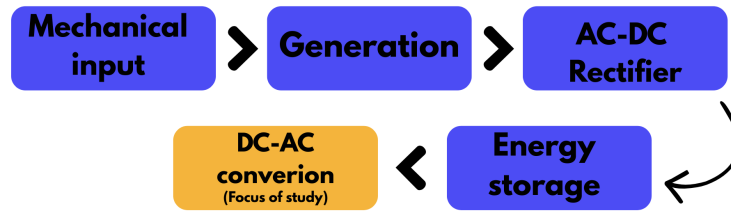


Figure 1.1: Different Stages of the Energy Conversion Chain

Each stage plays a specific role in the conversion, conditioning, and control of the energy to ensure optimal performance and compatibility with the requirements of the target application.

1.3 The Energy Conversion Chain components

1.3.1 Mechanical Input

The system begins with a mechanical rotational input representing the prime mover driving the synchronous generator. In renewable energy applications, this rotation is produced by

a natural source such as wind, where kinetic energy is converted into rotational motion at a speed close to the generator's rated synchronous speed. In this study, a constant rotational speed input is used in Simulink to keep the focus on the electrical conversion stages, particularly the inverter topologies, without adding complexity from the mechanical side.

1.3.2 Dual-Star Synchronous Generator (DSSG)

The DSSG is the core generation component that converts mechanical rotation into electrical power. It features a unique dual-star configuration with sinusoidal flux distribution. This multi-phase output configuration provides advantages in efficiency, stability with improved fault tolerance, and power density compared to conventional three-phase generators.

1.3.3 Pulse Width Modulation Rectifier

The PWM rectifier is an active power electronics converter that transforms the AC output from the DSSG into regulated direct current (DC). Unlike conventional diode-based rectifiers or rectifiers with thyristors, the PWM rectifier uses switches entirely controlled at switching on and off: MOSFET transistors for medium and small powers, IGBT for medium and large powers, and GTO for very large powers. The switches come with diodes in antiparallel. These switches are unidirectional in voltage and bidirectional in current. Thus, this converter, by its structure, is reversible in current and is able to control in an instantaneous way the current waveform drawn from the DSSG. It can also regulate the flow of reactive power through its control system, providing considerable advantages in power quality management.[2]

1.3.4 DC-Bus

The rectifier converts hexaphase alternating voltage produced by the DSSG into continuous voltage. However, this voltage remains rippled due to the switching operation of semiconductors. The high-capacity capacitor connected at the output of the rectifier smooths this voltage by attenuating the ripples by storing energy during overvoltage periods and restoring it during voltage dips. Thus, it ensures a more stable DC voltage necessary for proper operation of loads or downstream converters (inverters). Furthermore, it acts as a temporary energy reserve, capable of compensating for rapid voltage variations and reducing electromagnetic disturbances generated by switching operations. This role is particularly important in applications where the quality of DC power supply is crucial.[3]

1.3.5 Multilevel Inverter Topologies

The inverter is the crucial component responsible for converting regulated DC voltage back into three-phase AC voltage suitable for grid synchronisation and injection. This thesis investigates three distinct multilevel inverter families:

- Neutral-Point-Clamped (NPC) multilevel inverter

- Flying-Capacitor multilevel inverter
- Modular Multilevel Converter (MMC)

For the first two topologies, the levels investigated are: 3-level and 5-level. For the MMC, 3-level, 5-level, 9-level, and 19-level are investigated. For each topology, two modulation strategies are evaluated: level-shifted PWM and carrier phase-shifted PWM. The different topologies will be compared based on several criteria including redundancy, ease of development, ability to add levels, control strategy complexity and THD . These three multilevel inverter topologies will be studied in the context of using them with the DSSG, PWM rectifier, and DC bus to evaluate their overall system performance and suitability for renewable energy applications.

1.4 Key Contributions

This thesis contributes to the field of renewable energy conversion by:

- Proposing and analyzing an energy conversion chain for renewable energy applications, covering the stages from mechanical rotational energy conversion up to the DC-to-AC conversion stage.
- Providing a comprehensive comparison of three multilevel inverter topologies based on redundancy, ease of development, ability to add levels, control strategy complexity, and THD.
- Investigating the impact of different Pulse-Width Modulation strategies on inverter performance.
- Presenting detailed simulation-based validation of the energy conversion chain up to the inversion stage.

1.5 Conclusion

The energy conversion chain studied in this thesis represents a modern, efficient approach to converting renewable energy into regulated AC electrical power. By combining advanced generation technology that explores the use of polyphase machines, active power conversion , and multiple inverter topology options, the system offers flexibility and performance advantages over conventional approaches. The following chapters provide a detailed technical analysis of each stage, including mathematical modeling, control strategies, and performance evaluation. These aspects are essential to ensure a rigorous and reliable evaluation of the three multilevel inverter topologies investigated in this work.

Chapter 2: Dynamic Modeling of the DSSG-PWM Rectifier Energy Conversion Chain

2.1 Introduction

Synchronous machines with a high number of phases have attracted increasing research attention, particularly in large-scale electrical energy production and conversion. While the three-phase system long dominated, research into multiphase systems (four-phase, five-phase, and beyond) has emerged, with the six-phase system proving most practical and valuable. Six-phase synchronous machines offer significant advantages in efficiency, stability, and high power handling capabilities, with improved fault tolerance and power density. PWM rectifiers represent a major improvement over conventional diode-based rectifiers, offering instantaneous control of current waveform drawn from the AC source. They enable supplying regulated DC voltage while maintaining absorbed current in phase with the AC source voltage, achieving power factors very close to unity. Additionally, PWM rectifiers regulate reactive power flow through their control system, providing considerable advantages in power quality management. Their adaptability to load variations and grid disturbances allows seamless integration into existing infrastructure. [3] [4].

This chapter is dedicated to the modeling of the Dual-Star Permanent Magnet Synchronous Machine with round rotor and the mathematical modeling of the PWM rectifier associated with the DSSG in hexaphase configuration. A deep understanding of the dynamic models is essential, as it allows not only a precise analysis of their characteristics, but also an accurate prediction of their behavior under various operating conditions, particularly when the generated currents are non-sinusoidal, a case that is complex to apprehend. This study investigates the energy conversion chain comprising the DSSG, the PWM rectifier, and the capacitive DC bus feeding the resistive load.

2.2 Mathematical Modeling of the DSSG

The Dual-Star Synchronous Generator consists of a stator equipped with two sets of balanced and identical three-phase windings, designated (a, b, c) and (a', b', c') , spatially shifted by an electrical angle γ . The rotor, which constitutes the rotating part of the machine, is composed of permanent magnets and receives a mechanical power input, ensuring a constant rotational speed.[5]

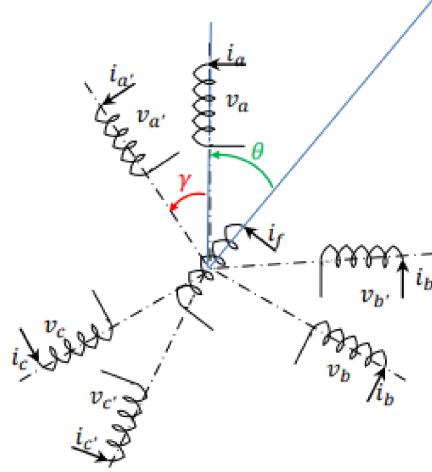


Figure 2.1: Winding arrangement of the Dual Star Synchronous Generator [3]

To establish simple relationships between the electrical quantities of the machine, a number of simplifying assumptions will be adopted:

- Magnetic circuit saturation, as well as hysteresis and eddy current losses, are neglected. This allows the self and mutual inductances of the windings to be defined in a straightforward manner.
- The windings of each stator phase, as well as those of the rotor, are assumed to produce a sinusoidally distributed flux. This simplifies the expression of the mutual inductances between the stator and rotor phases.
- The machine is assumed to have a symmetrical construction, with a constant air gap, and thermal effects on the magnetic circuit are not taken into account.

These assumptions imply, among other things, that the fluxes are additive, that the self-inductances are constant, and that the mutual inductances between the stator and rotor windings vary sinusoidally as a function of the electrical angle between their magnetic axes.

2.2.1 Model in the Canonical Basis (a, b, c) and (a', b', c')

Considering the machine as a six-phase machine and taking into account the magnetic coupling and the angular shift between the two stars, the electrical equation of the machine can be written as:

$$\begin{cases} v_a = \frac{d\Phi_{fa}}{dt} - R_s i_a - \frac{d}{dt} (l_s i_a + m_s (i_b + i_c) + m_1 i_{a'} + m_2 i_{b'} + m_3 i_{c'}) \\ v_b = \frac{d\Phi_{fb}}{dt} - R_s i_b - \frac{d}{dt} (l_s i_b + m_s (i_a + i_c) + m_3 i_{a'} + m_1 i_{b'} + m_2 i_{c'}) \\ v_c = \frac{d\Phi_{fc}}{dt} - R_s i_c - \frac{d}{dt} (l_s i_c + m_s (i_a + i_b) + m_2 i_{a'} + m_3 i_{b'} + m_1 i_{c'}) \\ v_{a'} = \frac{d\Phi_{fa'}}{dt} - R_s i_{a'} - \frac{d}{dt} (m_1 i_a + m_3 i_b + m_2 i_c + l_s i_{a'} + m_s (i_{b'} + i_{c'})) \\ v_{b'} = \frac{d\Phi_{fb'}}{dt} - R_s i_{b'} - \frac{d}{dt} (m_2 i_a + m_1 i_b + m_3 i_c + l_s i_{b'} + m_s (i_{a'} + i_{c'})) \\ v_{c'} = \frac{d\Phi_{fc'}}{dt} - R_s i_{c'} - \frac{d}{dt} (m_3 i_a + m_2 i_b + m_1 i_c + l_s i_{c'} + m_s (i_{a'} + i_{b'})) \end{cases} \quad (2.1)$$

The matrix representation of the equation is:

$$\begin{bmatrix} v_a \\ v_b \\ v_c \\ v_{a'} \\ v_{b'} \\ v_{c'} \end{bmatrix} = \frac{d}{dt} \begin{bmatrix} \Phi_{fa} \\ \Phi_{fb} \\ \Phi_{fc} \\ \Phi_{fa'} \\ \Phi_{fb'} \\ \Phi_{fc'} \end{bmatrix} - R_s \begin{bmatrix} i_a \\ i_b \\ i_c \\ i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix} - \begin{bmatrix} l_s & m_s & m_s & m_1 & m_2 & m_3 \\ m_s & l_s & m_s & m_3 & m_1 & m_2 \\ m_s & m_s & l_s & m_2 & m_3 & m_1 \\ m_1 & m_3 & m_2 & l_s & m_s & m_s \\ m_2 & m_1 & m_3 & m_s & l_s & m_s \\ m_3 & m_2 & m_1 & m_s & m_s & l_s \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \\ i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix} \quad (2.2)$$

In order to facilitate further development, the inductance matrix [6] can be conveniently decomposed. This decomposition allows the sub-matrices specific to each winding system to be distinguished, those of the first star and those of the second star, as well as the sub-matrices capturing the inter-star magnetic couplings.

Considering a balanced three-phase system, the stator inductance matrix L_{ss} can be formulated as follows, in terms of the leakage inductance l_f and the magnetizing inductance M_{ss} :

$$L_{ss} = \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_3 \\ m_s & m_s & L_s \end{bmatrix} = \begin{bmatrix} l_f + M_{ss} & M_{ss} \cos\left(\frac{2\pi}{3}\right) & M_{ss} \cos\left(\frac{4\pi}{3}\right) \\ M_{ss} \cos\left(\frac{4\pi}{3}\right) & l_f + M_{ss} & M_{ss} \cos\left(\frac{2\pi}{3}\right) \\ M_{ss} \cos\left(\frac{2\pi}{3}\right) & M_{ss} \cos\left(\frac{4\pi}{3}\right) & l_f + M_{ss} \end{bmatrix} \quad (2.3)$$

The magnetic interaction between the second star and the first star manifests itself through mutual couplings between their respective windings. These couplings are formalized and presented in the following matrix:

$$\begin{bmatrix} m_1 & m_2 & m_3 \\ m_3 & m_1 & m_2 \\ m_2 & m_3 & m_1 \end{bmatrix} = M_{ss} \begin{bmatrix} \cos(\gamma) & \cos\left(\gamma - \frac{2\pi}{3}\right) & \cos\left(\gamma - \frac{4\pi}{3}\right) \\ \cos\left(\gamma - \frac{4\pi}{3}\right) & \cos(\gamma) & \cos\left(\gamma - \frac{2\pi}{3}\right) \\ \cos\left(\gamma - \frac{2\pi}{3}\right) & \cos\left(\gamma - \frac{4\pi}{3}\right) & \cos(\gamma) \end{bmatrix} \quad (2.4)$$

Conversely, the magnetic influence of the first star on the second is characterized by mutual couplings between their windings. These interactions are modeled and represented by the following matrix:

$$\begin{bmatrix} m_1 & m_3 & m_2 \\ m_2 & m_1 & m_3 \\ m_3 & m_2 & m_1 \end{bmatrix} = M_{ss} \begin{bmatrix} \cos(\gamma) & \cos\left(\gamma - \frac{4\pi}{3}\right) & \cos\left(\gamma - \frac{2\pi}{3}\right) \\ \cos\left(\gamma - \frac{2\pi}{3}\right) & \cos(\gamma) & \cos\left(\gamma - \frac{4\pi}{3}\right) \\ \cos\left(\gamma - \frac{4\pi}{3}\right) & \cos\left(\gamma - \frac{2\pi}{3}\right) & \cos(\gamma) \end{bmatrix} \quad (2.5)$$

The arrangement of the two stars has a direct impact on the electromotive forces (e.m.f.) induced in each phase. This influence is explicitly incorporated into the matrix representation below:

$$\frac{d}{dt} \begin{bmatrix} \Phi_{fa} \\ \Phi_{fb} \\ \Phi_{fc} \\ \Phi_{fa'} \\ \Phi_{fb'} \\ \Phi_{fc'} \end{bmatrix} = \frac{d}{dt} \begin{bmatrix} \Phi_f \cos(p\theta) \\ \Phi_f \cos\left(p\theta - \frac{2\pi}{3}\right) \\ \Phi_f \cos\left(p\theta - \frac{4\pi}{3}\right) \\ \Phi_f \cos(p\theta + \gamma) \\ \Phi_f \cos\left(p\theta + \gamma - \frac{2\pi}{3}\right) \\ \Phi_f \cos\left(p\theta + \gamma - \frac{4\pi}{3}\right) \end{bmatrix} = - \begin{bmatrix} \omega_s \Phi_f \sin(p\theta) \\ \omega_s \Phi_f \sin\left(p\theta - \frac{2\pi}{3}\right) \\ \omega_s \Phi_f \sin\left(p\theta - \frac{4\pi}{3}\right) \\ \omega_s \Phi_f \sin(p\theta + \gamma) \\ \omega_s \Phi_f \sin\left(p\theta + \gamma - \frac{2\pi}{3}\right) \\ \omega_s \Phi_f \sin\left(p\theta + \gamma - \frac{4\pi}{3}\right) \end{bmatrix} \quad (2.6)$$

2.2.2 Equations of the DSSG in the Concordia plane

The Concordia transformation is a method used to convert a three-phase system into a static orthonormal two-phase system, to which a zero-sequence component is added. The axes of this two-phase coordinate system are denoted by α and β , which are in quadrature with respect to each other. The main advantage of this transformation lies in its ability to decouple the equations, which greatly simplifies the analysis of the DSSG.

- **Concordia Transformation for the First Star**

The Concordia transformation is performed using an orthogonal matrix C_1 that represents the projection of the three-phase system vectors (a, b, c) onto the (α, β) plane (Figure 2.2) by adding the zero-sequence component. This matrix allows the conversion of the three-phase system components (a, b, c) into the two-phase components (α, β) .

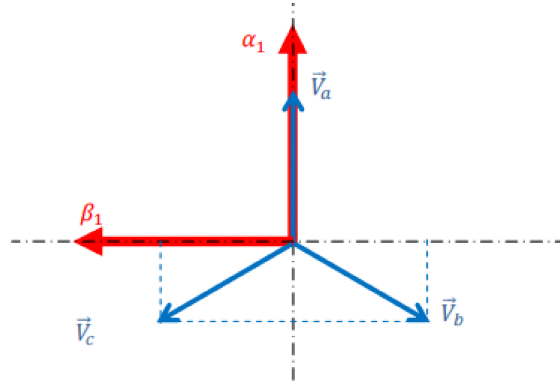


Figure 2.2: Orientation of the three-phase (a, b, c) system vectors in the (α_1, β_1) plane [3]

$$\begin{bmatrix} x_{\alpha 1} \\ x_{\beta 1} \\ x_{o1} \end{bmatrix} = C_1 \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (2.7)$$

The return to the three-phase system (the inverse transformation) is achieved by the transposed matrix C_1^T as follows:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = C_1^T \begin{bmatrix} x_{\alpha 1} \\ x_{\beta 1} \\ x_{o1} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} x_{\alpha 1} \\ x_{\beta 1} \\ x_{o1} \end{bmatrix} \quad (2.8)$$

• Concordia Transformation for the Second Star

Due to the angular offset between the first and second stars, a basic Concordia transformation alone is not sufficient to completely decouple the equations of the second star. It is therefore necessary to apply, in addition to the Concordia transformation, an additional rotation by an angle γ . This superimposed rotation ensures that the axes of the transformed coordinate system coincide with those of the coordinate system of the first star.

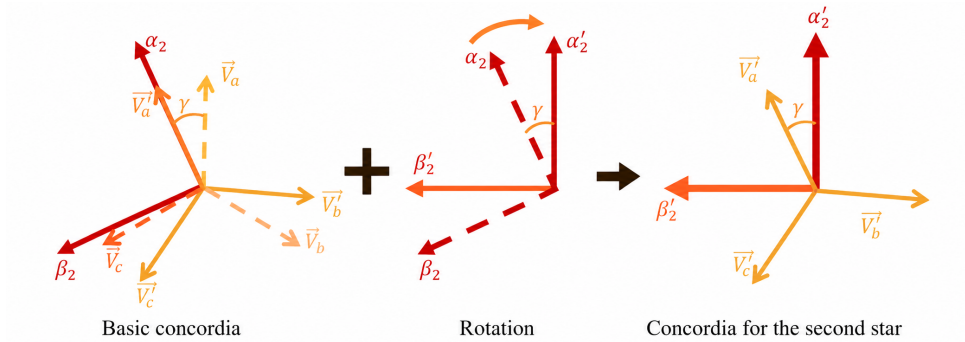


Figure 2.3: Concordia Transformation for the Second Star of the DSSG [3]

A specific orthogonal matrix C_2 is then used to transform the three-phase components (a', b', c') of the second star into the (α, β, o) coordinate system.

$$C_2 = \begin{bmatrix} \cos(\gamma) & \sin(\gamma) & 0 \\ -\sin(\gamma) & \cos(\gamma) & 0 \\ 0 & 0 & 1 \end{bmatrix} \times C_1 = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\gamma) & \cos\left(\gamma - \frac{2\pi}{3}\right) & \cos\left(\gamma - \frac{4\pi}{3}\right) \\ -\sin(\gamma) & -\sin\left(\gamma - \frac{2\pi}{3}\right) & -\sin\left(\gamma - \frac{4\pi}{3}\right) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (2.9)$$

Such as:

$$\begin{bmatrix} x_{\alpha' 2} \\ x_{\beta' 2} \\ x_{o2} \end{bmatrix} = C_2 \begin{bmatrix} x_{a'} \\ x_{b'} \\ x_{c'} \end{bmatrix} \quad (2.10)$$

The inverse transformation for the second star is performed using the transpose of the matrix, denoted C_2^T . Thus:

$$\begin{bmatrix} x_{a'} \\ x_{b'} \\ x_{c'} \end{bmatrix} = C_2^T \begin{bmatrix} x_{\alpha'2} \\ x_{\beta'2} \\ x_{o2} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\gamma) & -\sin(\gamma) & \frac{1}{\sqrt{2}} \\ \cos\left(\gamma - \frac{2\pi}{3}\right) & -\sin\left(\gamma - \frac{2\pi}{3}\right) & \frac{1}{\sqrt{2}} \\ \cos\left(\gamma - \frac{4\pi}{3}\right) & -\sin\left(\gamma - \frac{4\pi}{3}\right) & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} x_{\alpha'2} \\ x_{\beta'2} \\ x_{o2} \end{bmatrix} \quad (2.11)$$

2.3 Development of the machine model in the Concordia plane

Our goal is to develop the machine equations within Concordia space a vector space where the two three-phase systems are converted into simpler and decoupled equations.

2.3.1 Applying the Concordia transformation to the first star

With reference to equation (1.2) of the DSSG, the electrical equation for the first star in the three-phase system (a, b, c) is as follows:

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \frac{d}{dt} \begin{bmatrix} \Phi_{fa} \\ \Phi_{fb} \\ \Phi_{fc} \end{bmatrix} - R_s \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \begin{bmatrix} m_1 & m_2 & m_3 \\ m_3 & m_1 & m_2 \\ m_2 & m_3 & m_1 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix}$$

By substituting the vectors in the (a, b, c) coordinate system with their equivalents in the $(\alpha, \beta, 0)$ coordinate system and multiplying both sides by the Concordia matrix C_1 , the system simplifies as follows:

$$\begin{bmatrix} v_{\alpha 1} \\ v_{\beta 1} \\ v_{o1} \end{bmatrix} = \frac{d}{dt} \begin{bmatrix} \Phi_{f\alpha} \\ \Phi_{f\beta} \\ \Phi_{fo} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} - C_1 \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} C_1^T \frac{d}{dt} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} - C_1 \begin{bmatrix} m_1 & m_2 & m_3 \\ m_3 & m_1 & m_2 \\ m_2 & m_3 & m_1 \end{bmatrix} C_2^T \frac{d}{dt} \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} \quad (2.12)$$

After simplification

- The magnetic effects between the phases of the first star:

$$C_1 \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} C_1^T = \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} \end{bmatrix} \quad (2.13)$$

- The magnetic effect of the second star on the first star:

$$C_1 \begin{bmatrix} m_1 & m_2 & m_3 \\ m_3 & m_1 & m_2 \\ m_2 & m_3 & m_1 \end{bmatrix} C_2^T = \begin{bmatrix} \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (2.14)$$

Calculating the two inductance matrices within the Concordia coordinate system evidently produces two diagonal matrices. Thus, the system of electrical equations for the first star leads to decoupled equations in the $(\alpha, \beta, 0)$ coordinate system :

$$\begin{aligned} \begin{bmatrix} v_{\alpha 1} \\ v_{\beta 1} \\ v_{o1} \end{bmatrix} &= \frac{d}{dt} \begin{bmatrix} \Phi_{f\alpha 1} \\ \Phi_{f\beta 1} \\ \Phi_{fo1} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} - \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} \\ &\quad - \begin{bmatrix} \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} \end{aligned} \quad (2.15)$$

2.3.2 Applying the Concordia transformation to the second star

With reference to equation (1.2) of the DSSG, the electrical equation for the second star in the three-phase system (a', b', c') is as follows:

$$\begin{bmatrix} v_{a'} \\ v_{b'} \\ v_{c'} \end{bmatrix} = \frac{d}{dt} \begin{bmatrix} \Phi_{fa'} \\ \Phi_{fb'} \\ \Phi_{fc'} \end{bmatrix} - R_s \begin{bmatrix} i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix} - \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix} - \begin{bmatrix} m_1 & m_3 & m_2 \\ m_2 & m_1 & m_3 \\ m_3 & m_2 & m_1 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}$$

Transforming the quantities in this system into the $(\alpha, \beta, 0)$ frame and multiplying both sides of the equation by the Concordia matrix C_2 , we obtain the following simplified form:

$$\begin{aligned} \begin{bmatrix} v_{\alpha'2} \\ v_{\beta'2} \\ v_{o2} \end{bmatrix} &= \frac{d}{dt} \begin{bmatrix} \Phi_{f\alpha'2} \\ \Phi_{f\beta'2} \\ \Phi_{fo2} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} - C_2 \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} C_2^T \frac{d}{dt} \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} \\ &\quad - C_2 \begin{bmatrix} m_1 & m_3 & m_2 \\ m_2 & m_1 & m_3 \\ m_3 & m_2 & m_1 \end{bmatrix} C_1^T \frac{d}{dt} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} \end{aligned} \quad (2.16)$$

After simplification

- The magnetic effects between the phases of the second star:

$$C_2 \begin{bmatrix} L_s & m_s & m_s \\ m_s & L_s & m_s \\ m_s & m_s & L_s \end{bmatrix} C_2^T = \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} \end{bmatrix} \quad (2.17)$$

- The magnetic effect of the first star on the second star:

$$C_2 \begin{bmatrix} m_1 & m_3 & m_2 \\ m_2 & m_1 & m_3 \\ m_3 & m_2 & m_1 \end{bmatrix} C_1^T = \begin{bmatrix} \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (2.18)$$

Calculating the two inductance matrices within the Concordia coordinate system evidently produces two diagonal matrices. Thus, the system of electrical equations for the second star leads to decoupled equations in the $(\alpha, \beta, 0)$ coordinate system :

$$\begin{aligned} \begin{bmatrix} v_{\alpha'2} \\ v_{\beta'2} \\ v_{o2} \end{bmatrix} &= \frac{d}{dt} \begin{bmatrix} \Phi_{f\alpha'2} \\ \Phi_{f\beta'2} \\ \Phi_{fo2} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} - \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} \\ &\quad - \begin{bmatrix} \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha1} \\ i_{\beta1} \\ i_{o1} \end{bmatrix} \end{aligned} \quad (2.19)$$

By combining the equations of the two subsystems corresponding to the two stars, we obtain the overall mathematical model of the DSSG. This model is less complex than the original system (equation 1.2).

$$\begin{aligned} \begin{bmatrix} v_{\alpha1} \\ v_{\beta1} \\ v_{o1} \\ v_{\alpha'2} \\ v_{\beta'2} \\ v_{o2} \end{bmatrix} &= \begin{bmatrix} e_{\alpha1} \\ e_{\beta1} \\ e_{o1} \\ e_{\alpha'2} \\ e_{\beta'2} \\ e_{o2} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha1} \\ i_{\beta1} \\ i_{o1} \\ i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} - \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 & \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} & 0 & 0 & 0 \\ \frac{3}{2}M_{ss} & 0 & 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 & 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 & 0 & 0 & l_{fs} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha1} \\ i_{\beta1} \\ i_{o1} \\ i_{\alpha'2} \\ i_{\beta'2} \\ i_{o2} \end{bmatrix} \end{aligned} \quad (2.20)$$

Even though the axes specific to each star are uncoupled, the non-zero terms outside the diagonal blocks reveal the persistence of a magnetic coupling between the two stars. This means that the flux generated by one star influences the other, and vice versa.

2.4 Application for DSSG with sinusoidal flux distribution

Generators used in the industry are designed in a specific way to ensure the production of a balanced three-phase sinusoidal flux. With this configuration, the zero-sequence component is zero. This feature significantly simplifies the equations of magnetic flux, electromotive forces, and electromagnetic torque. In addition, the absence of the zero-sequence component makes the analysis and control of these systems much easier.

2.4.1 Calculation of the induced EMFs in the six phases in the $(\alpha, \beta, 0)$ coordinate system

The flux distribution along the air gap is assumed to be sinusoidal as a function of the position of the pole wheel relative to the axis of the coil of the first phase of the stator of the first star. This allows us to write:

$$\begin{bmatrix} \Phi_{fa} \\ \Phi_{fb} \\ \Phi_{fc} \\ \Phi_{fa'} \\ \Phi_{fb'} \\ \Phi_{fc'} \end{bmatrix} = \begin{bmatrix} \Phi_f \cos(p\theta) \\ \Phi_f \cos\left(p\theta - \frac{2\pi}{3}\right) \\ \Phi_f \cos\left(p\theta - \frac{4\pi}{3}\right) \\ \Phi_f \cos(p\theta + \gamma) \\ \Phi_f \cos\left(p\theta + \gamma - \frac{2\pi}{3}\right) \\ \Phi_f \cos\left(p\theta + \gamma - \frac{4\pi}{3}\right) \end{bmatrix} \quad (2.21)$$

Thus, the induced EMFs in the first and second stars are:

$$\begin{bmatrix} e_a \\ e_b \\ e_c \\ e_{a'} \\ e_{b'} \\ e_{c'} \end{bmatrix} = - \begin{bmatrix} \omega_s \Phi_f \sin(p\theta) \\ \omega_s \Phi_f \sin\left(p\theta - \frac{2\pi}{3}\right) \\ \omega_s \Phi_f \sin\left(p\theta - \frac{4\pi}{3}\right) \\ \omega_s \Phi_f \sin(p\theta + \gamma) \\ \omega_s \Phi_f \sin\left(p\theta + \gamma - \frac{2\pi}{3}\right) \\ \omega_s \Phi_f \sin\left(p\theta + \gamma - \frac{4\pi}{3}\right) \end{bmatrix} \quad (2.22)$$

For the first star, the components of the induced EMFs in the Concordia coordinate system are calculated as follows:

$$\begin{bmatrix} e_{\alpha 1} \\ e_{\beta 1} \\ e_{o1} \end{bmatrix} = C_1 \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} = -\sqrt{\frac{3}{2}} \omega_s \Phi_f \begin{bmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \end{bmatrix} \quad (2.23)$$

However, for the second star, the components of the induced EMFs in the Concordia coordinate system are calculated as follows:

$$\begin{bmatrix} e_{\alpha' 2} \\ e_{\beta' 2} \\ e_{o2} \end{bmatrix} = C_1 \begin{bmatrix} e_{a'} \\ e_{b'} \\ e_{c'} \end{bmatrix} = -\sqrt{\frac{3}{2}} \omega_s \Phi_f \begin{bmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \end{bmatrix} \quad (2.24)$$

2.4.2 Calculation of power and electromagnetic torque in the ' $\alpha\beta o'$ ' coordinate system

The instantaneous power delivered by the dual-star machine is:

$$p(t) = v_a i_a + v_b i_b + v_c i_c + v_{a'} i_{a'} + v_{b'} i_{b'} + v_{c'} i_{c'} = \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}^T \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} v_{a'} \\ v_{b'} \\ v_{c'} \end{bmatrix}^T \begin{bmatrix} i_{a'} \\ i_{b'} \\ i_{c'} \end{bmatrix} \quad (2.25)$$

Using the transformation to the Concordia coordinate system, this power becomes:

$$p(t) = \left(C_1^T \begin{bmatrix} v_{\alpha 1} \\ v_{\beta 1} \\ v_{o1} \end{bmatrix} \right)^T C_1^T \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} + \left(C_2^T \begin{bmatrix} v_{\alpha' 2} \\ v_{\beta' 2} \\ v_{o2} \end{bmatrix} \right)^T C_2^T \begin{bmatrix} i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} = \begin{bmatrix} v_{\alpha 1} \\ v_{\beta 1} \\ v_{o1} \end{bmatrix}^T \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \end{bmatrix} + \begin{bmatrix} v_{\alpha' 2} \\ v_{\beta' 2} \\ v_{o2} \end{bmatrix}^T \begin{bmatrix} i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} \quad (2.26)$$

This leads to a simplified expression, since the zero-sequence components are null.

$$p(t) = v_{\alpha 1} i_{\alpha 1} + v_{\beta 1} i_{\beta 1} + v_{\alpha' 2} i_{\alpha' 2} + v_{\beta' 2} i_{\beta' 2} \quad (2.27)$$

On the other hand, electromagnetic power is defined as:

$$P_e = \begin{bmatrix} e_{\alpha 1} \\ e_{\beta 1} \\ e_{o1} \\ e_{\alpha' 2} \\ e_{\beta' 2} \\ e_{o2} \end{bmatrix}^T \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} \quad (2.28)$$

The zero-sequence components are zero, hence

$$p_e = e_{\alpha 1} i_{\alpha 1} + e_{\beta 1} i_{\beta 1} + e_{\alpha' 2} i_{\alpha' 2} + e_{\beta' 2} i_{\beta' 2} \quad (2.29)$$

By multiplying the global equation of the six-phase synchronous generator by the current vector, we obtain:

$$\begin{bmatrix} e_{\alpha 1} \\ e_{\beta 1} \\ e_{o1} \\ e_{\alpha' 2} \\ e_{\beta' 2} \\ e_{o2} \end{bmatrix}^T \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} = p(t) + R_s \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix}^T \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix}^T \begin{bmatrix} l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 & \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 & \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & l_{fs} & 0 & 0 & 0 \\ \frac{3}{2}M_{ss} & 0 & 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 & 0 \\ 0 & \frac{3}{2}M_{ss} & 0 & 0 & l_{fs} + \frac{3}{2}M_{ss} & 0 \\ 0 & 0 & 0 & 0 & 0 & l_{fs} \end{bmatrix} \begin{bmatrix} i_{\alpha 1} \\ i_{\beta 1} \\ i_{o1} \\ i_{\alpha' 2} \\ i_{\beta' 2} \\ i_{o2} \end{bmatrix} \quad (2.30)$$

The electromagnetic power is equal to the sum of three terms. The first term represents the electrical power delivered to the load. The second term corresponds to the Joule losses dissipated in the stator winding resistances, while the third term represents the rate of

change of the magnetic energy stored in the double stator windings, which vanishes under steady-state conditions. Under balanced sinusoidal operation, both the instantaneous power and the electromagnetic power remain constant. The difference between them accounts solely for the Joule losses in the double stator windings.

$$P_e - P = p_{js} = R_s i_{\alpha 1}^2 + R_s i_{\beta 1}^2 + R_s i_{\alpha' 2}^2 + R_s i_{\beta' 2}^2 \quad (2.31)$$

The electromagnetic torque is expressed by the partial derivative of the electromagnetic energy with respect to the geometric angle of rotation of the rotating field as follows [7]

$$dE_e = P_e dt \quad (2.32)$$

$$T_e = \frac{dE_e}{d\theta} = \frac{P_e}{\Omega} = \frac{e_{\alpha 1} i_{\alpha 1} + e_{\beta 1} i_{\beta 1} + e_{\alpha' 2} i_{\alpha' 2} + e_{\beta' 2} i_{\beta' 2}}{\Omega}$$

Hence,

$$T_e = \frac{e_{\alpha 1}(i_{\alpha 1} + i_{\alpha' 2}) + e_{\beta 1}(i_{\beta 1} + i_{\beta' 2})}{\Omega} \quad (2.33)$$

2.5 DSSG model in the $(\alpha, \beta, z_1, z_2, z_3, z_4)$ coordinate system

Although the Concordia transformation simplifies the mathematical model of each three-phase star in the DSSG, further optimization is still possible. Therefore, an additional change of variables is applied. This transformation, which can be used for currents, fluxes, or voltages, decouples the coupled dynamics from the independent ones. As a result, the system equations are further simplified, which facilitates the design of advanced control strategies for the DSSG.

A common component along the α -axis and another along the β -axis are defined as follows:

$$\begin{cases} x_{\alpha} = \frac{1}{\sqrt{2}} (x_{\alpha 1} + x_{\alpha' 2}) \\ x_{\beta} = \frac{1}{\sqrt{2}} (x_{\beta 1} + x_{\beta' 2}) \end{cases} \quad (2.34)$$

In order to ensure the bijectivity of the transformation, a differential component along the α -axis and another along the β -axis are added, which are defined as follows:

$$\begin{cases} x_{z1} = \frac{1}{\sqrt{2}} (x_{\alpha 1} - x_{\alpha' 2}) \\ x_{z2} = \frac{1}{\sqrt{2}} (x_{\beta 1} - x_{\beta' 2}) \end{cases} \quad (2.35)$$

Whereas the two remaining components x_{z3} and x_{z4} are chosen equal to the two zero-sequence components previously defined in the Concordia transformation.

$$\begin{cases} x_{z3} = x_{o1} \\ x_{z2} = x_{o2} \end{cases} \quad (2.36)$$

Consequently, the expression of the electrical equations system of the DSSG in the $(\alpha\beta z_1 z_2 z_3 z_4)$ reference frame can be written as follows:

$$\begin{bmatrix} v_\alpha \\ v_\beta \\ v_{z1} \\ v_{z2} \\ v_{z3} \\ v_{z4} \end{bmatrix} = \begin{bmatrix} e_\alpha \\ e_\beta \\ e_{z1} \\ e_{z2} \\ e_{z3} \\ e_{z4} \end{bmatrix} - R_s \begin{bmatrix} i_\alpha \\ i_\beta \\ i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix} - \begin{bmatrix} l_{fs} + 3M_{ss} & 0 & 0 & 0 & 0 & 0 \\ 0 & l_{fs} + 3M_{ss} & 0 & 0 & 0 & 0 \\ 0 & 0 & l_{fs} & 0 & 0 & 0 \\ 0 & 0 & 0 & l_{fs} & 0 & 0 \\ 0 & 0 & 0 & 0 & l_{fs} & 0 \\ 0 & 0 & 0 & 0 & 0 & l_{fs} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_\alpha \\ i_\beta \\ i_{z1} \\ i_{z2} \\ i_{z3} \\ i_{z4} \end{bmatrix} \quad (2.37)$$

The electrical equations expressed in the $\alpha\beta z_1 z_2 z_3 z_4$ reference frame yield a significantly simplified and decoupled model of the DSSG. The transformation explicitly separates the torque-producing components (α and β axes) from the zero-sequence components (z_1, z_2, z_3, z_4 axes), which is a fundamental step toward the design and synthesis of control strategies, as this decoupling reduces controller complexity while improving dynamic performance and overall system efficiency.

The transformation also reveals four additional components whose inductance reduces to the sole leakage inductance l_{fs} . These components represent the zero-sequence modes and circulating currents inherent to the dual stator winding structure. They are primarily associated with non-sinusoidal current harmonics and asymmetry between the two star windings. In a well-designed and balanced system, the standard control objective is to minimize or eliminate these components, as they do not contribute to useful torque production but instead introduce additional copper losses.

2.5.1 Calculation of the α and β components of the EMFs in the $(\alpha, \beta, z_1, z_2, z_3, z_4)$ reference frame

The sinusoidal distribution of the flux along the air gap allows to derive the expressions of the EMFs of the DSSG

The components of the EMFs induced in the first star of the DSSG are:

$$\begin{pmatrix} e_{\alpha 1} \\ e_{\beta 1} \\ e_{o1} \end{pmatrix} = -\sqrt{\frac{3}{2}} \omega_s \Phi_f \begin{pmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \end{pmatrix} \quad (2.38)$$

Regarding the second star of the DSSG, the components of the EMFs are:

$$\begin{pmatrix} e_{\alpha' 2} \\ e_{\beta' 2} \\ e_{o2} \end{pmatrix} = -\sqrt{\frac{3}{2}} \omega_s \Phi_f \begin{pmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \end{pmatrix} \quad (2.39)$$

The EMFs in the $\alpha\beta$ reference frame are obtained by combining the contributions of both stars as follows:

$$\begin{cases} e_\alpha = \frac{1}{\sqrt{2}} (e_{\alpha 1} + e_{\alpha' 2}) \\ e_\beta = \frac{1}{\sqrt{2}} (e_{\beta 1} + e_{\beta' 2}) \end{cases} \quad (2.40)$$

These relations lead to the following expressions for the components:

$$\begin{pmatrix} e_\alpha \\ e_\beta \end{pmatrix} = -\sqrt{3} \omega_s \Phi_f \begin{pmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \end{pmatrix} \quad (2.41)$$

2.5.2 Calculation of the EMF components ' z_1 ' and ' z_2 '

The components e_{z_1} and e_{z_2} are defined from the EMFs of the two stars using equation (2.34), as follows:

$$\begin{cases} e_{z1} = \frac{1}{\sqrt{2}} (e_{\alpha 1} - e_{\alpha' 2}) \\ e_{z2} = \frac{1}{\sqrt{2}} (e_{\beta 1} - e_{\beta' 2}) \end{cases} \quad (2.42)$$

Given the equality of the components $e_{\alpha 1}$ and $e_{\alpha' 2}$, as well as $e_{\beta 1}$ and $e_{\beta' 2}$ between the two stars, it logically follows that these components take a zero value.

$$\begin{bmatrix} e_{z1} \\ e_{z2} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \quad (2.43)$$

Similarly, the zero-sequence components of the two stars, e_{o1} and e_{o2} , are also considered zero in this model, which leads to the nullity of the components e_{z3} and e_{z4} :

$$\begin{bmatrix} e_{z3} \\ e_{z4} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \quad (2.44)$$

Following the analysis of the EMF components, and given that e_{z1} , e_{z2} , e_{z3} and e_{z4} are zero, the overall EMF of the DSSG can be expressed in the following matrix, which highlights the dominant sinusoidal behavior of the machine:

$$\begin{pmatrix} e_\alpha \\ e_\beta \\ e_{z1} \\ e_{z2} \\ e_{z3} \\ e_{z4} \end{pmatrix} = -\sqrt{3} \omega_s \Phi_f \begin{pmatrix} \sin(p\theta) \\ -\cos(p\theta) \\ 0 \\ 0 \\ 0 \\ 0 \end{pmatrix} \quad (2.45)$$

This indicates that the system is perfectly symmetric and balanced in terms of induced EMFs, implying the absence of voltage potentials that could drive circulating currents between the two stars or zero-sequence currents in the system. This is a desirable condition for the stability and efficiency of the machine.

2.6 Model of the DSSG in the Park reference frame

The Park transformation converts the static components along the two axes α and β of the reference frame into a rotating Park reference frame dq , which is synchronized with the rotating field of the machine. This conversion simplifies the analysis and control of electrical machines by making the key quantities (state variables) continuous.

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = P_r(p\theta) \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = \begin{bmatrix} \cos(p\theta) & \sin(p\theta) \\ -\sin(p\theta) & \cos(p\theta) \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \quad (2.46)$$

While the return to the static components along the two axes α and β of the $\alpha\beta z_1 z_2 z_3 z_4$ reference frame (the inverse transformation) is ensured by the transpose of the matrix as follows:

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = P_r(\theta)^T \begin{bmatrix} x_d \\ x_q \end{bmatrix} = \begin{bmatrix} \cos(p\theta) & -\sin(p\theta) \\ \sin(p\theta) & \cos(p\theta) \end{bmatrix} \begin{bmatrix} x_d \\ x_q \end{bmatrix} \quad (2.47)$$

2.6.1 Application of the Park transformation to the DSSG model

Recalling the equations of the two components along the two axes α and β of the $\alpha\beta z_1 z_2 z_3 z_4$ reference frame:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} - R_s \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} - \begin{bmatrix} l_{fs} + 3M_{ss} & 0 \\ 0 & l_{fs} + 3M_{ss} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (2.48)$$

The transformation of the quantities from the stationary reference frame $\alpha\beta$ to the rotating reference frame dq is performed by applying the transpose of the Park matrix:

$$\begin{aligned} P_r^T(\theta) \begin{bmatrix} v_d \\ v_q \end{bmatrix} &= P_r^T(\theta) \begin{bmatrix} e_d \\ e_q \end{bmatrix} - R_s \times P_r^T(\theta) \begin{bmatrix} i_d \\ i_q \end{bmatrix} \\ &- \begin{bmatrix} l_{fs} + 3M_{ss} & 0 \\ 0 & l_{fs} + 3M_{ss} \end{bmatrix} \frac{d}{dt} \left(P_r^T(\theta) \begin{bmatrix} i_d \\ i_q \end{bmatrix} \right) \end{aligned} \quad (2.49)$$

After multiplying both sides by the Park matrix $P_r(p\theta)$, the system simplifies as follows:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} e_d \\ e_q \end{bmatrix} - R_s \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \begin{bmatrix} l_{fs} + 3M_{ss} & 0 \\ 0 & l_{fs} + 3M_{ss} \end{bmatrix} P_r(p\theta) \frac{d}{dt} \left(P_r^T(\theta) \begin{bmatrix} i_d \\ i_q \end{bmatrix} \right) \quad (2.50)$$

The decomposition of the derivative term is given by:

$$P_r(p\theta) \frac{d}{dt} \left(P_r^T(\theta) \begin{bmatrix} i_d \\ i_q \end{bmatrix} \right) = \begin{bmatrix} 0 & -\omega_s \\ \omega_s & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (2.51)$$

After simplification, we obtain:

$$\begin{aligned} \begin{bmatrix} v_d \\ v_q \end{bmatrix} &= \begin{bmatrix} e_d \\ e_q \end{bmatrix} - R_s \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \begin{bmatrix} 0 & \omega_s(l_{fs} + 3M_{ss}) \\ \omega_s(l_{fs} + 3M_{ss}) & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \\ &\quad - \begin{bmatrix} l_{fs} + 3M_{ss} & 0 \\ 0 & l_{fs} + 3M_{ss} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \end{aligned} \quad (2.52)$$

2.6.2 Calculation of the EMFs in the 'dq' reference frame

The calculation of the induced EMFs in the rotating synchronous 'dq' reference frame is a fundamental step, which greatly facilitates further studies on the behavior of the machine.

$$\begin{bmatrix} e_d \\ e_q \end{bmatrix} = \sqrt{6} \omega_s \Phi_f \begin{bmatrix} 0 \\ 1 \end{bmatrix} \quad (2.53)$$

In a dual star synchronous machine, the transformation of the EMFs from the static reference frame ' $\alpha\beta$ ' to the 'dq' reference frame shows that the direct component is zero.

2.6.3 Calculation of the powers and electromagnetic torque in the 'dq' reference frame

Recalling the equation of the instantaneous power involved in the dual star machine in the case where the zero-sequence components are zero:

$$p(t) = v_{\alpha 1} i_{\alpha 1} + v_{\beta 1} i_{\beta 1} + v_{\alpha' 2} i_{\alpha' 2} + v_{\beta' 2} i_{\beta' 2}$$

Knowing that:

$$v_{\alpha} i_{\alpha} + v_{\beta} i_{\beta} = \frac{1}{2} [(v_{\alpha 1} + v_{\alpha' 2})(i_{\alpha 1} + i_{\alpha' 2}) + (v_{\beta 1} + v_{\beta' 2})(i_{\beta 1} + i_{\beta' 2})] \quad (2.54)$$

And

$$v_{z1} i_{z1} + v_{z2} i_{z2} = \frac{1}{2} [(v_{\alpha 1} - v_{\alpha' 2})(i_{\alpha 1} - i_{\alpha' 2}) + (v_{\beta 1} - v_{\beta' 2})(i_{\beta 1} - i_{\beta' 2})] \quad (2.55)$$

This power can be expressed in the $\alpha\beta z_1 z_2 z_3 z_4$ reference frame as follows:

$$p(t) = v_{\alpha} i_{\alpha} + v_{\beta} i_{\beta} + v_{z1} i_{z1} + v_{z2} i_{z2} \quad (2.56)$$

In balanced sinusoidal operation, the d-axis of the Park reference frame is aligned with the flux of the pole wheel. This allows the flux components to be separated, leading to a

simple expression for the electromagnetic torque.

The electromagnetic power can be calculated in the same way as the instantaneous power:

$$P_e = e_\alpha i_\alpha + e_\beta i_\beta + e_{z1} i_{z1} + e_{z2} i_{z2} \quad (2.57)$$

Since e_{z1} and e_{z2} are null, the expression reduces to:

$$P_e = e_\alpha i_\alpha + e_\beta i_\beta \quad (2.58)$$

Applying the Park transformation to this equation leads to:

$$P_e = e_d i_d + e_q i_q \quad (2.59)$$

While the electromagnetic torque is:

$$T_e = \frac{e_d i_d + e_q i_q}{\Omega} \quad (2.60)$$

Given that the component e_d is zero according to equation (2.52), the electromagnetic torque in the Park reference frame is expressed as:

$$T_e = \frac{e_q i_q}{\Omega} \quad (2.61)$$

With :

$$e_q = 2\sqrt{\frac{3}{2}} \omega_s \Phi_f \quad (2.62)$$

2.7 Simulink Model of the DSSG

Our study will specifically address the case of electrical angular decoupling between the two stators with $\gamma = \pi/6$. The model of the DSSG is represented in the hexaphasic frame, which allows a dynamic representation of the machine. The system is modeled as a hexaphasic equilibrated system of EMFs in series with windings having mutual coupling.

Machine Parameters

Parameter	Value
Nominal rotation speed	400 RPM
Nominal phase current	6.7 A
Nominal power (per star)	2 kW
Number of pole pairs	6
Nominal frequency	40 Hz
EMF at nominal rotation speed	100 V
Phase resistance	$R_s = 1.4 \Omega$
Leakage inductance	$l_{fs} = 3 mH$
Mutual stator inductance at null angle	$M_{ss} = 1.4 mH$

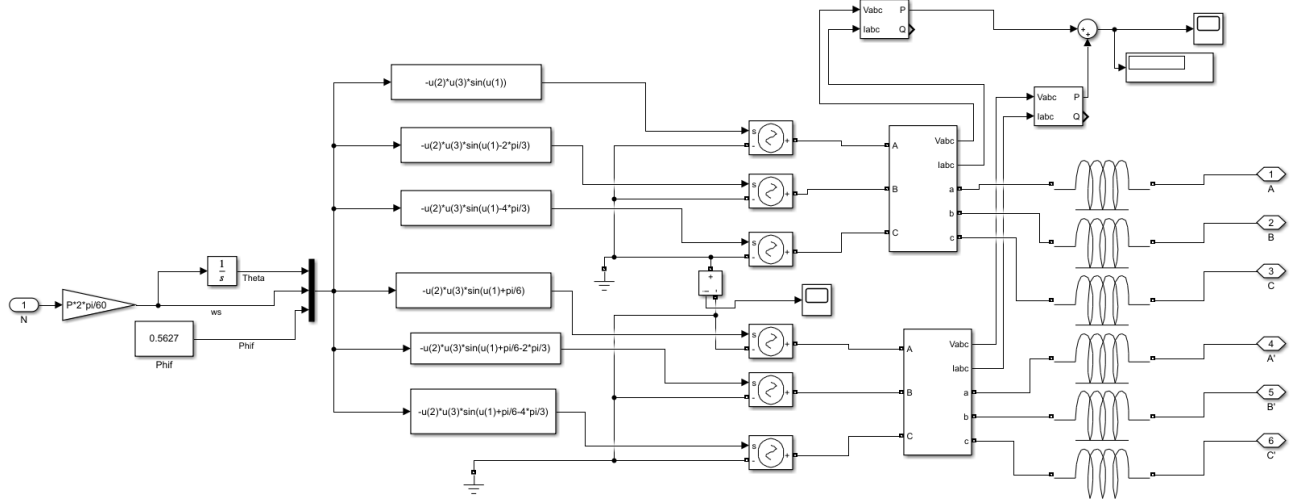


Figure 2.4: DSSG Simulink Model in MATLAB/Simscape

2.8 Simulation of Electrical Variables for Inductive Load

The figure below represents the Simulink model of the DSSG feeding an equilibrated inductive load ($10 \Omega, 8 mH$).

This operating point is characterized by conditions close to the nominal point of the generator. In addition to the model representation, we will also analyze the waveforms of voltages, EMFs, and power to validate the system performance.

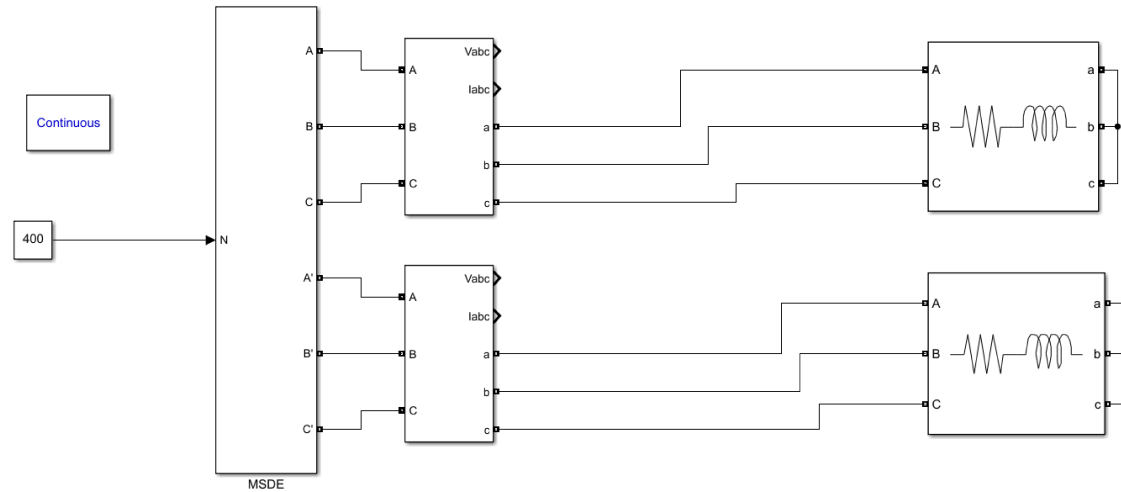


Figure 2.5: DSSG feeding an RL load

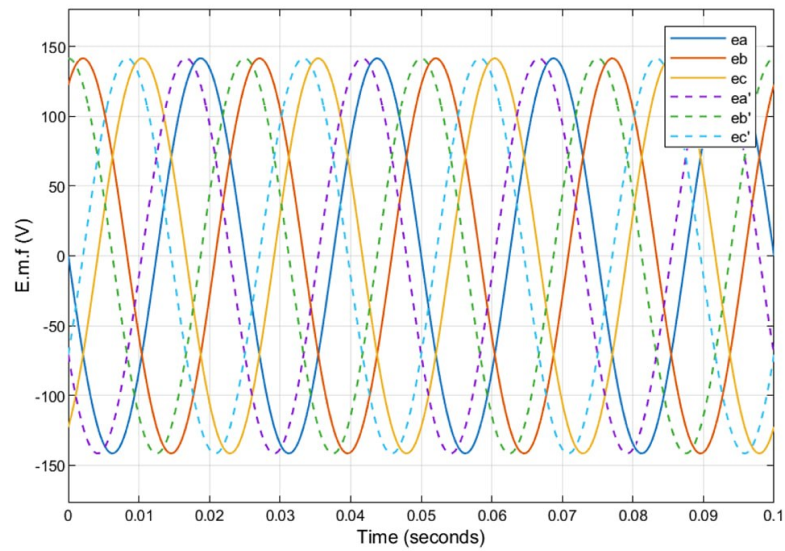


Figure 2.6: Induced EMF Waveforms

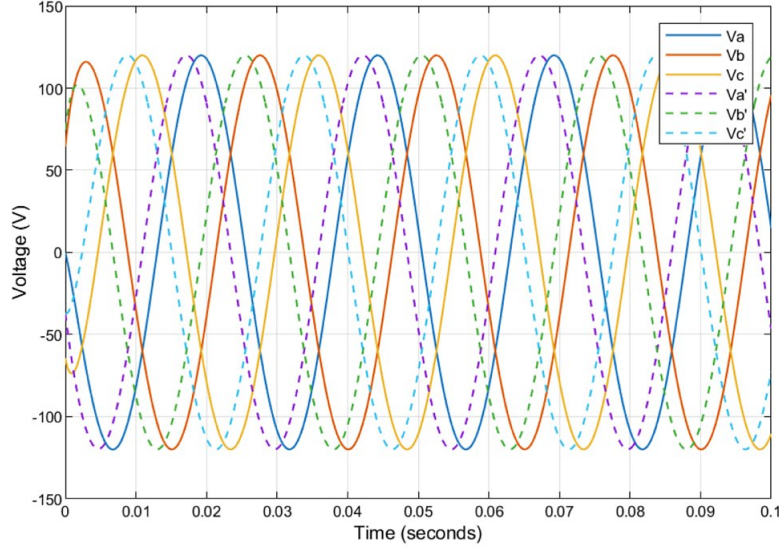


Figure 2.7: DSSG Voltage Waveforms with Inductive Load

Figures (2.6) and (2.7) show the induced EMFs and the corresponding phase voltages of the DSSG feeding the inductive load. The phase voltage is expressed as: $v = \text{EMF} - \text{Resistive drop} - \text{Inductive drop}$. The induced EMF ($d\Phi/dt$) represents the voltage generated by the rotor, the resistive drop ($R_s i$) accounts for losses in the stator windings, and the inductive drop represents the rate of change of flux linkage effects. As observed in the simulated waveforms, the phase voltage is lower than the induced EMF due to these resistive and inductive drops. This behavior validates Equation (2.1), which demonstrates the relationship between induced EMF and actual load voltage across the double stator windings.

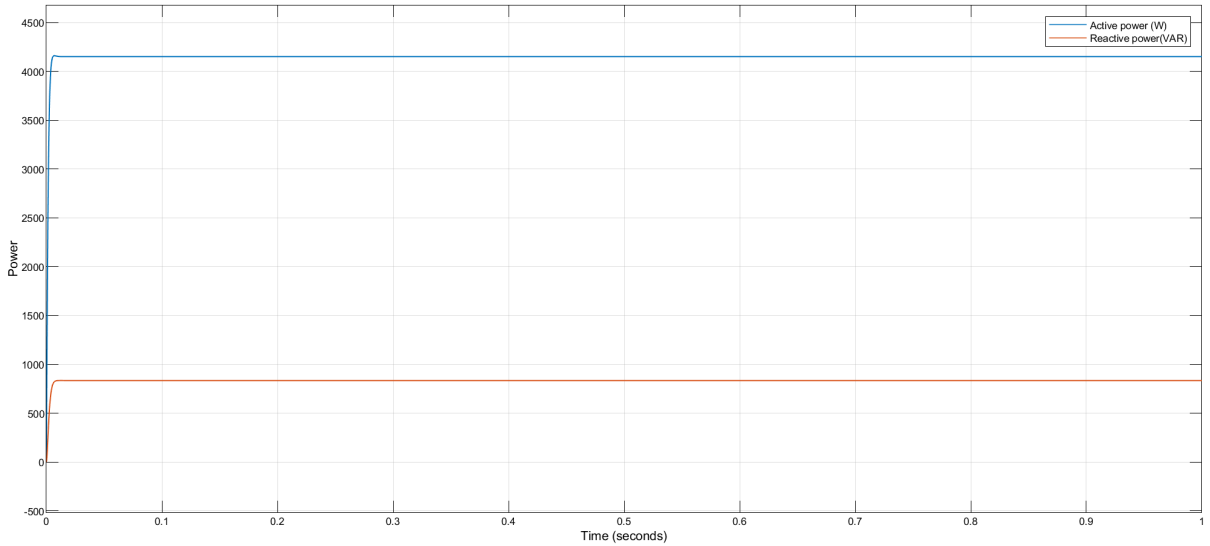


Figure 2.8: Behavior of Active and Reactive Power with RL Load

Figure (2.8) shows the behavior of transient and steady-state regimes of both active and reactive power of the DSSG feeding an RL load. After a very rapid transient regime, both powers stabilize at constant values (approximately 4150 W and 834 VAR), which validates stable and balanced operation of the DSSG.

2.9 PWM rectifiers

PWM rectifiers represent a major improvement compared to conventional diode-based rectifiers. They consist of a fully controllable semiconductor switch (controlled at both turn-on and turn-off) and a freewheeling diode. The switch is unidirectional in voltage and bidirectional in current. PWM rectifiers have the ability to instantaneously control the waveform of the current drawn from the AC source, which allows supplying any load (active or passive) with a regulated DC voltage from an AC voltage source, while ensuring that the absorbed current remains in phase with the corresponding AC source voltage. This allows achieving a power factor very close to unity. Furthermore, PWM rectifiers enable the regulation of reactive power flow through their control system, which represents a considerable advantage in terms of power quality management. Thanks to their adaptability to variations in load conditions and electrical grid disturbances, PWM rectifiers can be integrated into existing infrastructure without requiring significant modifications.[3]

2.9.1 Mathematical Modeling of the Three-Phase PWM Rectifier

The topology of the three-phase PWM rectifier is illustrated in the figure below. The input inductance is used as a filter, the AC side current is considered as a three-phase sinusoidal current, and a large capacitor is connected on the DC side to maintain a constant DC voltage.

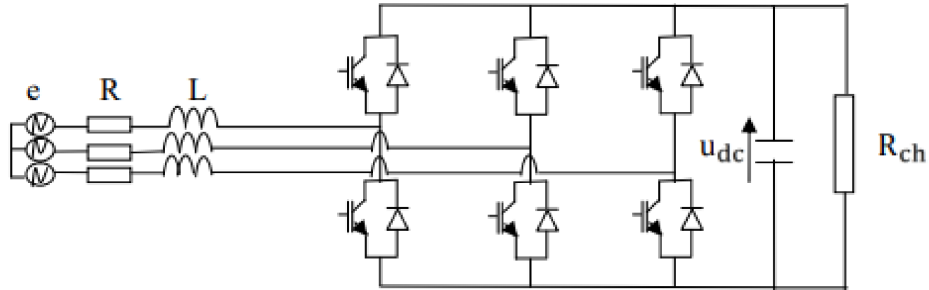


Figure 2.9: Topology of the Three-Phase PWM Rectifier[2]

The mathematical model of the PWM rectifier in the three-phase coordinate system is

$$\begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} = R \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + L \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \quad (2.61)$$

Since these quantities are time-varying, the design of the control system would be considerably complex. This is why it is necessary to transform the model from the abc frame to the rotating dq frame, which rotates at the same frequency as the AC source. The system is first transformed to a two-phase stationary frame $\alpha\beta$, under the hypothesis that the system is balanced, then the Park transformation is applied

$$P_r = \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix} \quad (2.62)$$

yielding

$$\begin{cases} e_d - V_d + L\omega i_q = Ri_d + L\frac{di_d}{dt} \\ e_q - V_q - L\omega i_d = Ri_q + L\frac{di_q}{dt} \end{cases} \quad (2.63)$$

It can be observed that the equations along the d and q axes exhibit a mutual coupling, which makes it impossible to control one variable independently without influencing the other. Therefore, a decoupling control strategy is required in order to control them independently. The feedforward decoupling control strategy is adopted here, which allows the regulation of currents by cancelling the coupling terms, which are added at the output of the controller in order to obtain the reference voltages required for regulation.

$$\begin{cases} Ri_d + L\frac{di_d}{dt} = V_{dr} \\ Ri_q + L\frac{di_q}{dt} = V_{qr} \end{cases} \quad (2.64)$$

The coupling voltages are

$$\begin{cases} V_{dc} = e_d + L\omega i_q \\ V_{qc} = e_q - L\omega i_d \end{cases} \quad (2.65)$$

therefore the reference voltages are

$$\begin{cases} V_d^* = V_{dc} - V_{dr} \\ V_q^* = V_{qc} - V_{qr} \end{cases} \quad (2.66)$$

Following the decoupling, the dynamics of i_d and i_q are governed by a first-order relation with their respective commands v_{dr} and v_{qr} , which allows the use of PI controllers, where the same PI parameters are applied to both current regulation loops.

2.10 PI Controller Design for Current and Voltage Regulation

2.10.1 Current PI Controller Design

The current controller plays a central role in the control of active and reactive power, by ensuring precise tracking of the reference currents i_d^* and i_q^* . Thus, the control law of the d -axis current can be expressed by the following equation:

$$Ri_d + L\frac{di_d}{dt} = \left(K_p + \frac{K_i}{s}\right)(i_d^* - i_d) \quad (2.67)$$

This equation is directly represented by the block diagram of the current regulation loop shown below:

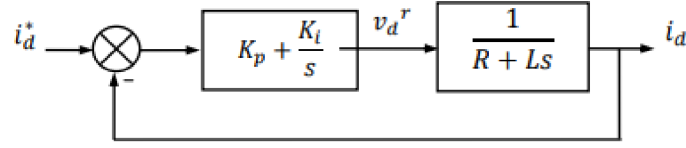


Figure 2.10: Current regulation loop[3]

The K_i and K_p parameters of the PI controller are calculated after establishing the open-loop transfer function and taking into consideration the RL filter time constant $\tau = L/R$. The dominant pole compensation method gives $K_p/K_i = \tau = L/R$, and the open-loop transfer function becomes $\frac{K_i}{Rs}$. After calculating the closed-loop transfer function and choosing the dynamics of the current regulation loop to be ten times faster than that of the RL filter, we obtain:

$$K_i = \frac{10R^2}{L}, \quad K_p = 10R \quad (2.68)$$

2.10.2 DC Bus Voltage Controller Design

The DC bus voltage U_{dc} regulation plays a crucial role in ensuring converter stability, by maintaining U_{dc} at its reference value U_{dc}^* despite load current fluctuations. This is achieved through a cascaded control approach, where the PI controller dynamically adjusts the phase current references to govern the active power transfer between the grid and the DC bus, while eliminating the steady-state error between the reference and the actual DC voltage.

The voltage regulation loop is illustrated by the block diagram below, where $\omega^* = (U_{dc}^*)^2$, $\omega = (U_{dc})^2$, and X is a variable representing the power difference between the input power of the rectifier and the load power. Under the hypothesis that switching losses, semiconductor conduction losses, and copper losses in the input filter are negligible, the process to be regulated is $\frac{2}{Cs}$. After calculating the open and closed-loop transfer functions and applying the pole placement method, the PI parameters are obtained as:

$$K_{iu} = \rho C, \quad K_{pu} = \rho^2 C \quad (2.69)$$

where ρ is chosen such that the inner loop (current regulation loop) is significantly faster than the outer loop (voltage regulation loop). In this study, ρ is selected as:

$$\rho = \frac{1}{4} \cdot \frac{R}{L} \quad (2.70)$$

Figure (2.11) represents the complete block diagram of the DC bus voltage regulation loop, which also provides a reference for the inner current regulation loop. The PWM rectifier is modeled as a unity gain system to simplify the design of this outer loop.

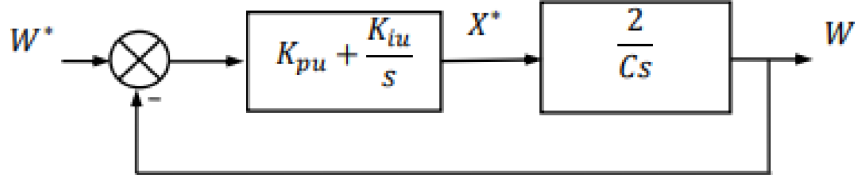


Figure 2.11: DC Bus Voltage Squared Regulation Loop[3]

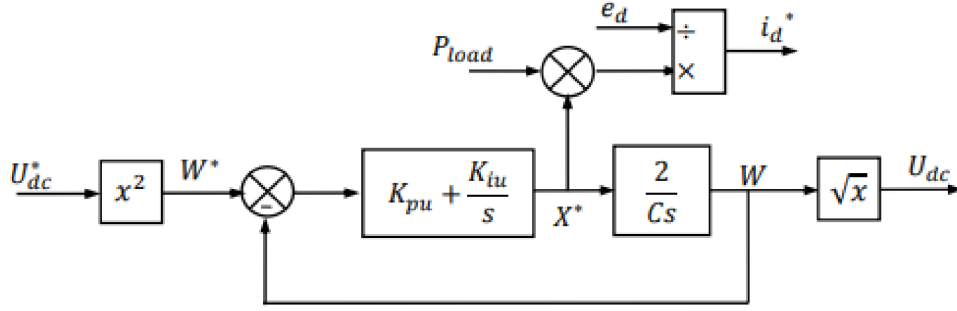


Figure 2.12: DC Bus Voltage Regulation Loop[3]

2.11 DSSG-PWM Rectifier Association with Capacitive DC Bus

The DSSG-PWM rectifier association presents some specificities compared to a conventional connection to an AC source. When the PWM rectifier is connected to a voltage source with negligible internal impedance, it is generally necessary to integrate an RL filter between the source and the rectifier to ensure harmonic filtration. However, in the case of the DSSG, this requirement is relaxed since the internal impedance of the machine inherently behaves as an RL filter, thus eliminating the need for an external filtering stage. [3]

Using the equations of the DSSG in the dq frame, which can be written as

$$\begin{cases} e_d - V_d + \omega_s (l_{fs} + 3M_{ss}) i_q = R_s i_d + (l_{fs} + 3M_{ss}) \frac{di_d}{dt} \\ e_q - V_q - \omega_s (l_{fs} + 3M_{ss}) i_d = R_s i_q + (l_{fs} + 3M_{ss}) \frac{di_q}{dt} \end{cases} \quad (2.71)$$

a fundamental similarity can be observed with equation (2.63). This analogy validates the use of the current and voltage control laws discussed earlier, where the control strategies in the dq frame are adapted to the DSSG. The parameters required for this adaptation are:

$$R = 1.4 \Omega, \quad L = l_{fs} + 3M_{ss} = 7.2 \text{ mH} \quad (2.72)$$

where R is the per-phase resistance of the DSSG, l_f is the leakage inductance, and M_{ss} is the mutual inductance between two stator windings at zero angular displacement.

In the Park rotating frame, aligning the d -axis with the supply voltage vector sets $e_q = 0$, where e_d represents the voltage vector amplitude. The PLL [8][9] achieves this

alignment by synchronizing the rotating dq frame with the voltage vector, which in the case of a three-phase RL filtered supply imposes $i_q^* = 0$.

In the DSSG case however, the e_d component is null (see equation(2.51)), which shifts the condition to $i_d^* = 0$, while i_q^* is imposed by the DC bus voltage regulation loop. Moreover, since the DSSG induced EMFs are not directly measurable at its terminals, a PLL [10] is used to estimate the rotor position angle required for the Park transformations.

Consequently, the new parameters of the PI controllers for both the current and voltage regulation loops become:

$$\rho = \frac{1}{4} \frac{R}{L} = 48.61, \quad K_{pu} = \rho C = 0.053 \quad \text{and} \quad K_{iu} = \rho^2 C = 2.6 \quad (2.73)$$

$$K_i = \frac{10R^2}{L} = 2722.22 \quad \text{and} \quad K_p = 10R = 14 \quad (2.74)$$

2.12 Simulation of the DSSG-PWM Rectifier System Feeding a Resistive Load via a Capacitive DC Bus

Figure 2.13 illustrates the Simulink/Simscape model of the energy conversion chain, comprising the DSSG, the PWM rectifier, and the capacitive DC bus feeding the resistive load.

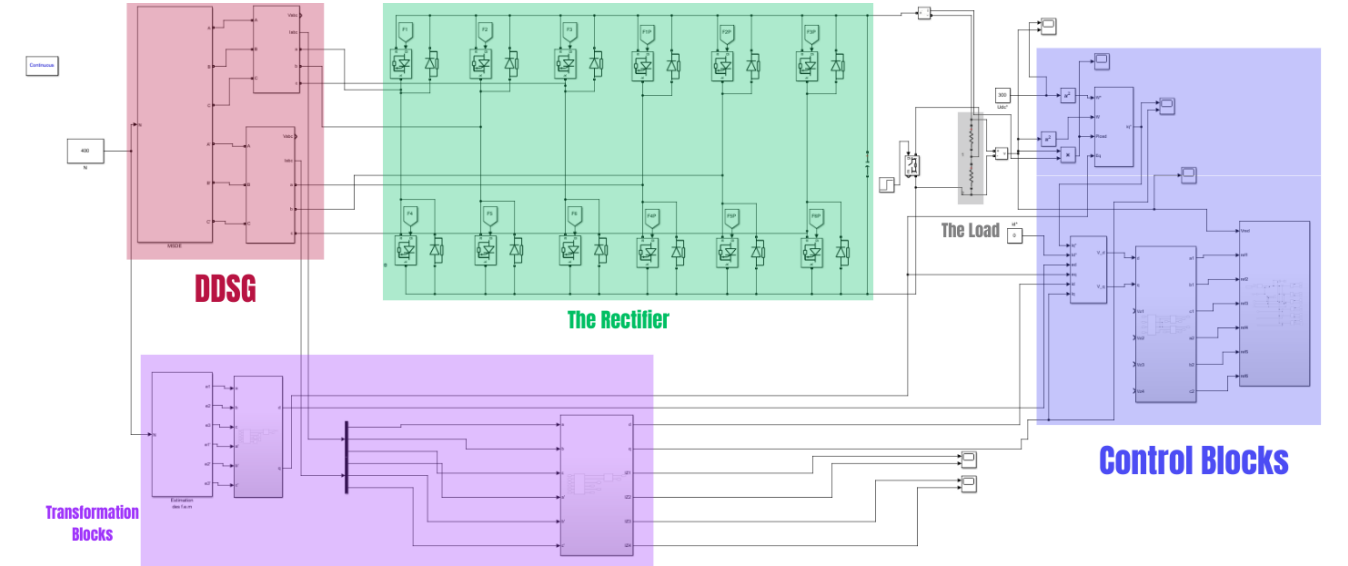


Figure 2.13: Simulink/Simscape model of the DSSG-PWM rectifier energy conversion chain

2.12.1 DC Bus Voltage Limitation

For proper operation of the PWM rectifier, the DC bus voltage U_{dc} must satisfy a minimum threshold condition. For a diode rectifier, this minimum voltage corresponds to the peak value of the line-to-line voltage [11]:

$$u_{dc_{min}} > \sqrt{2}v_{LL_{rms}} = \sqrt{2} \cdot \sqrt{3}v_{LN_{rms}} = \sqrt{3}E_{max} \quad (2.75)$$

For the PWM rectifier, an additional constraint is imposed:

$$u_{dc_{min}} \geq 2\sqrt{2} \cdot v_{LN_{rms}} = 2\sqrt{2} \cdot 100 = 282 \text{ V} \quad (2.76)$$

In this study, a DC bus voltage of $U_{dc}^* = 300 \text{ V}$ is chosen, which satisfies both constraints with sufficient margin.

2.12.2 Voltage and Current Waveforms

The simulation aims to validate the dimensioning of the PI controllers and to verify the capability of the PWM rectifier to instantaneously control the waveform of current drawn from the DSSG, which represents a major improvement compared to conventional diode-based rectifiers.

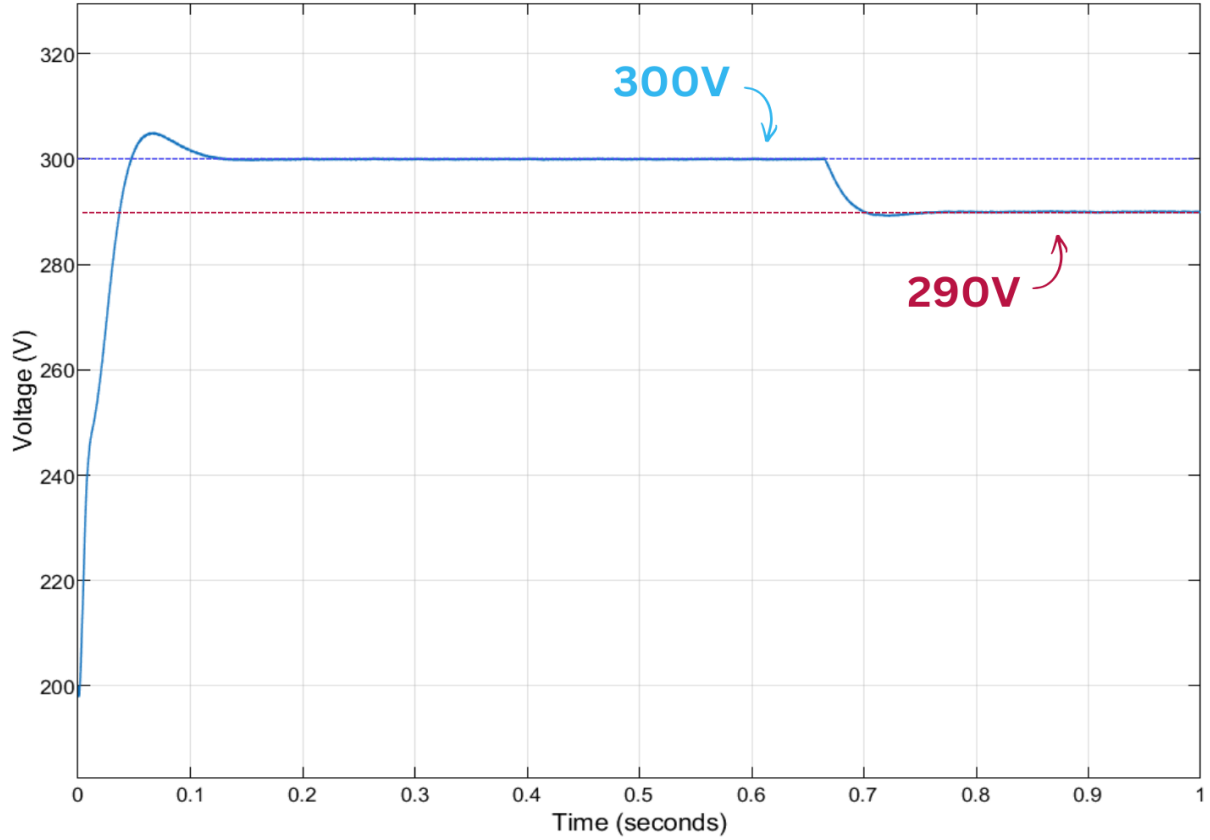


Figure 2.14: DC Bus Voltage Regulation

under two operating conditions. Initially, the voltage reference is set to 300 V, and the controller rapidly drives the DC bus voltage to the desired value with a small overshoot, after which the voltage settles quickly with negligible steady-state error. At 0.7s, the reference is changed from 300 V to 290 V. The DC bus voltage follows this new reference with a fast transient response, exhibiting a slight undershoot before converging smoothly to the new steady-state value of 290 V. The limited over/undershoot and short settling

time demonstrate the effectiveness of the PI controller in maintaining accurate DC bus voltage regulation and ensuring stable system operation under reference changes.

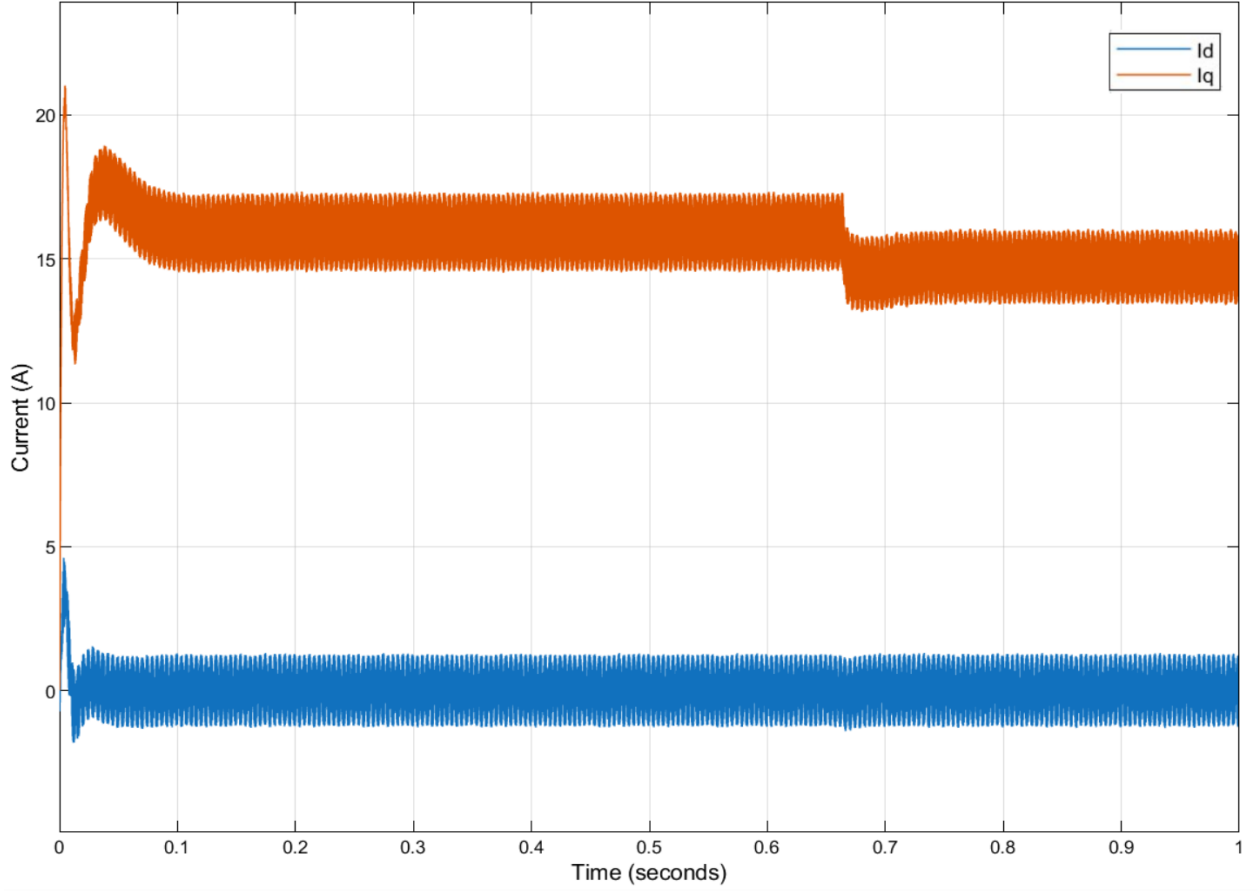
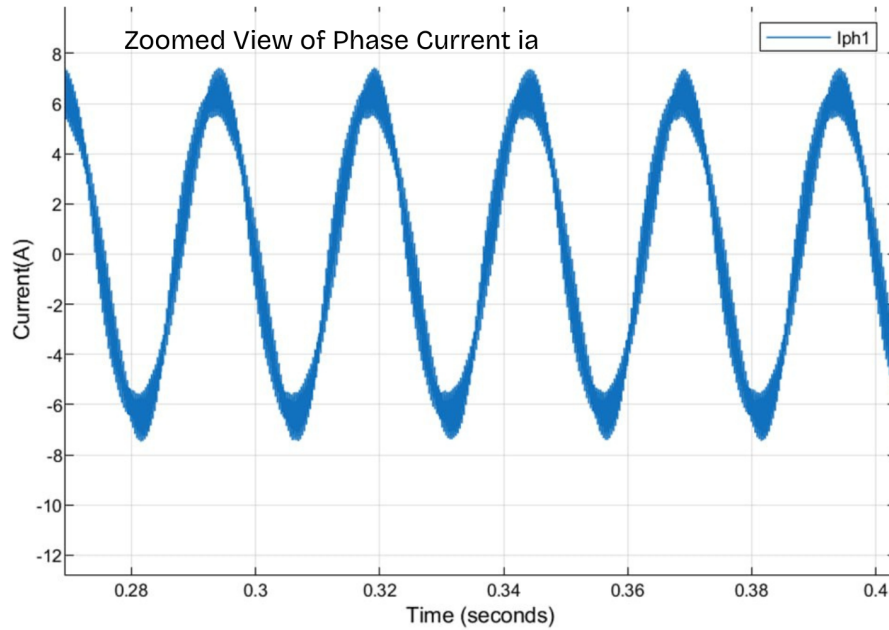


Figure 2.15: dq -Frame Currents

Figure 2.15 presents the evolution of the d -axis and q -axis currents, I_d and I_q , respectively. It can be observed that the d -axis current remains close to zero throughout the simulation, exhibiting only small switching-frequency ripples and brief transients during operating-point changes. This behavior confirms the effectiveness of the current control loop in maintaining proper field orientation. In contrast, the q -axis current tracks the active power demand and remains around 16–17 A during the first operating interval. When the DC bus voltage reference is reduced at $t \approx 0.7$ s, the q -axis current decreases to approximately 14–15 A, reflecting the lower power requirement of the system. The fast tracking capability and stable behavior of both current components demonstrate the good dynamic performance of the PWM rectifier control strategy, while the near-zero I_d current contributes to maintaining a high power factor and efficient power transfer.


 Figure 2.16: Phase current i_a

It can be observed from the figure above that the zoom on the phase current i_a shows in a clear way that the waveform of the phase current is quasi-sinusoidal. The superimposed high-frequency ripples are due to the commutation of the PWM, but the overall waveform is remarkably sinusoidal. This is one of the most significant improvements of the PWM rectifiers compared to conventional diode-based ones, as it ensures better current quality and reduced harmonic distortion.

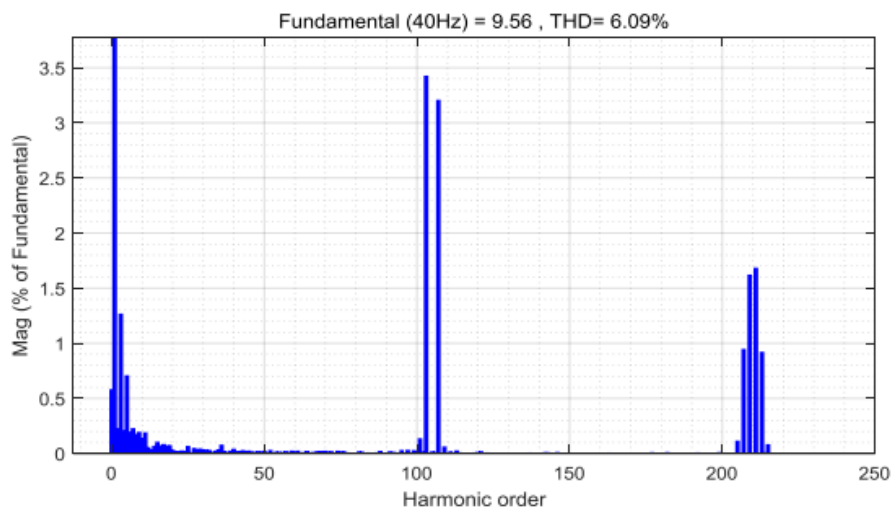


Figure 2.17: Harmonic Spectrum Analysis of Phase Current

The harmonic spectrum analysis of the phase current confirms the quasi-sinusoidal nature of the waveform. The fundamental frequency is observed at 40 Hz with a magnitude of 9.56, which represents the main sinusoidal component of the current. The Total Harmonic Distortion (THD) is measured at 6.09%, which indicates a good level of current

purity but confirms that the waveform is not perfectly sinusoidal due to the presence of harmonic components. The small peaks observed at higher harmonic orders (100, 200, etc.) correspond to the high-frequency ripples introduced by the PWM switching, which contribute to the overall distortion. This harmonic analysis validates that the PWM rectifier successfully produces a quasi-sinusoidal current from the DSSG with excellent power quality, demonstrating a major improvement over conventional diode-based rectifiers.

2.13 Conclusion

This chapter investigates the dynamic model of the DSSG-PWM rectifier energy conversion chain feeding a resistive load. The model of the DSSG was expressed in different bases (temporal, Concordia plane, Park plane, and $\alpha\beta z_1 z_2 z_3 z_4$), and we investigated the importance of these transformations in the analysis and control of polyphased machines. The application of the model to a dual-star synchronous machine with sinusoidal flux distribution in the inductor helped us to arrive at simple relations for EMFs, powers, and electromagnetic torque.

The simulation results confirmed the validation of the DSSG feeding a balanced RL load, which showed stable and equilibrated functioning in the steady-state regime. The EMFs and voltages exhibited two balanced three-phase systems in a diphasic configuration. We then investigated the PWM rectifier, its mathematical model, and its significant improvements compared to conventional diode-based rectifiers, thanks to its capacity for active control, which allows instantaneous control of the current waveform drawn from the DSSG, making it quasi-sinusoidal with significantly low THD.

PWM rectifiers, with the considerable advantages they bring in terms of power quality management and their adaptability to load variations and grid disturbances, make them the best solution for modern energy systems, especially in the field of renewable energy where precise control, bidirectionality in power flow, and low THD are crucial. While their complexity and higher cost are notable, these drawbacks are justifiable given their performance benefits.[1]

Chapter 3: Multi-Level Converters Applied In Energy Conversion Chains

3.1 Neutral Point Clamped Multilevel Inverter (NPC-MLI)

3.1.1 Introduction

Multilevel inverters have become widely used in modern power electronics due to their capability to generate high-quality voltage waveforms with reduced harmonic distortion. Compared to conventional two-level inverters, multilevel structures provide smoother output voltages and lower voltage stress across semiconductor devices, making them suitable for medium-voltage and high-power applications.

Among the different multilevel inverter topologies, the Neutral Point Clamped (NPC) inverter is considered one of the most important and commonly used structures because of its simple operating principle and good performance in industrial applications.

3.1.2 General Structure of the NPC Multilevel Inverter

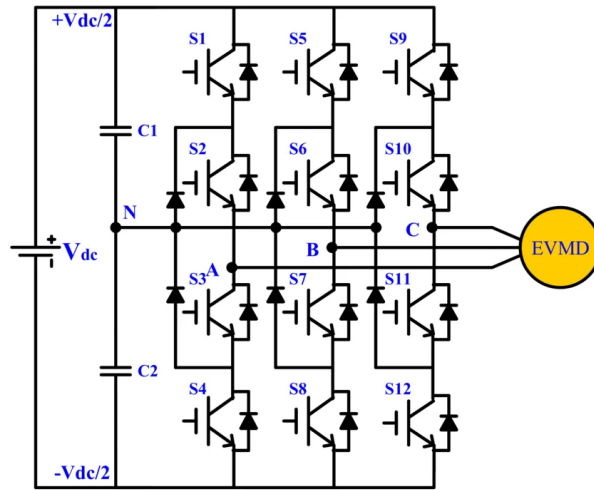


Figure 3.1: Three-level Neutral Point Clamped inverter[12]

The figure represents a three-level Neutral Point Clamped Multilevel Inverter (NPC-MLI). This topology belongs to the family of multilevel inverters, where the number of voltage levels can be increased depending on the application requirements. The inverter can therefore operate as a three-level, five-level, seven-level, or even higher-level converter.

The inverter is composed of three legs corresponding to the three phases. Each leg contains two arms:

- The upper arm
- The lower arm

Each arm contains a certain number of controlled switches. The number of switches is directly related to the number of output voltage levels generated by the inverter.

For a multilevel inverter, the number of voltage levels is generally linked to the number of carriers used in the modulation strategy according to the following relationship:

$$N_{levels} = N_{carriers} + 1 \quad (3.1)$$

For example:

- A three-level inverter requires two carriers
- A five-level inverter requires four carriers
- A seven-level inverter requires six carriers

The same logic is applied for most multilevel inverter topologies.

3.1.3 Carrier-Based PWM Control Strategy

Principle of Carrier Comparison

The control strategy used for multilevel inverters is based on the comparison between:

- A sinusoidal reference signal
- Multiple triangular carrier signals

The reference signal has:

- An amplitude equal to 1
- A frequency equal to 50 Hz

The carriers are distributed within the interval:

$$-1 \leq Carrier \leq 1 \quad (3.2)$$

This rule remains valid regardless of the number of carriers or the number of levels used.

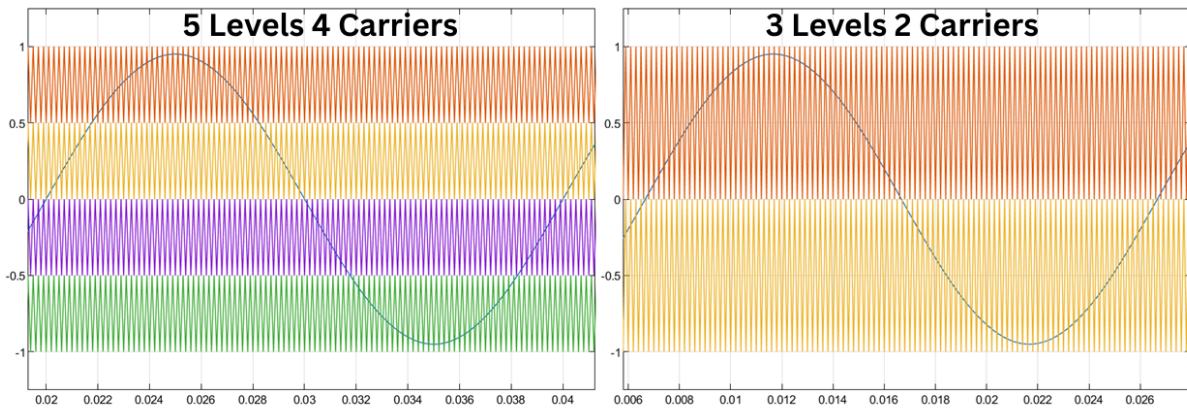


Figure 3.2: Three-level And Five-level Control Carriers

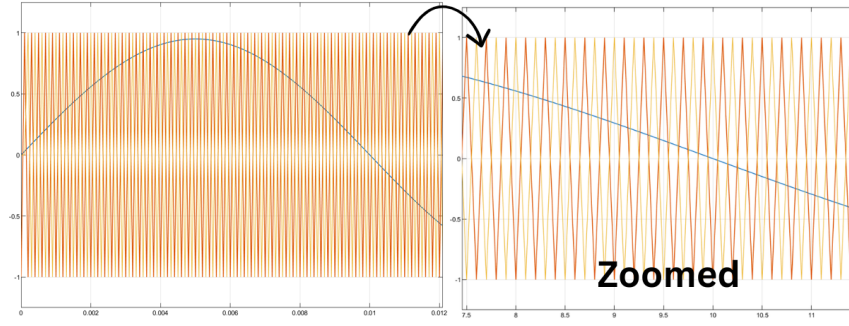


Figure 3.3: Phase Shifted Carriers

3.1.4 Simplified Control of the Three-Level NPC Inverter

The control of the three-level NPC inverter is based on comparing a sinusoidal reference signal with two level-shifted carriers, which divides operation into three switching regions.

When the reference is above the upper carrier, the inverter produces the positive voltage level. When it is below the lower carrier, the negative voltage level is generated. When the reference lies between the two carriers, the zero voltage level is produced.

Each region directly corresponds to a specific switching state of the power devices, forming the basis of the inverter switching logic.

3.1.5 Practical Representation of Power Switches

In simulations, each switch is modeled as a single ideal device for simplicity. In practice, however, high-voltage operation requires several semiconductor devices connected in series to withstand the applied stress.

These devices operate simultaneously under the same gating signal and share the total voltage across the switch. This voltage-sharing approach improves reliability and makes the system suitable for medium- and high-voltage applications.

It also increases robustness, as the stress is distributed across multiple devices, reducing the risk of failure and improving overall converter safety.

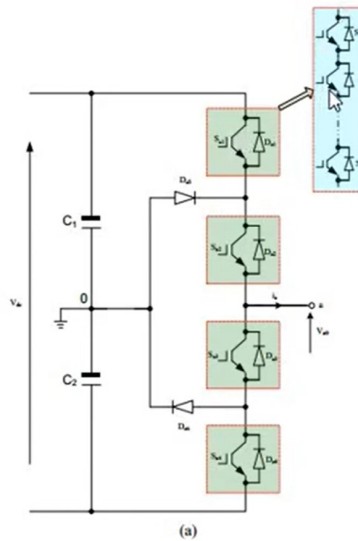


Figure 3.4: Cells Real Structure[17]

3.1.6 MATLAB/Simulink Implementation of the NPC Multilevel Inverter

The theoretical model of the three-level and five-level NPC inverter was implemented in MATLAB/Simulink in order to validate the modulation strategy and switching logic described previously.

The implementation process was carried out in a structured manner. First, a single inverter arm was designed, representing one phase leg of the converter. This arm includes the complete switching structure required for a three-level NPC configuration. After validating the operation of a single arm, the full phase leg was constructed. Finally, three identical legs were combined to form the complete three-phase inverter system.

Each leg corresponds to one phase of the system, and therefore the final model represents a three-phase multilevel inverter.

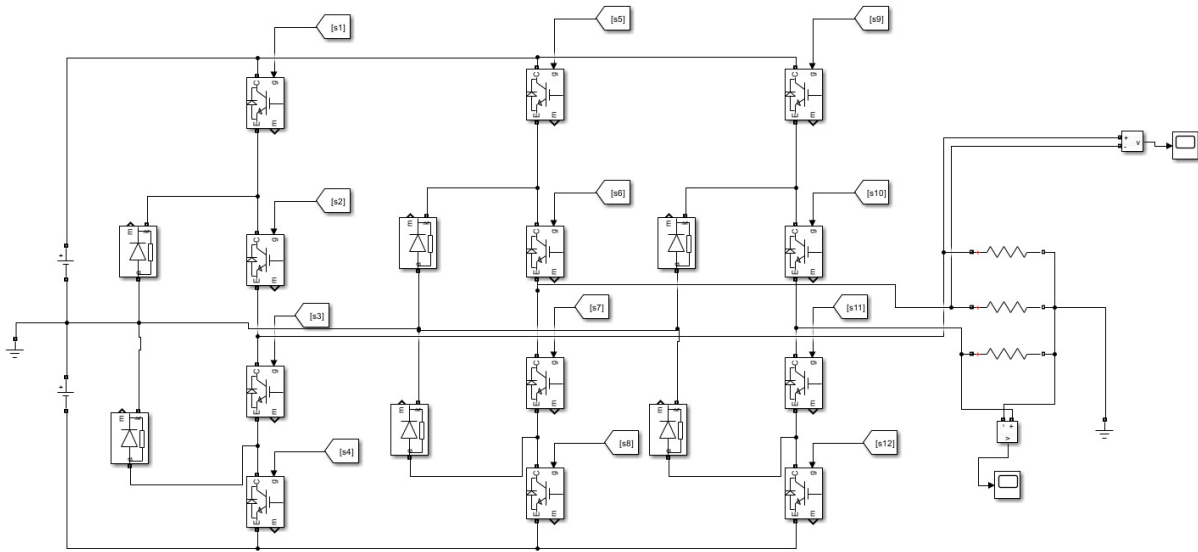


Figure 3.5: Three-level Neutral Point Clamped inverter

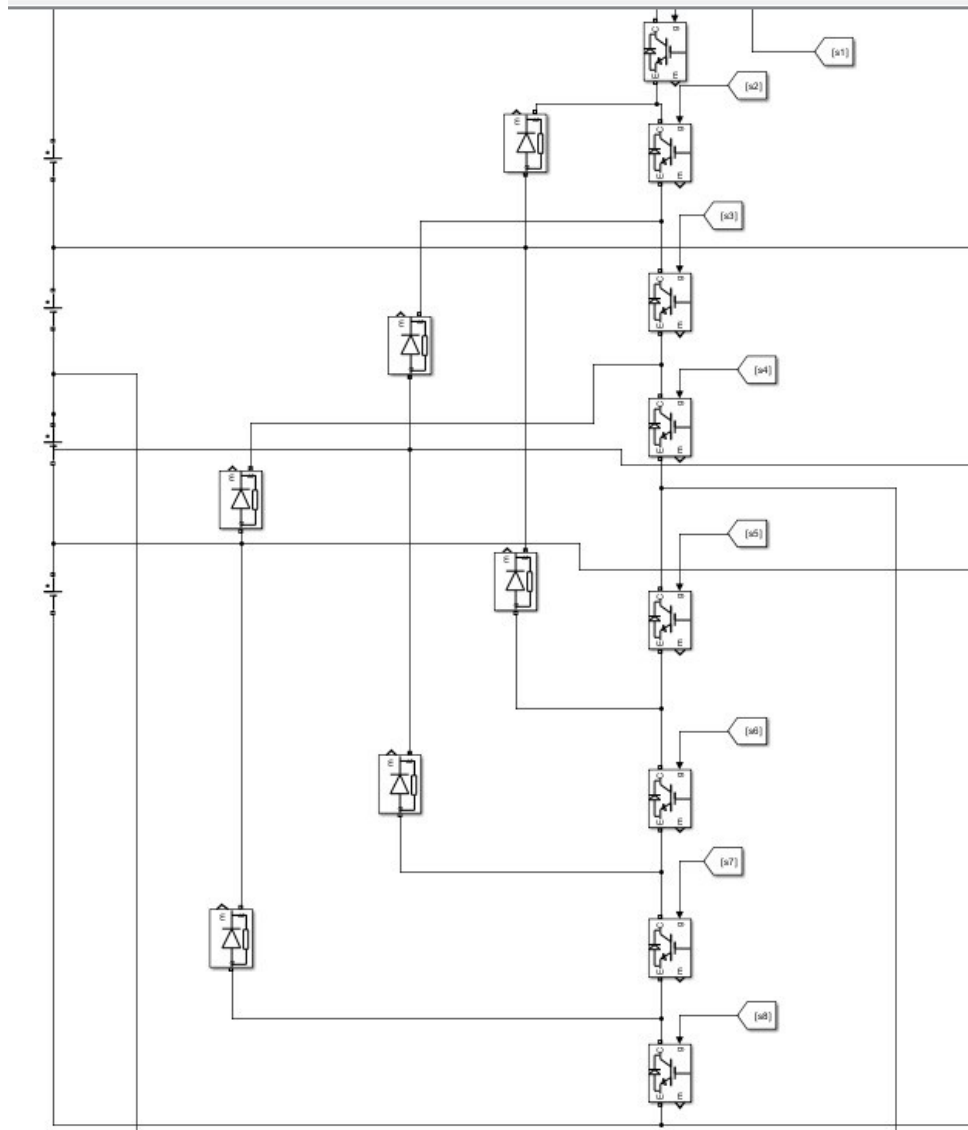


Figure 3.6: Five-level Neutral Point Clamped inverter

Control Strategy Implementation

The control block was developed based on a carrier-based PWM technique. For the three-level case, two carrier signals were used. The sinusoidal reference signals are compared continuously with these carriers in order to determine the switching states.

The switching logic is defined as follows:

- When the reference signal is greater than the upper carrier, the system operates in the positive region.
- When the reference signal is lower than the lower carrier, the system operates in the negative region.
- When the reference signal is between the two carriers, the zero state is generated.

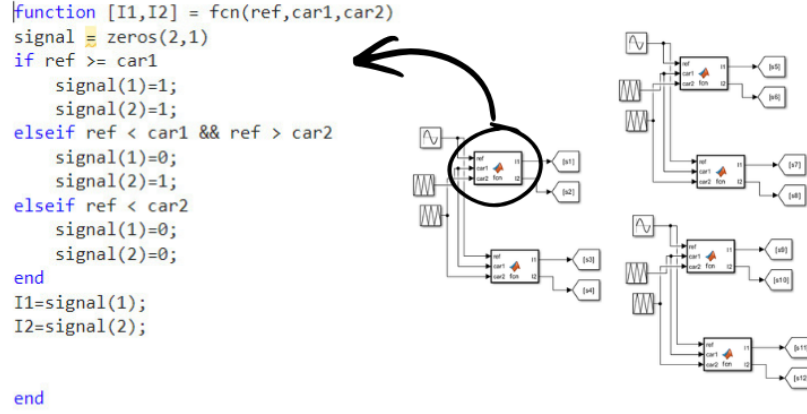


Figure 3.7: Control Code Upper Arm

This logic directly follows the switching table previously defined for the NPC inverter. The gating signals are then applied to the semiconductor switches to generate the required output voltage levels.

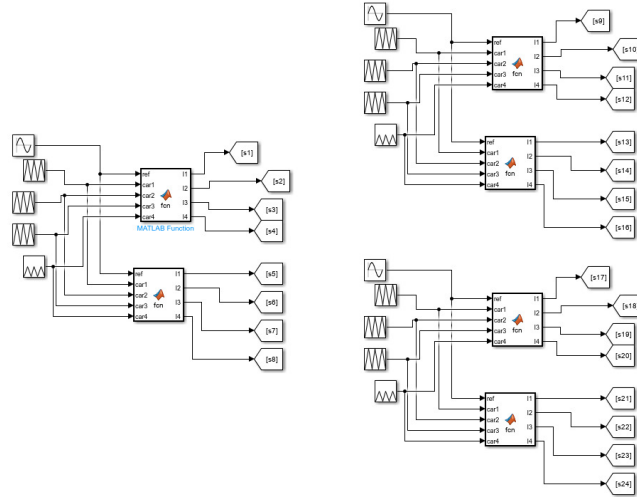


Figure 3.8: The Whole Control System

Three-Level Inverter Results

The implemented three-level inverter successfully generates the expected output voltage levels:

$$V_{out} \in \left\{ +\frac{V_{DC}}{2}, 0, -\frac{V_{DC}}{2} \right\} \quad (3.3)$$

The simulation results show that the output waveform closely follows the reference signal.

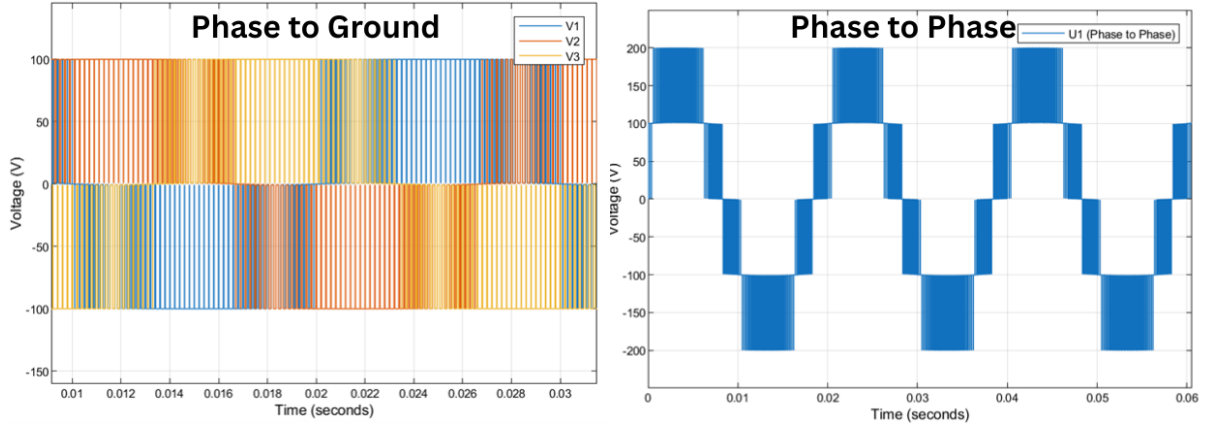


Figure 3.9: Three-level Neutral Point Clamped Results

The Total Harmonic Distortion (THD) obtained for the three-level configuration is approximately in the range of 58%.

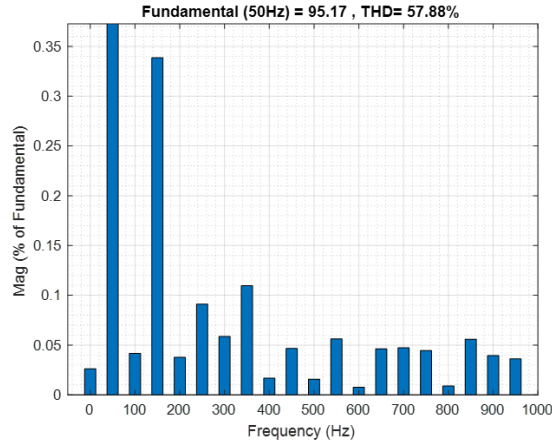


Figure 3.10: Three-level Neutral Point Clamped THD

Five-Level Inverter Extension

The same modeling approach was extended to a five-level NPC inverter. In this case, four carrier signals are required, which increases the complexity of the control block.

The switching logic remains identical in principle, but with finer voltage steps:

$$V_{out} \in \{+V_{DC}/2, +V_{DC}/4, 0, -V_{DC}/4, -V_{DC}/2\} \quad (3.4)$$

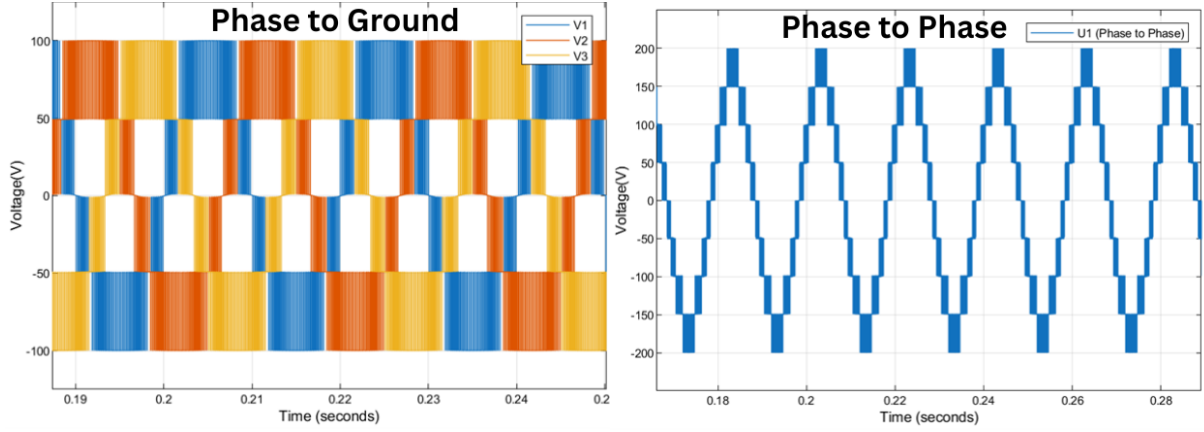


Figure 3.11: Five-level Neutral Point Clamped Results

The increase in the number of voltage levels significantly improves the quality of the output waveform. The Total Harmonic Distortion is reduced to approximately 30%, demonstrating the advantage of increasing the number of levels.

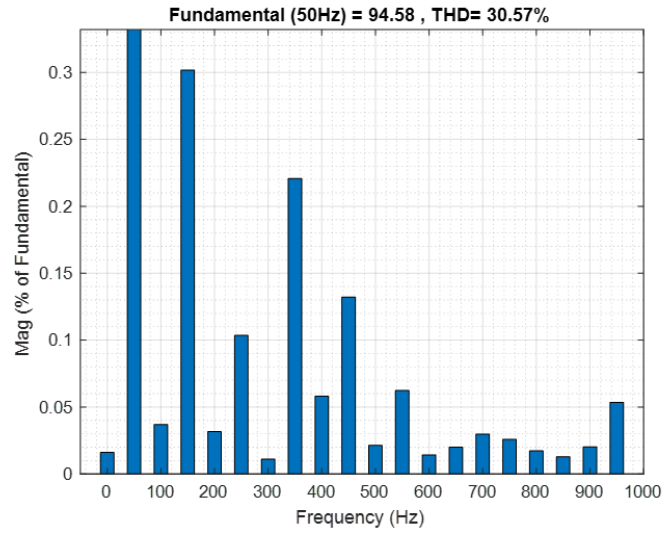


Figure 3.12: Five-level Neutral Point Clamped THD

3.1.7 Phase Displacement Between Control Signals

To generate a balanced three-phase system, three sinusoidal reference signals were used. Each reference signal has the same amplitude and frequency but is phase-shifted by: 120 degrees

This phase displacement ensures the correct generation of a three-phase output system and allows proper operation of the inverter under balanced load conditions.

3.1.8 Discussion

The obtained results demonstrate the effectiveness of the NPC multilevel inverter in generating high-quality output voltages from a DC source. Increasing the number of levels from three to five significantly improved the waveform quality and reduced the THD from approximately 58

The study also showed that one of the main advantages of the NPC topology is its simple control strategy. The switching signals are generated using a straightforward carrier-comparison method, making the controller easy to implement and validate in MATLAB/Simulink.

However, the improvement in harmonic performance is accompanied by an increase in converter complexity. Higher-level structures require additional switches, carrier signals, and more complex switching logic.

Overall, the NPC inverter offers a good compromise between control simplicity and harmonic performance, especially when supplied by a stable DC source.

3.1.9 Conclusion

This work compared two modulation strategies applied to the Neutral Point Clamped (NPC) multilevel inverter: Level-Shifted Carrier (LSC) and Phase-Shifted Carrier (PSC). The results showed that both techniques perform similarly, with only minor differences in waveform quality and harmonic content.

The study also analyzed the impact of increasing the number of voltage levels from three to five. The results clearly showed that higher-level inverters significantly improve output waveform quality and reduce Total Harmonic Distortion (THD), confirming the advantage of multilevel structures.

However, this improvement comes at the cost of increased system complexity. Moving from three to five levels requires more switches, more carrier signals, and a more complex control structure, which makes implementation more demanding.

Despite this, the NPC topology remains attractive due to its relatively simple control principle based on carrier comparison and switching logic tables. This makes it practical for implementation while still achieving good harmonic performance.

Overall, the NPC multilevel inverter provides a balanced trade-off between control simplicity and output power quality, particularly for medium- and high-power DC-AC conversion applications.

3.2 Flying Capacitor Multilevel Inverter (FCMLI)

The Flying Capacitor Multilevel Inverter (FCMLI) is a multilevel converter topology designed to produce multiple voltage levels and improve output waveform quality by reducing harmonic distortion. Unlike the Neutral Point Clamped (NPC) inverter, it uses flying capacitors instead of clamping diodes to generate intermediate voltage levels and distribute voltage stress.

This structure improves voltage sharing among semiconductor devices and allows flexible voltage level generation. However, it introduces additional complexity in capacitor voltage balancing, especially as the number of levels increases. Consequently, higher-level implementations require more capacitors and more advanced control strategies.

Figure 3.6 illustrates the general structure of a three-phase Flying Capacitor Multilevel Inverter.

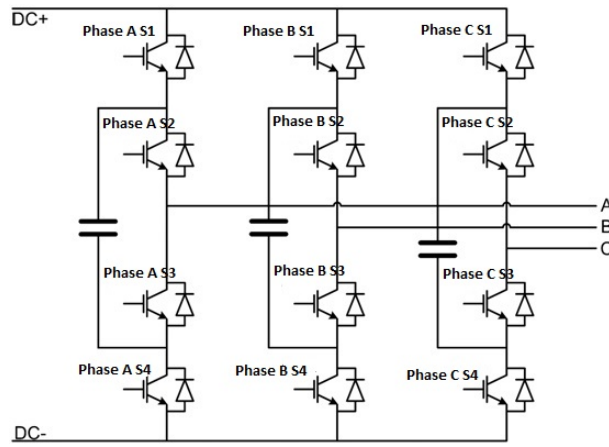


Figure 3.1: Three-phase Flying Capacitor Multilevel Inverter topology[14]

In the presented structure, each phase leg contains several controlled semiconductor switches associated with flying capacitors. These capacitors allow the converter to produce multiple voltage levels while sharing the voltage stress across the switching devices.

The Flying Capacitor Multilevel Inverter (FCMLI) is structurally close to the Neutral Point Clamped (NPC) inverter in terms of its goal: generating multiple voltage levels from a DC source to improve output waveform quality and reduce harmonic distortion. The main difference lies in the way intermediate voltage levels are achieved. Instead of using clamping diodes as in the NPC topology, the FCMLI replaces them with flying capacitors placed within each phase leg. These capacitors are not simply passive replacements; they actively participate in shaping the voltage levels by storing and redistributing energy during switching transitions.

The number of flying capacitors increases with the number of output voltage levels. In general, for an n -level converter, the number of required capacitors per phase leg is $n - 2$. For example, a three-level FCMLI requires one flying capacitor per phase leg, while a five-level structure requires three capacitors per leg. In addition, these capacitors are also used to split or share the input DC bus voltage across multiple intermediate levels. As the number of levels increases, both the number of capacitors and the number of possible switching states increase significantly.

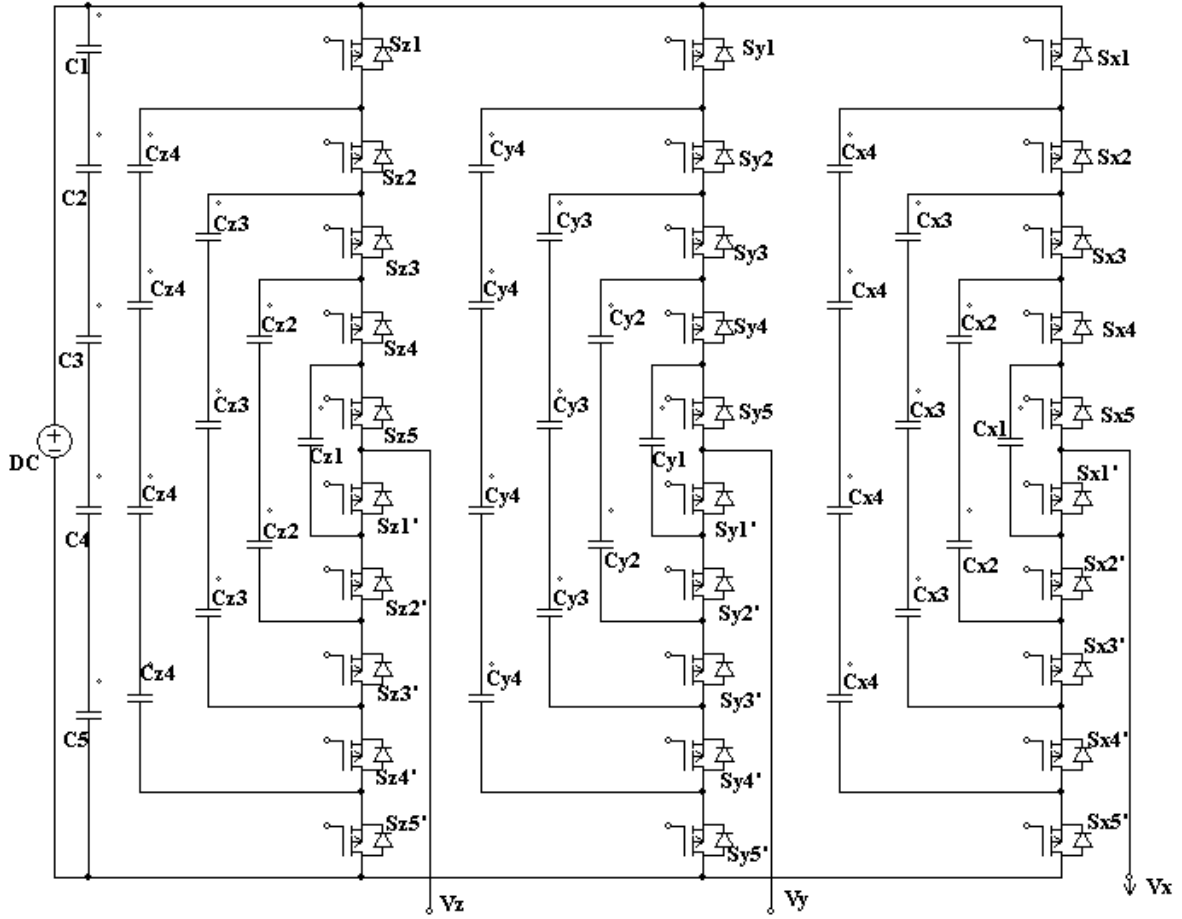


Figure 3.2: Five-phase Flying Capacitor Multilevel Inverter[15]

The main challenge of the FCMLI topology is the balancing of flying capacitor voltages. Unlike the NPC inverter, where the neutral point is fixed, the FCMLI relies on capacitors whose voltages vary dynamically and must be continuously regulated. Their charging and discharging depend on the switching states and current direction, meaning the converter behavior is influenced not only by the switching pattern but also by the internal capacitor energy.

For a given output voltage level, multiple redundant switching states may exist. These states are not equivalent, as each one affects the capacitor voltage differently. The control strategy must therefore select appropriate switching combinations to maintain voltage balance while still producing the desired output waveform.

This complexity increases significantly with higher voltage levels, as both the number of capacitors and redundant states grow. For this reason, FCMLI structures become more difficult to control in high-level applications. In a three-level configuration, even the zero-voltage state can be generated using different switching combinations, which must be managed carefully to avoid long-term capacitor imbalance.

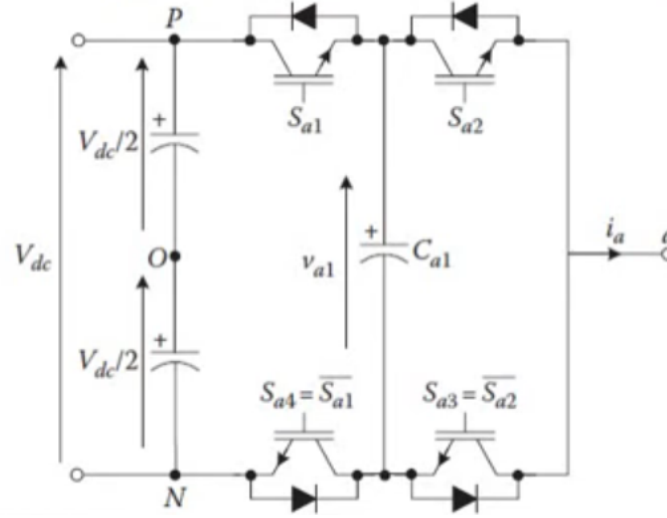


Figure 3.3: Three-phase Flying Capacitor Multilevel Inverter topology

S_{a1}	S_{a2}	S_{a3}	S_{a4}	Pole voltage, V_{a0}
1	1	0	0	$V_{dc}/2$
1	0	1	0	0
0	1	0	1	0
0	0	1	1	$-V_{dc}/2$

Figure 3.4: Three-phase Flying Capacitor Multilevel Inverter Truth Table

In the following implementation, the three-level FCMLI is derived from the NPC structure by replacing the clamping diodes with flying capacitors. The modulation framework, including the carrier-based approach and reference signals, remains similar. However, the switching logic is modified to include capacitor voltage balancing considerations, ensuring stable operation of the converter while preserving the same multicarrier PWM structure used in the NPC case.

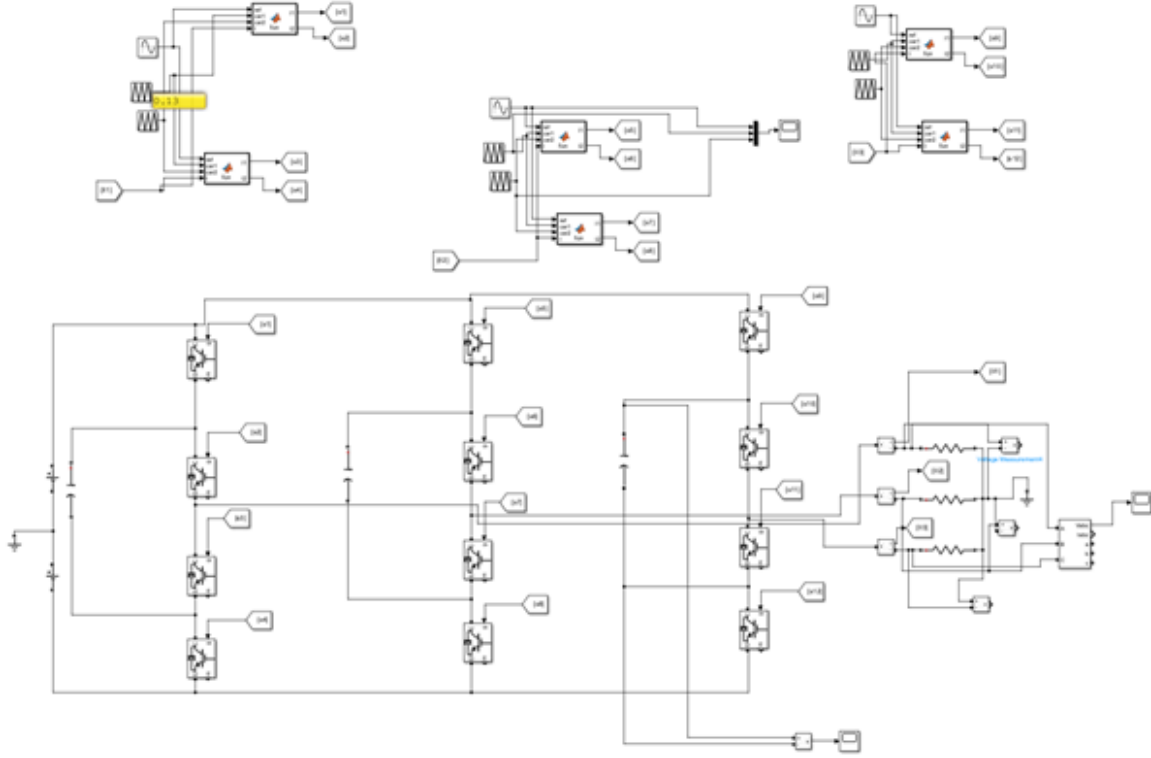


Figure 3.5: Three-phase Flying Capacitor Multilevel Inverter Simulink

The switching strategy is based on comparing the reference signal V_{ref} with two carrier signals Crr_1 and Crr_2 , while also considering the phase current i when the reference lies between the carriers.

For the upper arm, extreme regions set both switches ON or OFF depending on whether the reference is above or below the carriers. In the intermediate region, the switching state depends on the current direction to ensure proper capacitor charging and discharging.

The lower arm operates in a complementary manner, with inverted switching states and the same current-dependent rule in the intermediate region.

This combined logic is applied to each phase of the FCMLI, ensuring correct multilevel voltage generation while maintaining flying capacitor voltage balance.

```

function [I1,I2] = fcn(ref,car1,car2,I)
signal = zeros(2,1)
if ref > car1
    signal(1)=1;
    signal(2)=1;
elseif ref < car1 && ref > car2
    if I > 0
        signal(1)=1;
        signal(2)=0;
    elseif I < 0
        signal(1)=0;
        signal(2)=1;
    end
elseif ref < car2
    signal(1)=0;
    signal(2)=0;
end
I1=signal(1);
I2=signal(2);

end

```

Upper Arm

```

function [I1,I2] = fcn(ref,car1,car2,I)
signal = zeros(2,1)
if ref > car1
    signal(1)=0;
    signal(2)=0;
elseif ref < car1 && ref > car2
    if I < 0
        signal(1)=0;
        signal(2)=1;
    elseif I > 0
        signal(1)=1;
        signal(2)=0;
    end
elseif ref < car2
    signal(1)=1;
    signal(2)=1;
end
I1=signal(1);
I2=signal(2);

end

```

Lower ARM

Figure 3.6: Three-phase Flying Capacitor Multilevel Inverter Control Codes

At the steady-state operating condition, the inverter successfully generates the expected three voltage levels:

$$+\frac{V_{DC}}{2}, \quad 0, \quad -\frac{V_{DC}}{2}$$

In addition, the flying capacitor voltage remains properly balanced around its reference value of 100 V. The charging and discharging process of the capacitor is therefore effectively controlled by the switching algorithm, ensuring stable converter operation and proper generation of the multilevel waveform.

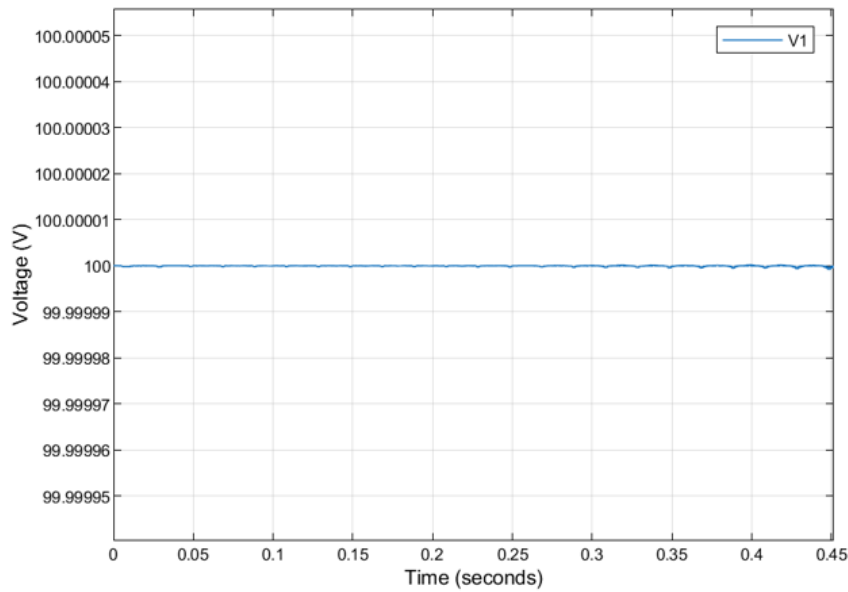


Figure 3.7: Three-level FCMLI flying capacitor voltage

We can now visualize the three output phases together. As shown in Figure 3.12, the three phase voltages are perfectly phase-shifted with respect to one another by 120° , ensuring balanced three-phase operation of the inverter.

The obtained waveform confirms the proper operation of the three-level FCMLI and the correct synchronization between the three phases. In addition, the Total Harmonic Distortion (THD) measured from the output voltage is approximately 58.08%, as illustrated in the corresponding FFT analysis figure.

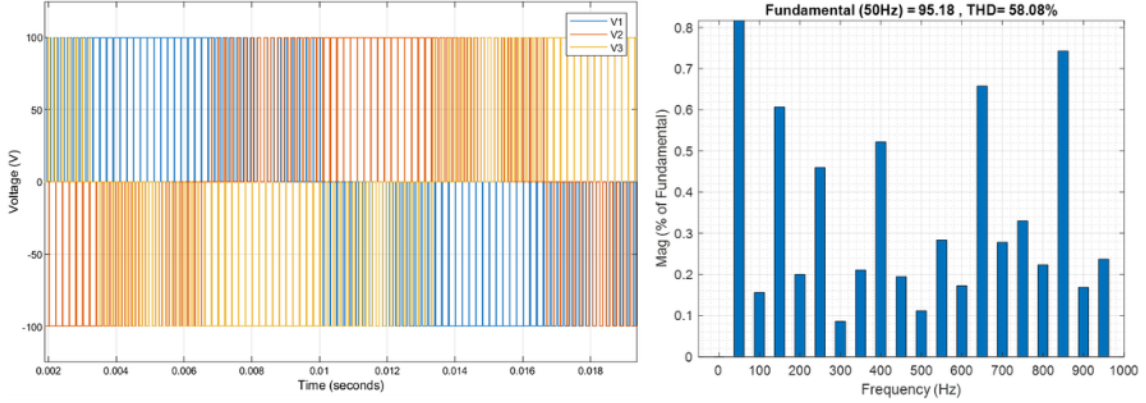


Figure 3.8: Three-phase output voltages and THD of the three-level FCMLI

After validating the three-level FCMLI, the study is extended to the five-level configuration. Increasing the number of levels improves waveform quality but also increases the complexity of both the switching strategy and capacitor voltage balancing.

This section focuses on generating the five voltage levels, analyzing the added control complexity, and comparing the THD performance with the three-level case to evaluate the effectiveness and trade-offs of the FCMLI topology.

3.2.1 Five-Level Flying Capacitor Multilevel Inverter

The five-level FCMLI is developed by extending the previously studied NPC-based structure. The main modification consists of replacing the clamping diodes with flying capacitors, which define the FCMLI topology.

The number of flying capacitors is determined by the relation:

$$N_{fc} = N_{level} - 2$$

For a five-level inverter, this gives:

$$N_{fc} = 3$$

meaning that each phase leg requires three flying capacitors to generate the required voltage levels and ensure proper operation.

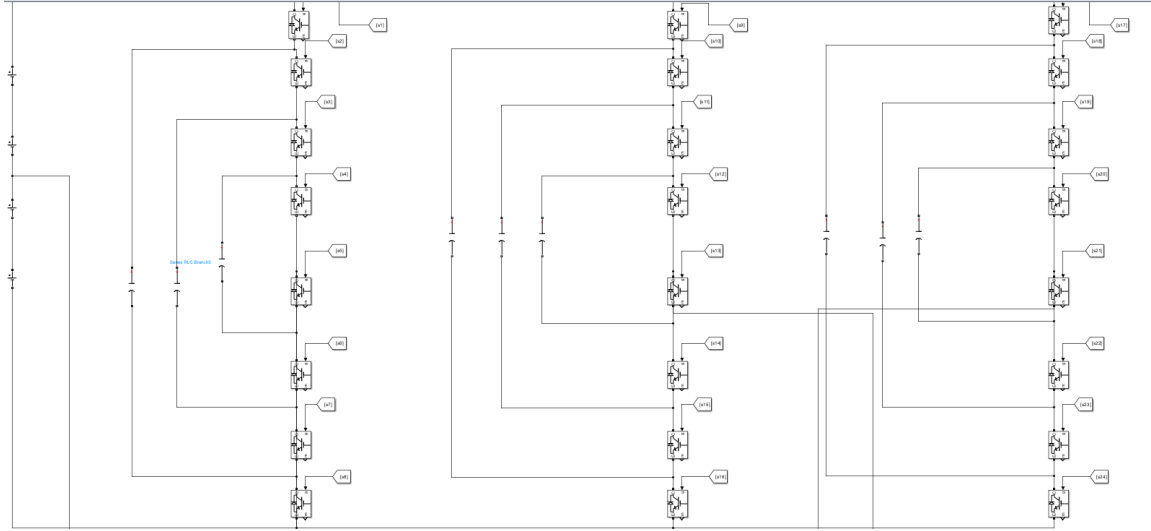


Figure 3.9: Five-levels Flying Capacitor Multilevel Inverter Simulink

Regarding the control method, the general modulation principle remains unchanged. Three sinusoidal reference signals are used, phase-shifted by 120° from one another in order to generate a balanced three-phase system. In addition, four carrier signals are required because the number of carriers is equal to the number of levels minus one:

$$N_{carrier} = N_{level} - 1$$

Thus:

$$N_{carrier} = 5 - 1 = 4$$

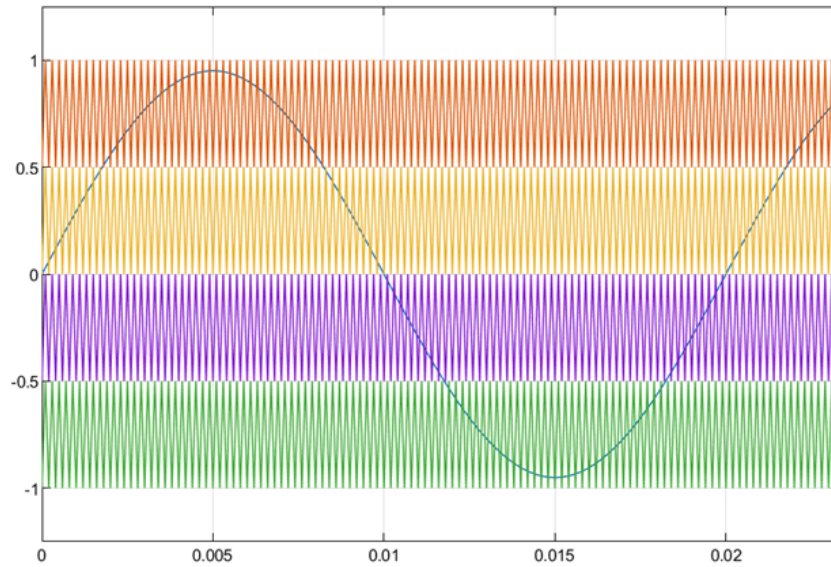


Figure 3.10: Five-levels Control Carriers

The obtained simulation results confirm the successful generation of the five output voltage levels. The phase-to-phase voltage waveform clearly presents the expected mul-

tilelevel staircase shape. In addition, the three output phases can be visualized simultaneously in the same scope, showing proper synchronization and the expected 120° phase shift between phases.

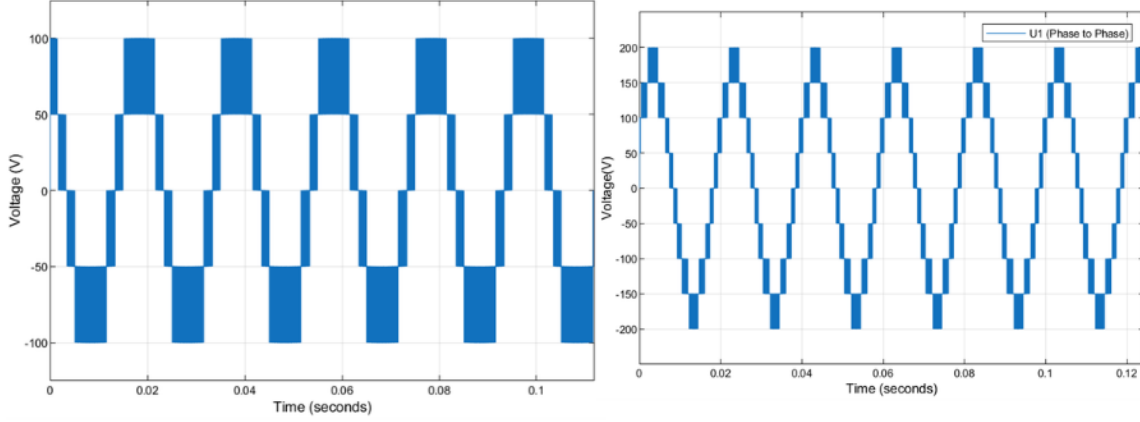


Figure 3.11: Output voltages of the Five-level FCMLI

The FFT analysis also shows that the Total Harmonic Distortion (THD) decreases significantly compared to the three-level configuration. The obtained THD value is approximately 30.95%, which confirms the improvement in waveform quality when increasing the number of voltage levels.

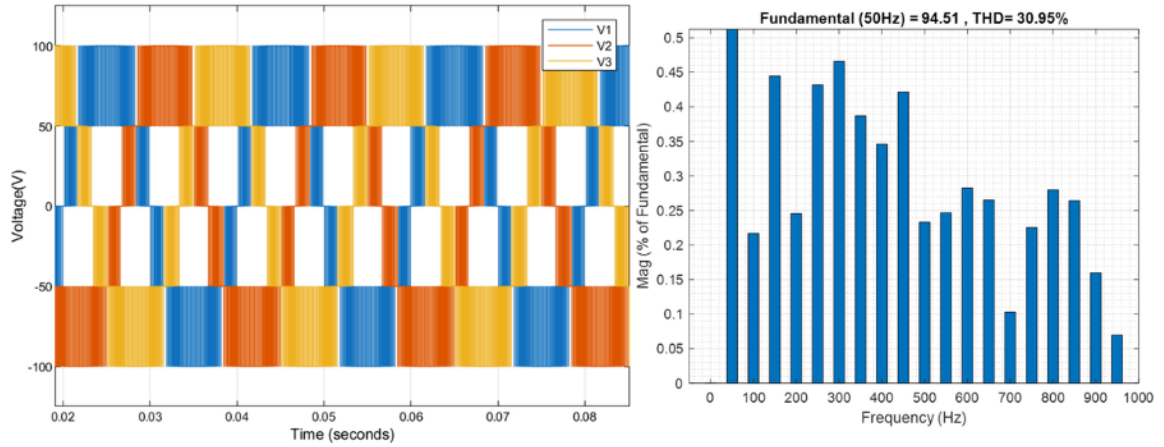


Figure 3.12: Three-phase output voltages and THD of the Five-level FCMLI

3.2.2 Discussion

The obtained results show that the Flying Capacitor Multilevel Inverter produces performance that is very similar to the Neutral Point Clamped topology in terms of output waveform quality and harmonic reduction. In both the three-level and five-level cases, the inverter successfully generates the expected voltage levels with a significant improvement in THD when the number of levels increases.

However, this improvement comes with a major drawback related to control complexity. Unlike the NPC inverter, the FCMLI requires continuous balancing of all flying capacitor voltages. This makes the control strategy more demanding, since the switching

states must be carefully selected not only to generate the correct output voltage but also to maintain constant capacitor voltages.

The presence of redundant switching states further increases this complexity. Each state can affect capacitor charging and discharging differently, meaning the controller must actively manage these effects in real time. As a result, the implementation becomes significantly more complicated, especially when moving from three-level to five-level operation.

Despite this complexity, the simulation results confirm that stable operation is achievable, with properly balanced capacitor voltages and acceptable THD values. However, the overall gain in performance compared to simpler topologies remains limited when considering the increased control effort required.

3.2.3 Conclusion

The Flying Capacitor Multilevel Inverter presents output voltage waveforms and harmonic performances that are generally comparable to those obtained using the Neutral Point Clamped topology. Increasing the number of levels improves the waveform quality and reduces the Total Harmonic Distortion, as observed in the transition from the three-level to the five-level FCMLI.

However, the main limitation of the FCMLI topology lies in the increasing complexity associated with the flying capacitors. Unlike the NPC inverter, the capacitor voltages must be continuously monitored and balanced during operation. The charging and discharging behavior of the capacitors becomes an essential part of the control strategy.

For a three-level FCMLI, only one flying capacitor is required per phase leg, making the balancing process relatively manageable. However, as the number of levels increases, the number of capacitors rises significantly. Consequently, the number of switching states and balancing conditions also increases, making the control algorithm considerably more complicated.

Although higher levels provide lower THD values, the improvement is not sufficiently remarkable to justify the large increase in control complexity and hardware constraints. For this reason, Flying Capacitor Multilevel Inverters are generally less preferred in very high-level industrial applications. In practical systems, engineers usually favor converter topologies that offer a better compromise between performance, implementation complexity, reliability, and control simplicity.

Therefore, despite the good waveform quality produced by the FCMLI topology, its practical use becomes limited at higher levels because the additional complexity introduced by capacitor voltage balancing does not provide a proportional improvement in performance compared to other multilevel inverter topologies.

3.3 Introduction and Context

The Modular Multilevel Converter (MMC) is a multilevel voltage-source converter topology that has gained significant attention in high-power and high-voltage applications. It was first introduced in the early 2000s by **R. Marquardt** and **A. Lesnicar**, and has since become a dominant solution in modern power electronics, particularly in HVDC transmission and renewable energy integration systems.

The increasing interest in MMC technology is mainly due to its excellent performance in terms of scalability, output waveform quality, and efficiency. These characteristics make it a strong alternative to conventional multilevel inverter topologies such as the Cascaded H-Bridge (CHB), Neutral Point Clamped (NPC), and Flying Capacitor (FC) converters.

3.3.1 The Converter and The Submodule Structure

The MMC is composed of three-phase legs, where each phase consists of an upper and a lower arm. Each arm is formed by a series connection of multiple identical submodules, together with an arm inductor connected in series.

Each phase leg can be described as follows:

- One upper arm
- One lower arm

Each arm contains a number of submodules (SMs) connected in series. The DC source voltage is distributed across the arms, and the midpoint defines the AC output terminal. The converter operates with a single DC source, typically divided conceptually between the upper and lower arms depending on the system reference.

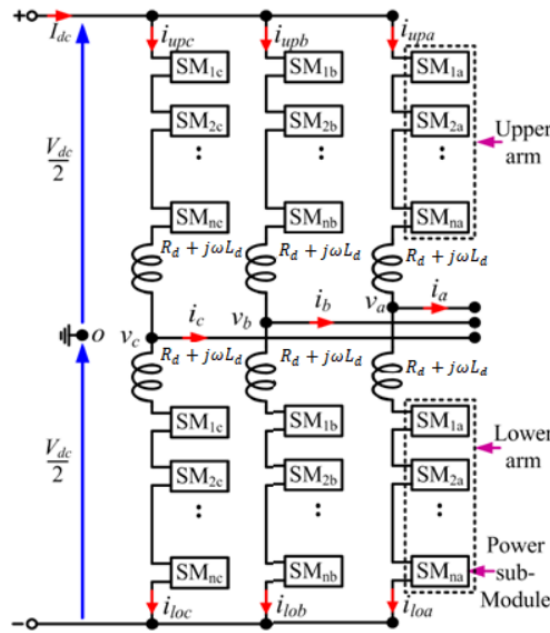


Figure 3.1: Three phase N level MMC topology [17]

Each Submodule can be implemented using either:

- Half-Bridge (HB) submodule, also known as a chopper cell.
- Full-Bridge (FB) submodule, also known as a bridge cell.

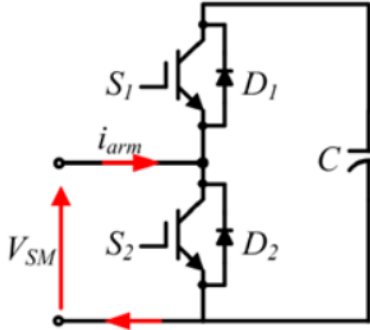


Figure 3.2: Half-Bridge (HB) submodule[16]

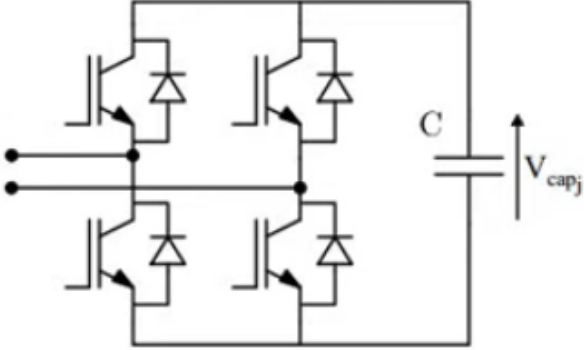


Figure 3.3: Full-Bridge (FB) submodule[16]

Each arm contains a series inductor, which plays a critical role in shaping the current behavior.

The arm inductors:

- Limit the rate of change of current $\left(\frac{di}{dt}\right)$
- Reduce current ripple
- Ensure quasi-continuous arm current

As a result, the internal arm currents are continuous rather than chopped, which improves overall system stability and reduces stress on semiconductor devices.

3.3.2 The Advantages and disadvantages of each Submodule Topology

The half-bridge submodule consists of two controlled switches and one capacitor connected in parallel with the switching structure.

Its main advantages are a reduced number of switches, which leads to lower cost and a simpler implementation, making it widely used in simulations and in basic studies of modular multilevel converters.

However, its main limitation is the lack of fault tolerance, since the failure of a single switch can render the entire submodule inoperative.

In contrast, the full-bridge submodule is composed of four switches and a capacitor, and its main advantage is improved fault tolerance and redundancy, allowing the system to continue operating even under certain fault conditions.

On the other hand, this improved performance comes at the expense of higher cost, increased structural complexity, and greater switching losses compared to the half-bridge configuration.

3.3.3 Capacitor Role and Voltage Balancing

Each submodule contains a capacitor connected in parallel with the switching structure, serving as a distributed energy storage element. Proper MMC operation requires maintaining balanced capacitor voltages to ensure equal voltage sharing among submodules and stable converter performance.

During operation, the capacitors continuously charge and discharge depending on the switching states and current direction. Therefore, a voltage balancing control strategy is required to regulate the capacitor voltages. Without proper balancing, voltage deviations may occur, leading to output waveform distortion and reduced system stability.

3.3.4 Control Principle

The control of the MMC is based on two main aspects:

3.3.5 Switching Control

The switching control of MMC submodules is relatively simple, as the switches operate in a complementary manner, ensuring safe operation and preventing capacitor short circuits. This approach requires only basic switching logic for each submodule.

However, although the switching process itself is straightforward, the overall control becomes more challenging when determining which submodules should be inserted or bypassed to maintain capacitor voltage balance and achieve proper converter performance.

3.3.6 Control Strategy and Structural Development of the Modular Multilevel Converter (MMC)

The control of the MMC is based on managing the switching states of its submodules to generate the desired output voltage. In half-bridge submodules, the switches operate in a complementary manner, allowing simple switching logic with a single control signal.

To facilitate understanding, the converter can be developed progressively by analyzing a single submodule, then assembling submodules into arms, arms into legs, and finally constructing a complete MMC phase. The number of output voltage levels depends on the number of submodules per arm, with higher numbers of submodules producing more voltage levels and improved waveform quality.

The proposed development approach begins with simple MMC configurations and gradually increases in complexity. Different control and switching strategies are then evaluated through performance indicators such as Total Harmonic Distortion (THD). Once the converter model is fully validated, it can be integrated with renewable energy sources, such as wind turbine generation systems.

3.3.7 Single Submodule Modeling (Half Bridge)

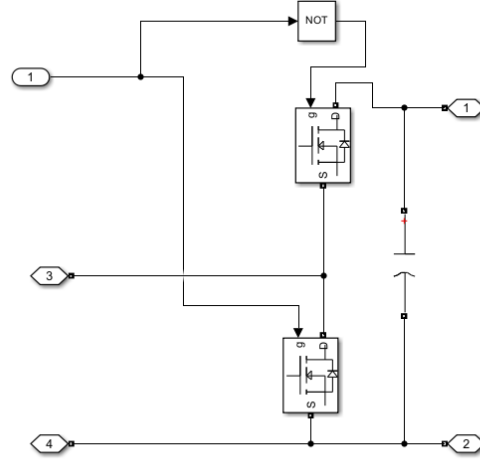


Figure 3.4: Single half-bridge submodule model

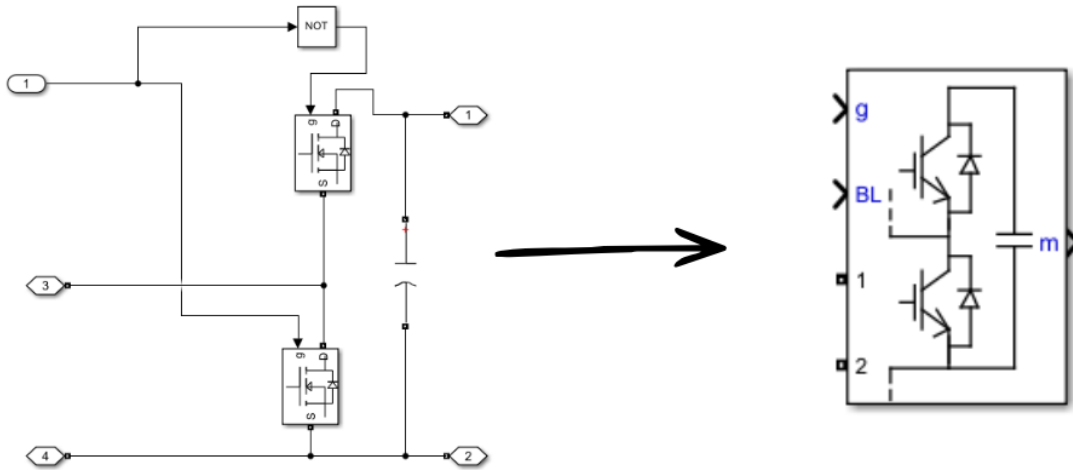


Figure 3.5: Submodule Ready to use block in MATLAB/Simulink

Next, voltage and current measurement blocks are added to monitor the converter operation. Voltage sensors are connected across the capacitors and the load, while current sensors are placed within the converter arms to determine the magnitude and direction of current flow. A common sensor orientation is adopted for all arms, simplifying the implementation of the control algorithm and allowing the same control structure to be applied throughout the converter.

To ensure proper capacitor voltage balancing, the controller continuously monitors the voltage of each submodule capacitor. The measured voltages are collected, processed, and sorted to identify the highest and lowest voltage levels within each arm. This information enables the control system to select the appropriate submodules for charging or discharging, thereby maintaining balanced capacitor voltages and ensuring stable converter operation.

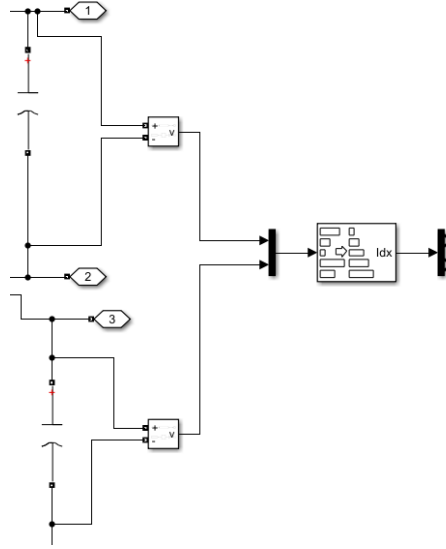


Figure 3.6: Capacitor-voltage sorting and balancing block diagram

The Level-Shifted Carrier PWM technique is used to control the MMC inverter. Each arm is modulated by comparing a sinusoidal reference signal with two level-shifted carrier signals, which define the switching boundaries.

In a conventional implementation, four carrier signals would be required for a phase leg containing four submodules. However, in this work, the number of carriers is reduced to two by exploiting the consistent current sign convention across all measurements. This allows the lower arm to be treated as the complementary behavior of the upper arm, simplifying the modulation structure.

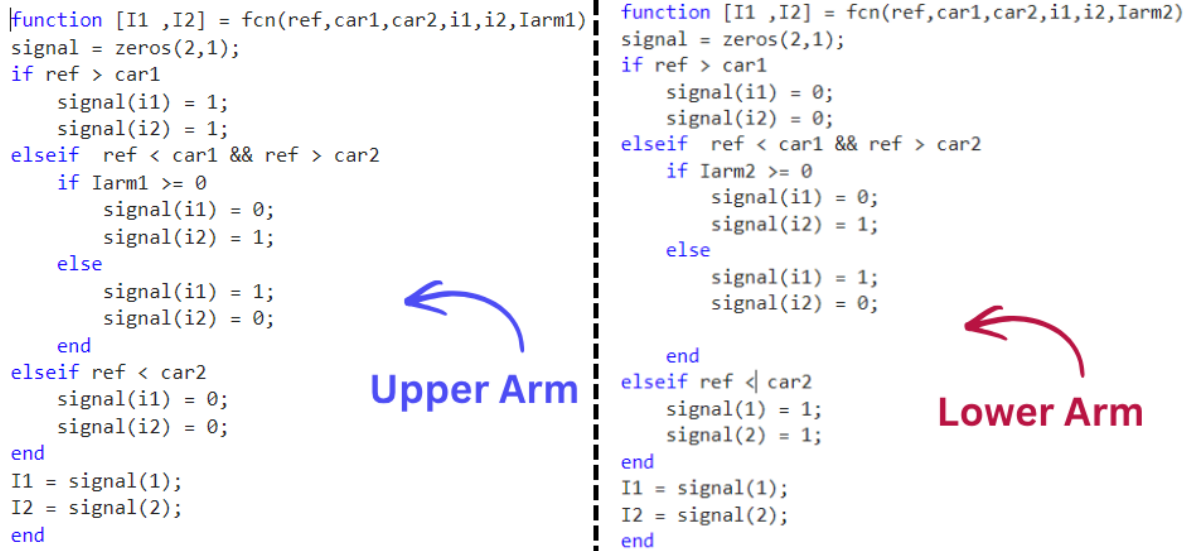
The control system is implemented using two MATLAB Function blocks, one per arm. Each block receives the same modulation inputs (reference sine wave and carrier signals), along with arm-specific signals including the sorted capacitor voltages and the measured arm current.

If the reference signal is above the upper carrier, both switches are set to bypass mode. If it is below the lower carrier, both switches are set to insertion mode.

When the reference lies between the two carriers, capacitor voltage balancing is applied. The sorting algorithm provides the highest and lowest capacitor voltages, and the switching decisions are made accordingly based on current direction.

For positive current, the submodule with the lower capacitor voltage is inserted while the higher one is bypassed, allowing controlled charging of the lower-voltage capacitor. For negative current, the roles are reversed to promote discharge of the higher-voltage capacitor.

The “first” and “second” submodules correspond to the sorted capacitor voltages rather than fixed physical positions, ensuring that the control always targets the lowest and highest voltage cells in real time. And these are the MATLAB function Codes :



```

function [I1 ,I2] = fcn(ref,car1,car2,i1,i2,Iarm1)
signal = zeros(2,1);
if ref > car1
    signal(i1) = 1;
    signal(i2) = 1;
elseif ref < car1 && ref > car2
    if Iarm1 >= 0
        signal(i1) = 0;
        signal(i2) = 1;
    else
        signal(i1) = 1;
        signal(i2) = 0;
    end
elseif ref < car2
    signal(i1) = 0;
    signal(i2) = 0;
end
I1 = signal(1);
I2 = signal(2);
end
    
```

Upper Arm

```

function [I1 ,I2] = fcn(ref,car1,car2,i1,i2,Iarm2)
signal = zeros(2,1);
if ref > car1
    signal(i1) = 0;
    signal(i2) = 0;
elseif ref < car1 && ref > car2
    if Iarm2 >= 0
        signal(i1) = 0;
        signal(i2) = 1;
    else
        signal(i1) = 1;
        signal(i2) = 0;
    end
elseif ref < car2
    signal(1) = 1;
    signal(2) = 1;
end
I1 = signal(1);
I2 = signal(2);
end
    
```

Lower Arm

Figure 3.7: MATLAB function code for upper-arm and lower-arm switching logic

The same procedure is then repeated to create the second and third phases of the converter. These phases are implemented using exactly the same operating principle and control logic as the first phase. Each arm contains its own MATLAB Function block responsible for generating the switching signals according to the carrier comparison, arm current direction, and capacitor voltage balancing strategy.

Since the converter is composed of three phases and each phase contains two arms, the complete MMC model requires a total of six arms and therefore six MATLAB Function blocks, one for each arm independently.

The three-phase MMC model is implemented using a modular structure in MATLAB/Simulink. Although each arm appears to contain a single submodule block, this block is actually a configurable half-bridge submodule that can represent multiple cascaded cells by adjusting its internal parameters.

This approach allows the converter structure to be easily scaled by increasing the number of submodules without manually constructing each switching cell. As a result, higher-level MMC configurations can be obtained simply by modifying the block settings.

However, increasing the number of submodules also requires updating the modulation strategy, since the number of level-shifted carrier signals must match the number of voltage levels to ensure proper switching control.

With the full three-phase system implemented, the simulation enables observation of phase and line voltages, providing a complete view of the converter behavior. The modular design also facilitates straightforward expansion toward higher-resolution output waveforms.

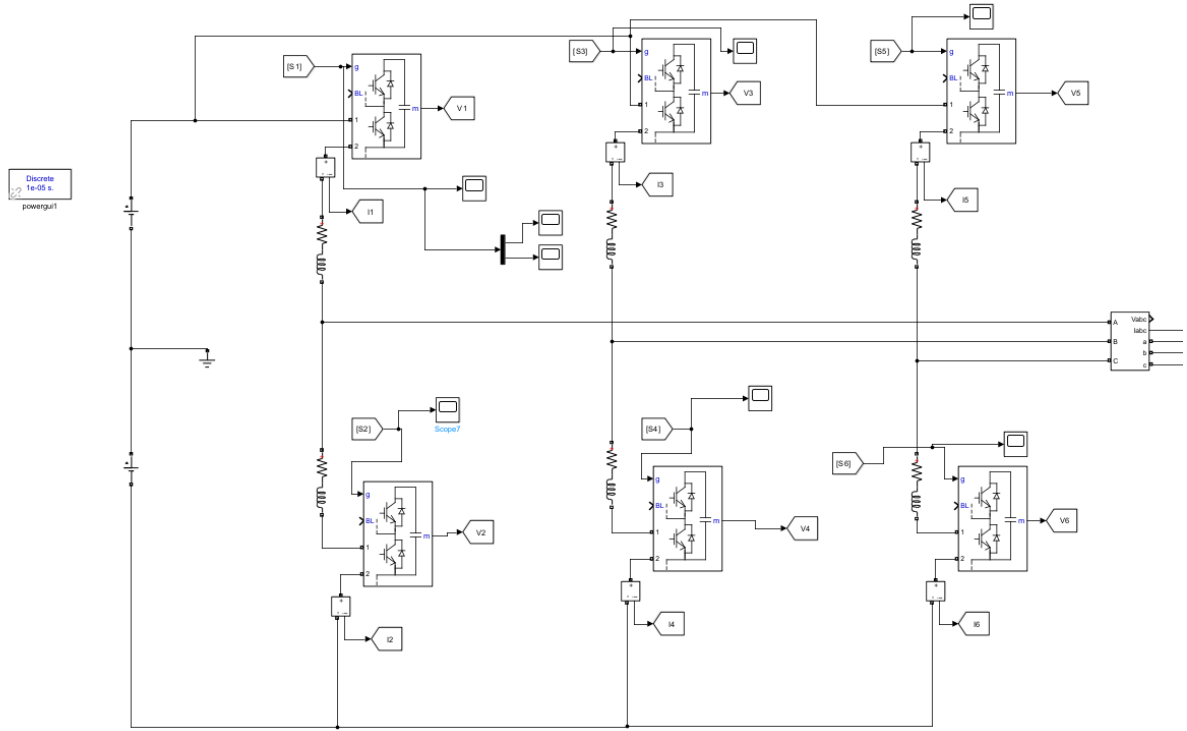


Figure 3.8: Complete three-phase MMC model

The main difference is that, instead of using a single sinusoidal reference signal, three sinusoidal reference waveforms are employed, one for each phase of the converter. These reference signals all operate at the same frequency and are phase-shifted by 120° ($\frac{2\pi}{3}$) relative to one another in order to generate a balanced three-phase system.

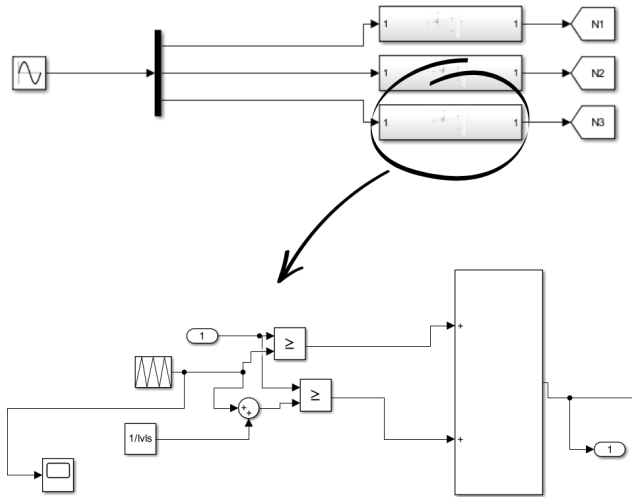


Figure 3.9: Three-phase sinusoidal references and comparator logic

Thus, the first reference signal is taken as the phase reference, the second is shifted by -120° , and the third by $+120^\circ$. These three reference waveforms are then compared with the corresponding level-shifted carrier signals to generate the switching commands for the six converter arms.

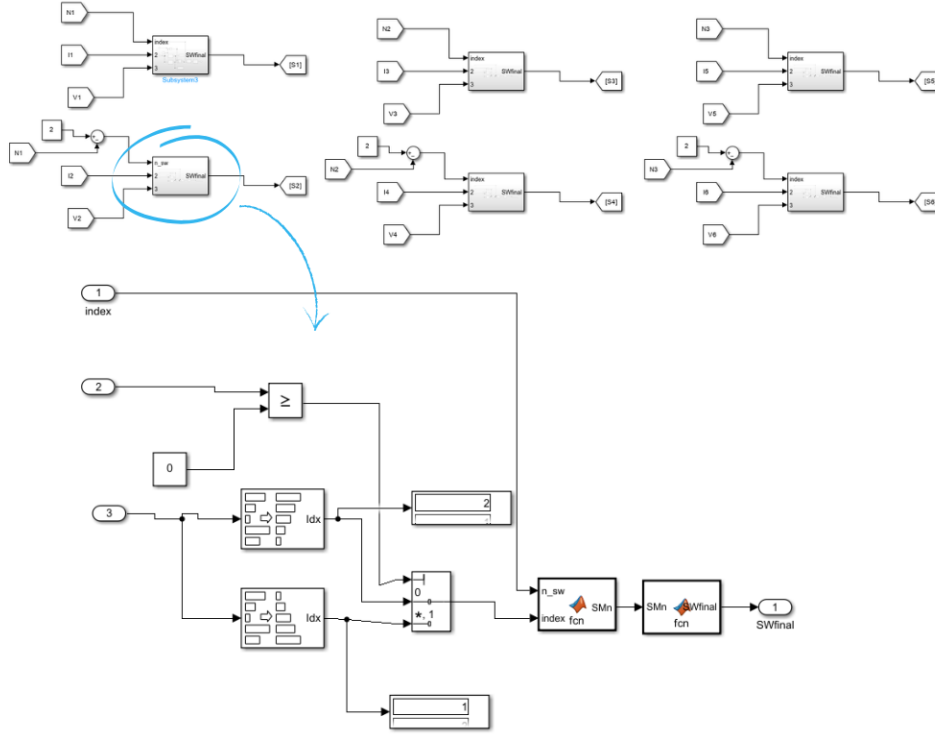


Figure 3.10: Sorting, index selection, and gate-signal generation logic

3.3.8 Results And Interpretations

After simulation, the voltage levels of the three phases can be clearly observed. The phase voltage takes the values $+V_{dc}$, 0 and $-V_{dc}$, which corresponds to a three-level inverter.

The following figure shows the voltage levels measured between two phases. Although each phase contains only three levels, the line-to-line voltage presents five levels. This occurs because the line voltage is obtained by subtracting the voltages of two different phases. As a result, additional voltage combinations appear, producing the levels $+2V_{dc}$, $+V_{dc}$, 0 , $-V_{dc}$, and $-2V_{dc}$. Therefore, the voltage measured between phases contains five distinct levels instead of three.

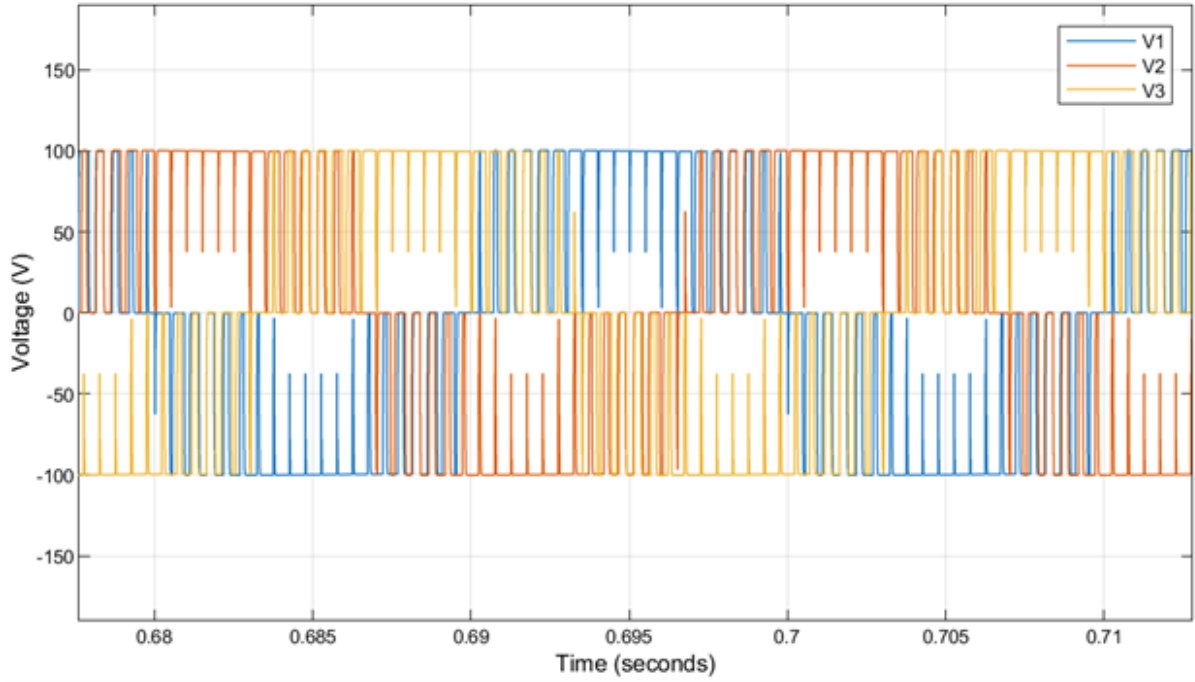


Figure 3.11: Three phase voltages of the 3-level MMC

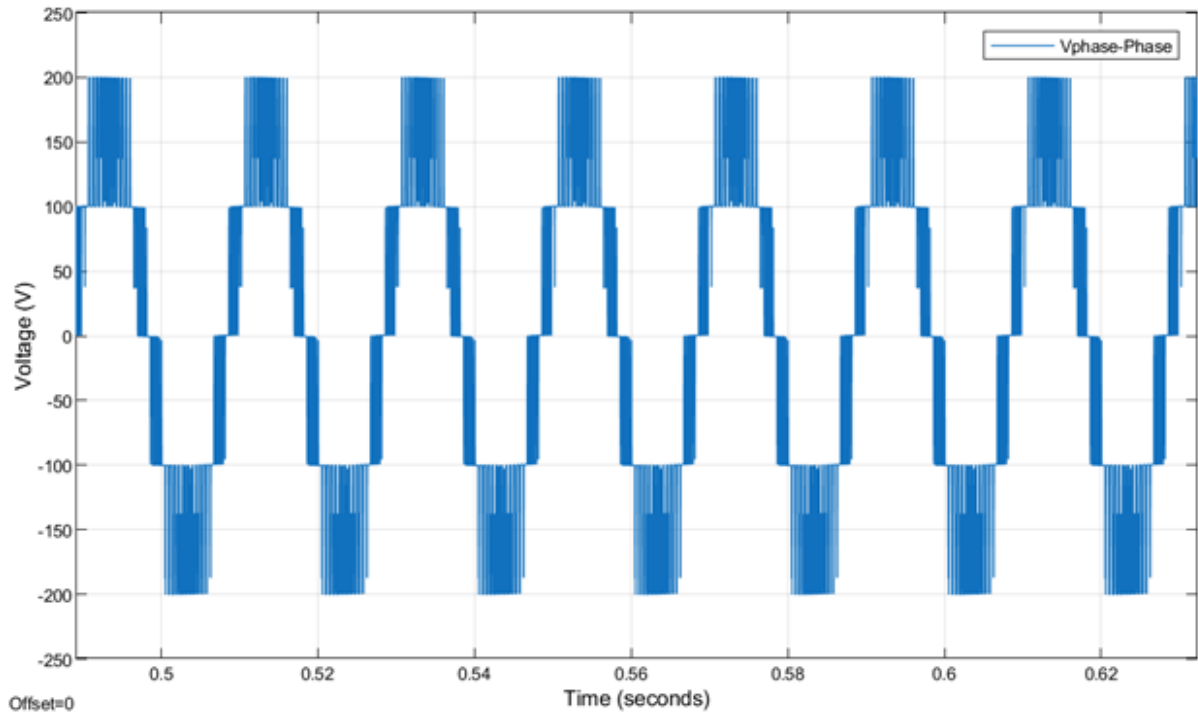


Figure 3.12: Line-to-line voltage waveform of the MMC

The Figure Next shows the output voltages of the MMC inverter for 3, 5, 15, and 19 levels. The values below each waveform represent the Total Harmonic Distortion (THD) calculated with respect to a 50 Hz sinusoidal reference signal.

The obtained THD values are:

- 3-level MMC: THD = 36%
- 5-level MMC: THD = 27.9%
- 15-level MMC: THD = 8%
- 19-level MMC: THD = 6.12%

It can be observed that increasing the number of voltage levels reduces the THD significantly. This is because the output waveform becomes closer to an ideal sinusoidal signal, which decreases the harmonic content and improves the output voltage quality.

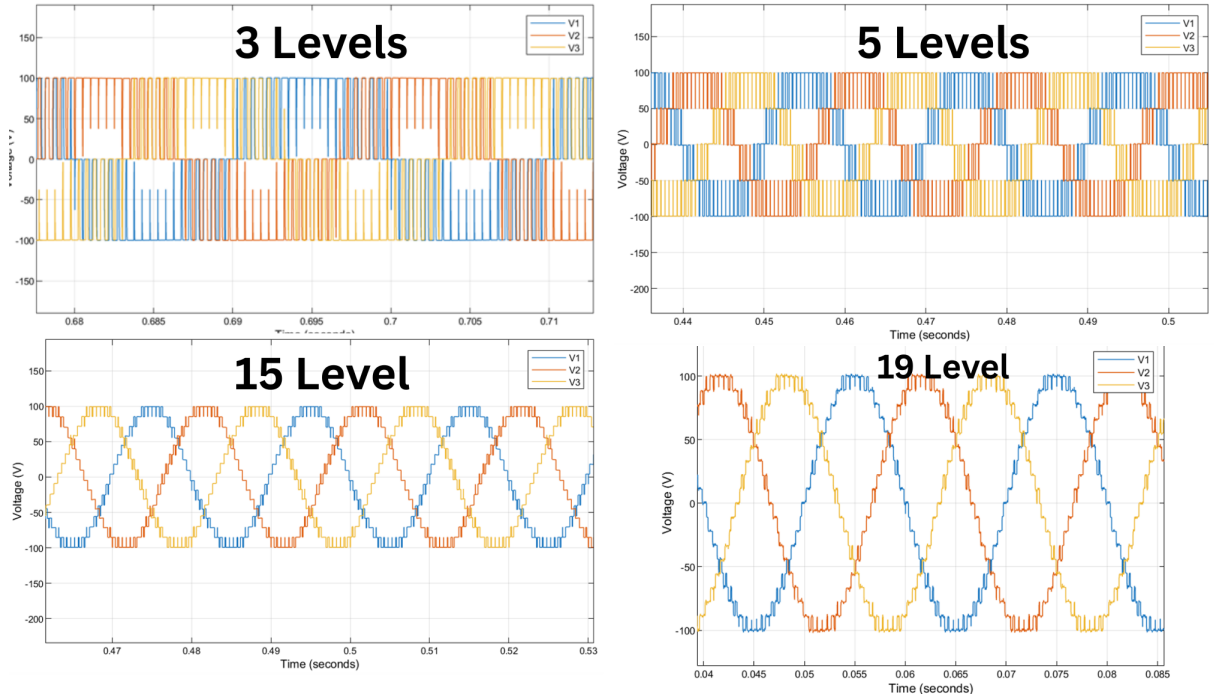


Figure 3.13: Output voltages for 3-, 5-, 15-, and 19-level MMCs

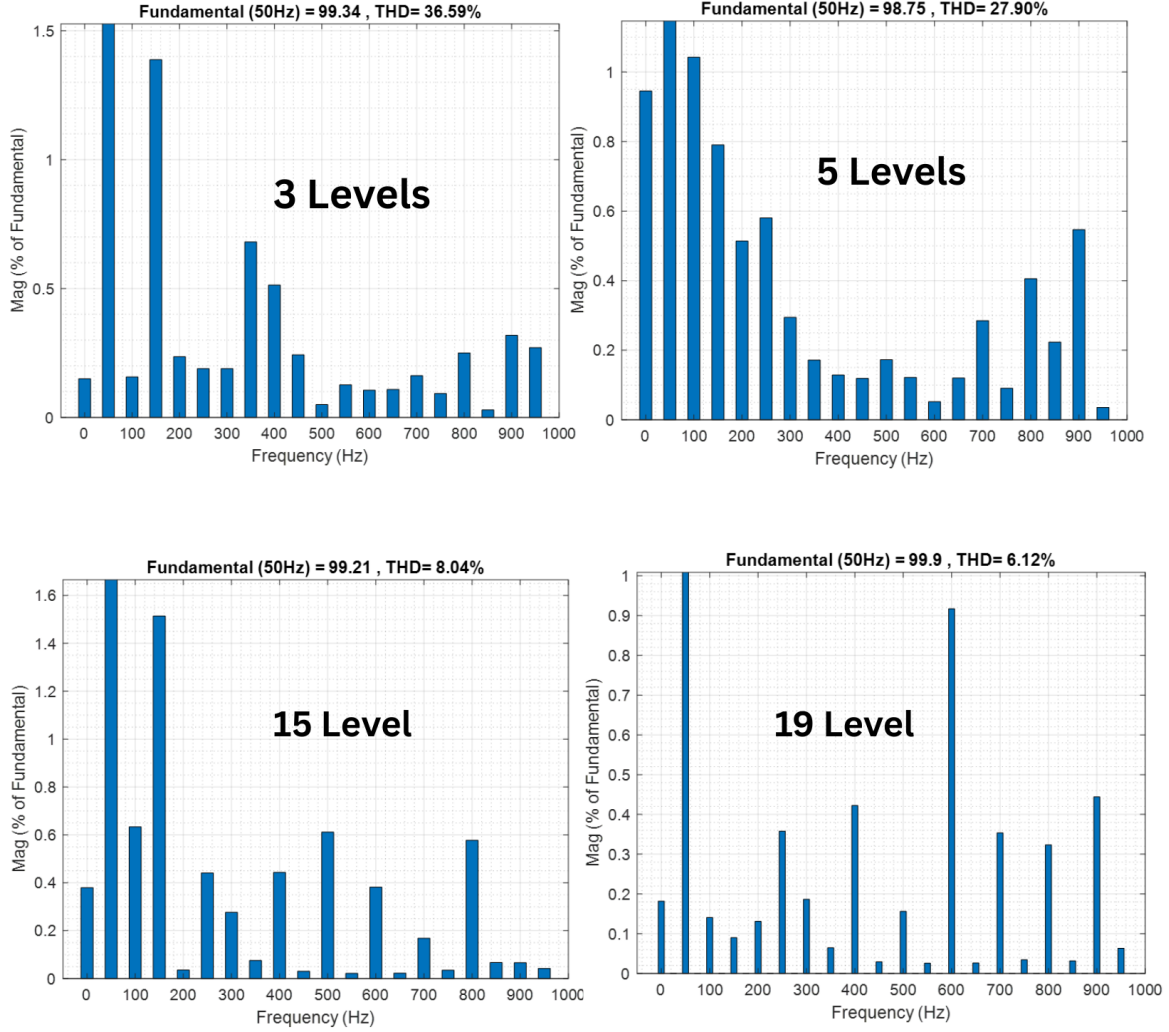


Figure 3.14: Harmonic spectra and THD comparison for different MMC levels

3.3.9 Conclusion

This chapter presented the Modular Multilevel Converter (MMC), including its structure, operating principles, submodule configurations, capacitor voltage balancing strategy, and control methodology. A progressive modeling approach was adopted, starting from a single half-bridge submodule and extending to a complete three-phase MMC system implemented in MATLAB/Simulink.

The simulation results demonstrated the influence of the number of voltage levels on output waveform quality. As the number of levels increased, the Total Harmonic Distortion (THD) decreased significantly, confirming the capability of the MMC to generate output voltages that closely approximate sinusoidal waveforms.

The study also highlighted the importance of capacitor voltage balancing and modulation strategies in ensuring stable converter operation. Despite the increased control complexity associated with balancing and switching management, the MMC remains a highly scalable topology capable of achieving excellent output voltage quality.

Overall, the obtained results confirm the effectiveness of the MMC in multilevel power conversion applications and provide a solid understanding of its behavior, advantages, and practical limitations.

Chapter 4: Integration of Multilevel Converters into the Wind Energy Conversion Chain Based on a Dual Star Synchronous Generator

4.1 Introduction

The previous chapters were dedicated to the design, modeling, and simulation of three multilevel inverter topologies: the Neutral Point Clamped Multilevel Inverter (NPC-MLI), the Flying Capacitor Multilevel Inverter (FCMLI), and the Modular Multilevel Converter (MMC). Each topology was studied and validated independently under controlled DC source conditions.

The objective of this chapter is to go a step further by integrating each of these inverters with a six-phase generator through a rectifier stage, forming a complete energy conversion chain. This allows us to evaluate the real behavior of each inverter under more realistic and dynamic supply conditions, where the DC-link voltage is no longer ideal but is instead provided by the rectified output of the generator. The three topologies are tested under the same generator, and the results are compared at the end of the chapter in order to draw a final conclusion on the performance and suitability of each topology.

4.2 NPC Multilevel Inverter Connected to the Six-Phase Generator

4.2.1 Connection with the Six-Phase Generator

As a first integration test, the developed NPC multilevel inverter models were connected to the six-phase generator system in order to evaluate their performance under realistic operating conditions. The three-level and five-level configurations were both tested with the generator. The obtained results show the output voltage waveforms generated by each inverter structure after connection with the six-phase system.

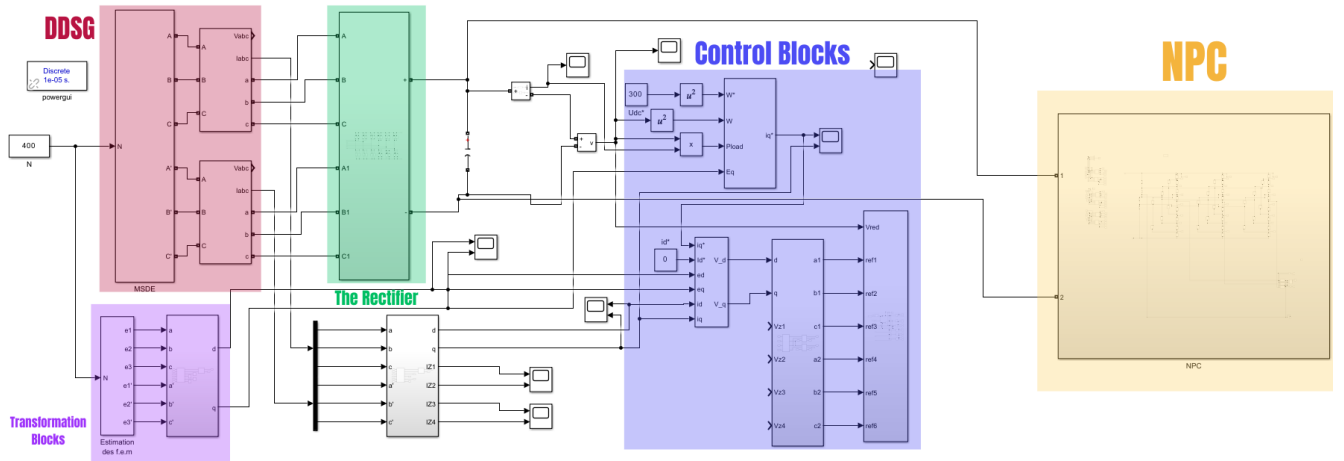


Figure 4.1: NPC inverter connected to the six-phase generator

4.2.2 Simulation Results

The simulation results confirm that the five-level inverter produces a smoother voltage waveform compared to the three-level configuration due to the increased number of voltage levels.

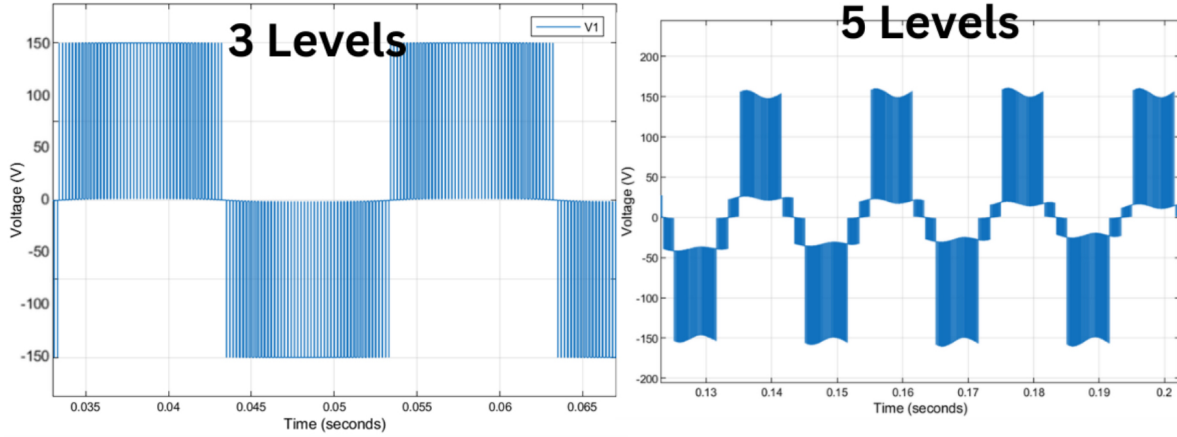


Figure 4.2: NPC inverter connected to the six-phase generator Results

4.3 Flying Capacitor Multilevel Inverter Connected to the Six-Phase Generator

4.3.1 Integration with the Six-Phase Generator

Following the NPC integration, the same procedure is applied to the Flying Capacitor Multilevel Inverter. The five-level FCMLI was connected to the six-phase generator, as shown in Figure 4.3. The corresponding simulation results are presented in Figure 4.4. During the startup phase, the output waveform presents some distortion while the rectifier is establishing the DC-link voltage. Once the rectifier reaches approximately 300 V, the converter operates under more stable conditions and the waveform quality improves noticeably.

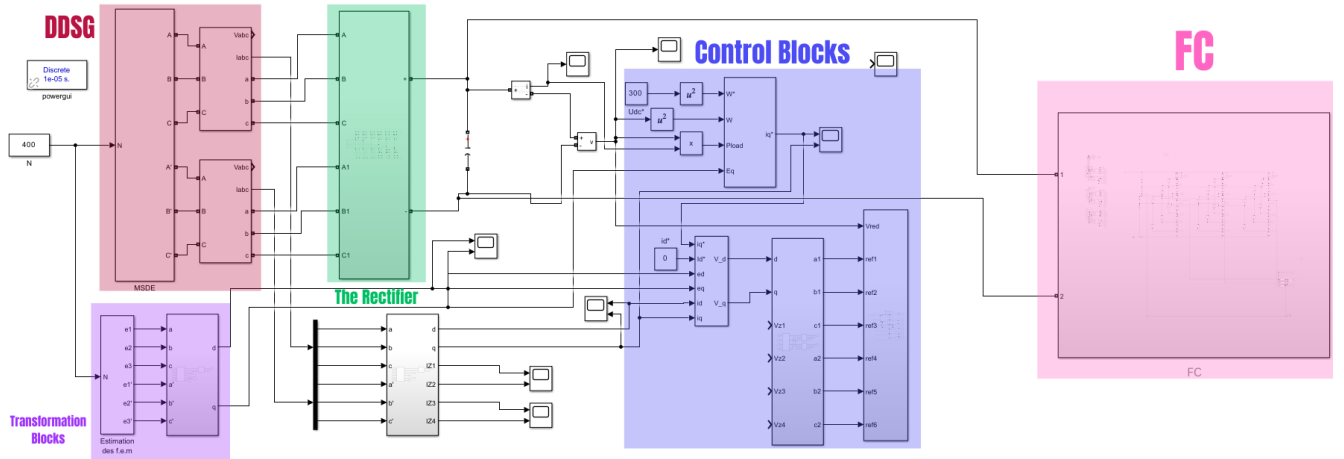


Figure 4.3: Five-level FCMLI connected to the six-phase generator

4.3.2 Simulation Results

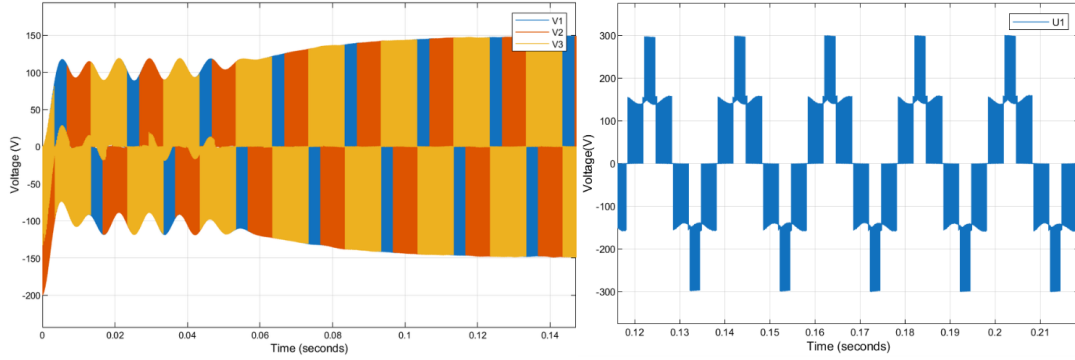


Figure 4.4: Output waveforms obtained after the integration of the FCMLI with the six-phase generator

4.4 Modular Multilevel Converter Connected to the Six-Phase Generator

4.4.1 Integration with the Six-Phase Generator

The final integration test concerns the Modular Multilevel Converter. The approach here follows the same principle as the previous two topologies, but with a slight modification to the model structure. The MMC is first gathered into a subsystem in order to avoid complicating the overall model. We use the same 19-level MMC that was previously designed, and we simply connect the positive terminal to the positive output of the rectifier and the negative terminal to the negative output of the rectifier. In other words, the DC voltage generators used previously in the MMC are replaced by capacitors connected to the rectifier outputs.

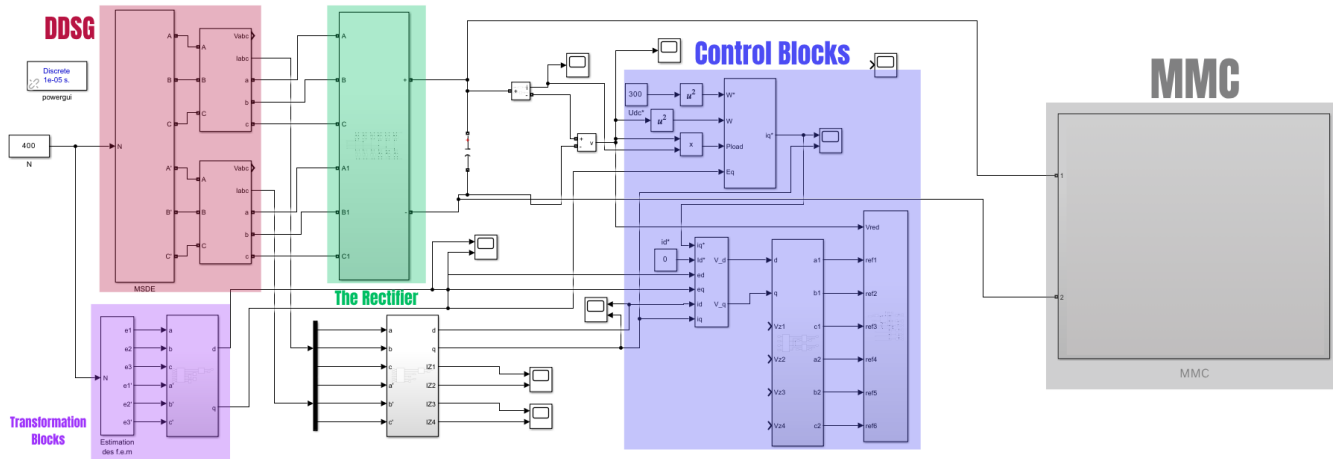


Figure 4.5: MMC subsystem connected to the rectifier

4.4.2 Simulation Results

Each capacitor has a capacitance of $3300 \mu F$ and is initially charged to $150 V$, for both the positive and negative sides, with the ground naturally placed between them. After that, the simulation is launched. The obtained results are quite expected. First, the MMC encounters some difficulty maintaining perfectly balanced three-phase voltage waveforms

during the first milliseconds of the simulation. This is because the rectifier itself does not instantly provide a stable 300 V DC output; instead, the voltage increases progressively from its initial value. As a consequence, the MMC initially struggles to generate stable three-phase waveforms. However, once the rectifier output reaches approximately 300 V, it can clearly be observed that the MMC becomes stable and succeeds in generating proper three-phase 19-level voltage waveforms with a relatively low THD and good waveform quality.

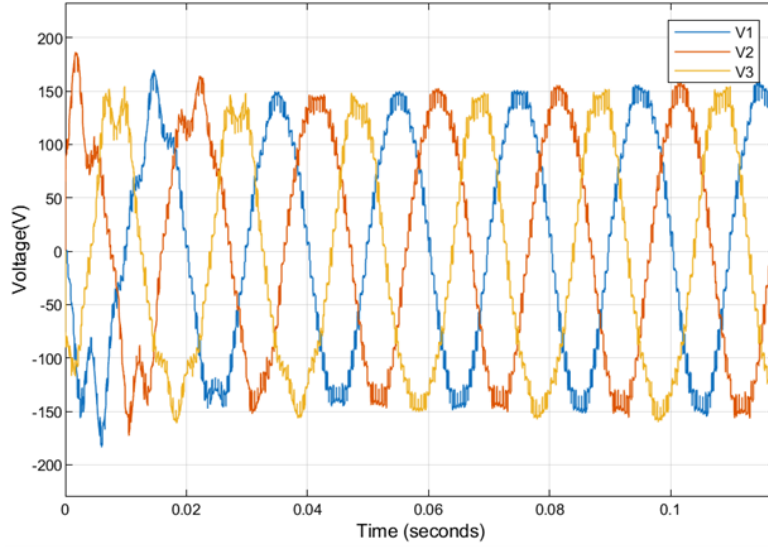


Figure 4.6: Three-phase output voltages after rectifier connection

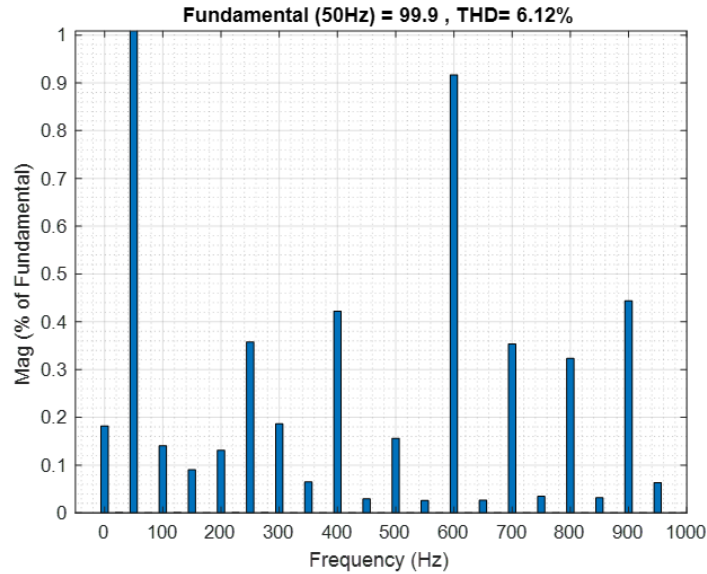


Figure 4.7: Harmonic spectrum of the 19-level MMC after rectifier connection

4.5 Comparative Analysis of the Three Topologies

Having tested all three inverter topologies under the same six-phase generator and rectifier supply, it is now possible to draw a direct comparison between them. Table 4.1 summarizes

the key criteria, advantages, limitations, and THD results obtained for each topology across the different voltage levels studied throughout this work.

Table 4.1: Comparison of NPC-MLI, FCMLI, and MMC

Criteria	NPC-MLI	FCMLI	MMC
Main principle	Uses clamping diodes to generate multiple voltage levels.	Uses flying capacitors instead of clamping diodes.	Uses a modular structure made of series-connected submodules.
Control complexity	Relatively simple carrier-based PWM and truth-table control.	More complex because capacitor voltage balancing must be continuously managed.	The Simulink model is easier thanks to ready-to-use submodule blocks, but the control codes are complicated because they require sorting, balancing, and switching logic.
Voltage balancing	No major balancing issue is emphasized.	Main limitation of the topology.	Essential for all submodules and must be continuously maintained.
Main advantage	Good compromise between performance and control simplicity.	Good voltage sharing and improved waveform quality.	Excellent scalability and very low THD at high numbers of levels.
Main disadvantage	Increasing the number of levels increases the control and hardware complexity.	The number of capacitors and redundant switching states increases rapidly.	The control system is advanced and computationally demanding.
THD results	About 58% for the three-level case and about 30% for the five-level case.	About 58.08% for the three-level case and about 30.95% for the five-level case.	36% for the 3-level case, 27.9% for the 5-level case, 8% for the 15-level case, and 6.12% for the 19-level case.
Scalability	Useful, but the control becomes more complex as the number of levels increases.	Less preferred for very high-level applications because balancing becomes difficult.	Highly scalable and suitable for very high-level operation.
Overall conclusion	A good compromise between performance and simplicity.	Good waveform quality, but balancing complexity limits practical use at higher levels.	Best waveform quality and scalability, but the control remains the most complex.

4.6 Conclusion

This chapter presented the integration of the three multilevel inverter topologies — NPC-MLI, FCMLI, and MMC — with a six-phase generator through a rectifier stage. The results obtained for each topology confirmed that all three inverters are capable of operating correctly under realistic supply conditions, with only a short transient disturbance observed at startup due to the progressive rise of the DC-link voltage.

The comparative analysis shows that while the NPC-MLI and FCMLI offer acceptable performance at low voltage levels, the MMC clearly stands out in terms of waveform quality

and scalability, achieving a THD as low as 6.12% at the 19-level configuration. These results validate the MMC as the most suitable topology for high-performance applications requiring a large number of voltage levels.

General Conclusion

This thesis has presented a comprehensive study of the DSSG-PWM rectifier feeding a DC bus, building upon the foundational work of our fellow colleagues. Through this work, we have fully explored the use of the DSSG-PWM rectifier in an energy conversion chain up to the DC-AC conversion stage. We have investigated different multilevel inverter topologies, comparing them based on several criteria including redundancy, ease of development, ability to add levels, control strategy complexity, and THD, and evaluated the overall performance and suitability of each topology within the conversion chain.

The first chapter presented a complete description of the conversion chain with a general overview of each stage and its role in the overall system.

Chapter two was dedicated to the modeling of the DSSG-PWM rectifier conversion system. Understanding the dynamic modeling of these two components was essential to fully grasp the work of our fellow colleagues and to explore the advantages of this system in the energy conversion chain.

Chapter three investigated three distinct multilevel inverter families: neutral point clamped, flying capacitor, and modular multilevel inverter (MMC). We examined 3-level and 5-level configurations for the first two topologies, and 3-level, 5-level, 9-level, and 19-level configurations for the MMC. Two modulation strategies were evaluated for each topology: level-shifted PWM and carrier phase-shifted PWM. The topologies were compared based on control complexity, ease of adding levels, and THD performance.

Chapter four presents the integration of the three multilevel inverter topologies with a dual-star synchronous generator through a rectifier stage, forming a complete energy conversion chain. The DC-link voltage is provided by the rectified output of the generator rather than an ideal source, and the results are compared to conclude on the performance and suitability of each topology within the conversion chain.

In conclusion, this thesis has comprehensively explored the energy conversion chain for renewable energy applications, covering the stages from mechanical rotational energy conversion up to the inversion stage. We developed a deep understanding of the DSSG-PWM rectifier system through dynamic modeling, building upon the foundational work of our fellow colleagues. Understanding their work was essential to enable us to fully explore the potential of this system within the energy conversion chain. We investigated three distinct multilevel inverter topologies (neutral point clamped, flying capacitor, and modular multilevel inverter), comparing their performance based on redundancy, ease of development, ability to add levels, control strategy complexity, and THD.

The work done in this thesis opens several opportunities towards future research and development. Key points for future investigation include:

- **Experimental Validation:** Experimental validation of the models and control strategy through a real test bench implementation is essential. This would allow confirmation of the simulation results and identification of the real practical challenges associated with the implementation of the energy conversion chain. Laboratory testing on a physical prototype would bridge the gap between simulation and practical deployment, revealing constraints and behaviors that simulation alone cannot capture. Furthermore, exploring the use of these multilevel inverter topologies

for grid-connected renewable energy applications represents a natural extension of this work, where the key challenges associated with grid synchronization, power quality, and compliance with grid codes would need to be thoroughly investigated.

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Abstract

This thesis presents a comprehensive study of the energy conversion chain for renewable energy applications using a dual-star synchronous generator associated with a six-phase rectifier with pulse width modulation, as well as multilevel inverter topologies. Chapter one describes the energy conversion chain covering each stage from mechanical rotational energy conversion up to the DC-AC conversion stage. Chapter two develops a deep understanding of the dual-star synchronous generator system associated with a six-phase PWM rectifier through dynamic modeling. Chapter three investigates three distinct families of multilevel inverters: neutral point clamped, flying capacitor, and modular multilevel inverter, comparing their performance based on redundancy, ease of development, ability to add levels, control strategy complexity, and THD. Chapter four presents the integration of the three multilevel inverter topologies with a dual-star synchronous generator through a rectifier stage. The obtained results are then compared to conclude on the performance and suitability of each topology within the conversion chain.

This work paves the way for significant future research. Future work should focus on experimental validation through real test bench implementation, exploration of advanced control strategies using machine learning approaches, and investigating the use of these multilevel inverter topologies for grid-connected renewable energy applications, addressing the key challenges associated with grid synchronization, power quality, and compliance with grid codes.

Résumé

Ce mémoire présente une étude approfondie de la chaîne de conversion d'énergie pour les applications d'énergies renouvelables, utilisant un générateur synchrone à double étoile associé à un redresseur hexaphasé à modulation de largeur d'impulsion, ainsi que des topologies d'onduleurs multiniveaux. Le premier chapitre décrit la chaîne de conversion d'énergie en couvrant chaque étape, depuis la conversion de l'énergie mécanique rotative jusqu'au stade de l'inversion. Le deuxième chapitre développe une compréhension approfondie du système générateur synchrone à double étoile associé à un redresseur MLI hexaphasé à travers la modélisation dynamique. Le troisième chapitre étudie trois familles distinctes d'onduleurs multiniveaux : l'onduleur à point neutre fixe, l'onduleur à capacités flottantes et le convertisseur modulaire multiniveau, en comparant leurs performances sur la base de la redondance, de la facilité de développement, de la capacité à ajouter des niveaux, de la complexité de la stratégie de commande et du taux de distorsion harmonique. Le chapitre quatre présente l'intégration des trois topologies d'onduleurs multiniveaux avec un générateur synchrone à double étoile à travers un étage redresseur. Les résultats obtenus sont ensuite comparés afin de conclure sur les performances et l'adéquation de chaque topologie au sein de la chaîne de conversion.

Ce travail ouvre la voie à d'importantes recherches futures, les travaux futurs devront se concentrer sur la validation expérimentale par la mise en œuvre d'un banc d'essai réel, l'exploration de stratégies de commande avancées utilisant des approches d'apprentissage automatique, ainsi que l'étude de l'utilisation de ces topologies d'onduleurs multiniveaux pour des applications d'énergies renouvelables connectées au réseau, en abordant les défis associés à la synchronisation au réseau, à la qualité de l'énergie et à la conformité aux codes du réseau.

ملخص

يقدم هذا البحث دراسة شاملة لسلسلة تحويل الطاقة في تطبيقات الطاقة المتجددة، باستخدام مولد تزامني ثنائي النجمة مرتبط بمقوم سداسي الأوجه يتضمن عرض النبضة، فضلاً عن طوبولوجيات العاكسات متعددة المستويات. يصف الفصل الأول سلسلة تحويل الطاقة بتغطية كل مرحلة، من تحويل الطاقة الميكانيكية الدورانية وصولاً إلى مرحلة العكس. يطور الفصل الثاني فهماً عميقاً لنظام المولد التزامني ثنائي النجمة المرتبط بمقوم يتضمن عرض النبضة سداسي الأوجه من خلال النمذجة الديناميكية. يدرس الفصل الثالث ثلاث عائلات متميزة من العاكسات متعددة المستويات: عاكس النقطة المحايدة المثبتة، وعاكس المكثفات الطائفة، والمحول المعياري متعدد المستويات، مقارناً أدائها على أساس التكرارية، وسهولة التطوير، والقدرة على إضافة مستويات، وتعقيد استراتيجية التحكم، ومعدل التشويه التوافقي. تناول الفصل الرابع دمج الأنظمة الثلاثة لمحولات الجهد متعددة المستويات مع المولد التزامني ذي النجمة المزدوجة عبر مرحلة التقييم. ثم تتم مقارنة النتائج المتحصل عليها للخروج باستنتاج حول أداء وملاءمة كل نظام ضمن سلسلة التحويل. يفتح هذا العمل الباب أمام أبحاث مستقبلية مهمة. ينبغي أن تركز الأعمال المستقبلية على التحقق التجريبي من خلال تنفيذ منصة اختبار حقيقية، واستكشاف استراتيجيات تحكم متقدمة باستخدام مناهج التعلم الآلي، فضلاً عن دراسة استخدام هذه الطوبولوجيات من العاكسات متعددة المستويات في تطبيقات الطاقة المتجددة المتصلة بالشبكة، مع معالجة التحديات المرتبطة بمزامنة الشبكة وجودة الطاقة والامتثال لقواعد الشبكة.