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Railway Traction Control Unit: Simulation and Implementation

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Resumo

Esta dissertação apresenta um estudo centrado na modernização do controlador do conversor de tração das locomotivas, particularmente das séries LE5600 da Comboios de Portugal, alinhado com os objetivos estratégicos da empresa para evitar a obsolescência de componentes essenciais. A necessidade de atualização visa não só prolongar a vida útil dos sistemas, mas também melhorar a eficiência energética e a adaptabilidade operacional das locomotivas.

O problema abordado neste trabalho foca-se no controlo avançado de um motor de indução trifásico em duas principais vertentes. A primeira é o controlo do conversor trifásico específico das LE5600, otimizando as tensões e correntes nominais para maximizar o desempenho e a fiabilidade. A segunda vertente envolve o controlo de um motor AC de 400V num banco de ensaios motor-gerador, alargando os estudos para aplicações além das locomotivas específicas.

Esta dissertação dedica-se à simulação detalhada das novas configurações propostas. Os resultados dos estudos de simulação são cruciais, destacando melhorias significativas no controlo de tração e na eficiência energética através da implementação de novas tecnologias, como transístores IGBT em substituição aos GTOs obsoletos. Esta abordagem não apenas supera limitações técnicas existentes, mas também promove uma integração mais eficiente com os sistemas ferroviários modernos.

Por fim, os resultados da simulação são fundamentais para demonstrar a arquitetura proposta do novo controlador. A análise dos dados obtidos valida a eficácia das soluções propostas, garantindo que as locomotivas série LE5600 estejam alinhadas com os mais recentes avanços da indústria ferroviária. Este estudo não apenas propõe uma modernização tecnológica, mas também estabelece um novo padrão de desempenho e fiabilidade para o sistema de tração ferroviária.

Abstract

This dissertation presents a study focused on the modernization of traction converter controllers for locomotives, particularly the LE5600 series by Comboios de Portugal, aligned with the company's strategic goals to mitigate the obsolescence of critical components. The need for upgrade aims not only to extend the systems' lifespan but also to enhance energy efficiency and operational adaptability of locomotives.

This work addresses the advanced control of a three-phase induction motor in two main aspects. Firstly, it involves optimizing the control of the specific three-phase converter used in the LE5600 series, tuning voltage and current settings to maximize performance and reliability. Secondly, it encompasses the control of a 400V AC motor in a motor-generator test bench, extending the scope beyond specific locomotive applications.

The dissertation focuses on detailed simulation of proposed new configurations. Simulation results are pivotal, demonstrating significant improvements in traction control and energy efficiency through the adoption of modern technologies such as Insulated-Gate Bipolar Transistors (IGBTs) replacing obsolete Gate Turn-off Thyristors (GTOs). This approach not only overcomes current technical limitations but also facilitates better integration with modern railway systems.

Finally, simulation outcomes are essential for showcasing the proposed architecture of the new controller. Analysis of these results validates the effectiveness of proposed solutions, ensuring that the LE5600 series locomotives align with the latest advancements in the railway industry. This study not only proposes technological modernization but also sets a new standard for performance and reliability in railway traction systems.

Objetivos de Desenvolvimento Sustentável (ODS) das Nações Unidas

Os Objetivos de Desenvolvimento Sustentável (ODS) da ONU, conforme descritos por Ban Ki-moon, ex-Secretário-Geral das Nações Unidas, representam uma visão comum para a humanidade e um compromisso coletivo entre líderes globais e as suas populações. Esses 17 objetivos que se incluem na Agenda 2030, implementada em 2016, visam orientar os esforços globais na direção do desenvolvimento sustentável. São abordadas questões como a erradicação da pobreza, a melhoria da saúde e da educação, a redução das desigualdades e o estímulo ao crescimento económico, bem como o combate às alterações climáticas, tendo em conta a preservação dos ecossistemas.

O trabalho efetuado nesta dissertação, contribui para alguns destes objetivos, destacando-se o Objetivo nove, *Indústria, Inovação e Infraestruturas* e o Objetivo onze, *Cidades e Comunidades Sustentáveis*.

A *Indústria, Inovação e Infraestruturas* inclui garantir a inovação e a presença de infraestruturas sustentáveis na indústria, permitindo a atualização e modernização, com foco na sustentabilidade. A procura e substituição de componentes mais atuais é, neste caso, acompanhada pela necessidade de aumento da eficiência, das locomotivas, tornando-as meios de transporte mais sustentáveis.

O objetivo, *Cidades e Comunidades Sustentáveis*, visa, proporcionar o acesso a sistemas de transporte seguros, de custo reduzido, sustentáveis e acessíveis para todos, melhorando a segurança rodoviária através da expansão da rede de transportes públicos. O investimento de recursos e a pesquisa na ferrovia, tal como é feito nesta dissertação, alinha-se com este objetivo promovendo intrinsecamente a utilização dos transportes públicos, nomeadamente os comboios.

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*“The difference between theory and practice
is smaller in theory than it is in practice.”*

Albert Einstein

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Symbols and Abbreviations

AC	Alternating Current
BJT	Bipolar Junction Transistors
DC	Direct Current
DFOC	Direct Field-Oriented Control
DTC	Direct Torque Control
EMF	Electromotive force
EMU	Electric Multiple Units
FOC	Field-Oriented Control
FSM	Finite State Machine
GTO	Gate Turn-Off Thyristors
IFOC	Indirect Field-Oriented Control
IGBT	Insulated-Gate Bipolar Transistor
IIR	Infinite Impulse Response
MMF	Magnetomotive Force
MOSFET	Metal-Oxide-Semiconductor Field-Effect Transistor
NPN	Negative-Positive-Negative
PWM	Pulse Width Modulation
PI	Proportional-Integral
RMSF	Reference Modulating Signal Frequency
RPM	Revolutions Per Minute
SOA	Safe Operating Area
SPWM	Sinusoidal Pulse Width Modulation
SVM	Space Vector Modulation
TCU	Traction Control Unit
VCE	Collector-to-Emitter Voltage
V/F	Voltage over Frequency
ZOH	Zero-Order-Hold

Chapter 1

Introduction

This dissertation proposes a modernized controller for the three-phase inverter of the LE5600 locomotives used by Portuguese Railways. The aim is to enhance operational efficiency and support the principles of the circular economy by refurbishing the Traction Control Unit, thereby improving the overall performance of the traction converter system.

1.1 Context

The railway industry stands as a bastion of technological advancement, continuously embracing innovations to optimize infrastructure and operational efficiency. However, amidst this rapid evolution, certain critical components within the 5600 series locomotives, notably the traction converter controller, have fallen behind contemporary technological standards. These locomotives, which began service in 1993, now have components nearing the end of their life. After more than 30 years of operation, there is significant potential for upgrades, particularly in the Traction Control Unit, due to the substantial advancements in technology during this period. This controller, reliant on outdated semiconductor technology like Gate Turn-Off Thyristors (GTOs), requires urgent upgrades due to its constraints on efficiency and integration within modern railway infrastructure. Modernizing this controller can enhance operational efficiency and extend the service life of the locomotives.

1.2 Main Goals of the Project

The main goals of this project are as follows:

- Conduct a comprehensive literature review, exploring what makes up the Traction Control Unit and the different control methods that can be employed in it;
- Validate findings and knowledge acquired through software simulations, creating a working TCU;
- Develop a physical prototype of the TCU in a laboratory environment;

By establishing these goals, the execution of this project in phases enables a linearized and organized development process.

1.3 Structure of the Dissertation

The structure of this dissertation is divided in 7 chapters

- Chapter 1: Introduction: Sets the context, highlights the problem statement, and outlines the project goals.
- Chapter 2: State of the Art: Explores the required knowledge that is going to provide thorough comprehension of the current and future Traction Converter.
- Chapter 3: Theoretical Fundamentals: Details the information and theory required for this work.
- Chapter 4: Validation and Simulation: Focuses on a proposed modernization plan, implications, and simulation models.
- Chapter 5: Implementation: Outlines strategies for integrating upgrades into the 5600 series locomotives, demonstrating the approaches made for this goal.
- Chapter 6: Discussion of the results obtained along this work, as well as comparing the ones obtained through simulation and the ones obtained experimentally.
- Chapter 7: Final discussion, how this work accomplishes its goals and future work.

This structure guides the comprehensive exploration of existing limitations, proposed modernization strategies, and potential implications and future applications of these upgrades within the locomotive systems.

Chapter 2

Literature Review

2.1 Overview of traction control

The role of traction control systems in locomotives and trainsets is essential for ensuring safe, efficient, and versatile railway operations globally. These systems, integral to managing speed, safety, and performance across various terrains, demand a thorough understanding of their technical aspects, advantages, disadvantages, and specific applications.

The objective of this dissertation is focused on the series LE5600 from *Comboios de Portugal*, which is a locomotive used in both passenger and cargo transportation.

In figure 2.1, the LE5600 locomotive series from *Comboios de Portugal* is depicted.



Figure 2.1: LE5600 series locomotive retrieved from [1].

This model features a 5600 kW nominal power consumption with a top speed of 200 km/h, and traction control system built by Siemens [1].

Alongside the LE5600 series, there are other similar locomotives like the LE4700 series, more suitable for cargo transport. This train was also constructed with Siemens' collaboration [1][2].

Railways are not solely reliant on locomotives. While these are generally employed for long-distance transport or heavier loads, other vehicles, such as Electric Multiple Units (EMU), can cater to different requirements.

However, the CPA4000 series, commonly referred to as the *Alfa Pendular*, is an EMU versatile and capable of extended journeys, spanning almost the entire country. EMUs are frequently utilized for regional and intercity transport, albeit not exclusively for short distances like some models. They generally accommodate lower weight capacities and are slower compared to locomotives.

While both the LE5600 and the *Alfa Pendular* attain a top speed of 200 km/h and cater to long-distance travel, their propulsion systems differ. In the case of the standard trainset, a distinct characteristic lies in the arrangement of motors per passenger car along the entire length. For instance, the *Alfa Pendular* is equipped with 2 motors per passenger car in 4 out of the 6 total cars. Conversely, in locomotives such as the LE5600, the motors are positioned at the locomotive itself, with 4 motors situated in the locomotive's four axles [1][2].

Even though these two models appear different due to their motor disposition, they share a similar traction converter architecture for the traction control inverter. What can be found in both of these trains is similar to what figure 2.2 illustrates.

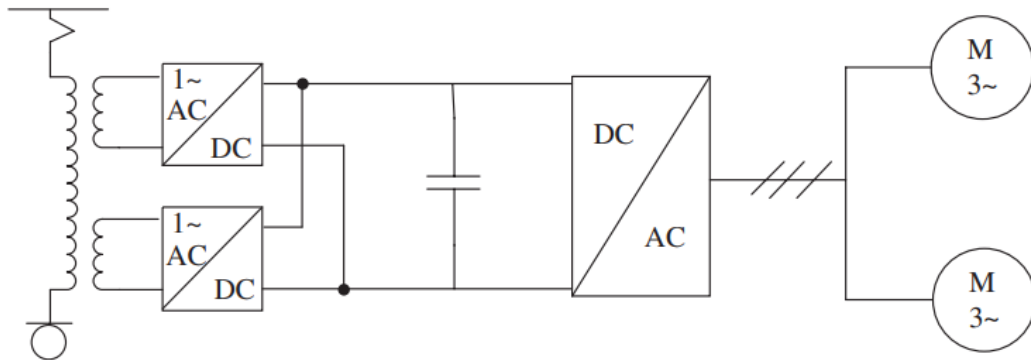


Figure 2.2: Electric drive configuration retrieved from [3].

The primary focus of this dissertation lies on the AC/DC inverter positioned to the right, responsible for converting DC current — originating from the AC power obtained through the catenary network to feed the DC bus — back into three-phase AC current to power the motors. Within this specific inverter, several semiconductors referred to as GTOs govern the power supply to the motor, regulating its operation [2]. A detailed illustration depicting the arrangement of GTOs within the inverter is presented in figure 2.3.

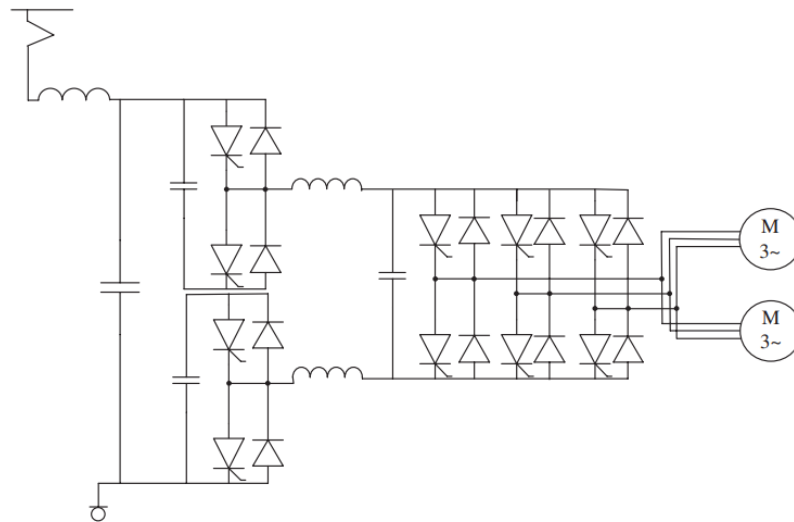


Figure 2.3: GTO disposition inside the inverters for a DC catenary retrieved from [3].

2.2 Semiconductor technology

In this section, it will be discussed different options of semiconductors that can be found inside the Inverter of a locomotive.

2.2.1 Gate Turn-off Thyristors (GTO)

A gate turn-off thyristor (known as a GTO) is a three terminal power semiconductor device. These are devices that have the ability for full control of on- and off-states via the control terminal (gate). And unlike standard thyristors, a GTO is designed to turn-off by applying a negative gate signal. Meaning, applying a positive gate signal to its gate terminal can turn-on a GTO. GTOs are of two types: asymmetrical and symmetrical. In the case of the Series LE5600, the GTOs are symmetrical [2] since they allow for an equal forward and reverse blocking capability, which is something the asymmetrical GTOs are not able to accomplish [4].

Following, is figure 2.4 of the structure of a GTO.

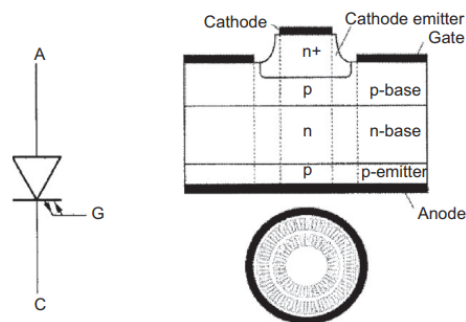


Figure 2.4: On the left is the GTO symbol and on the right is the GTO's structure retrieved from [4].

The GTO functions akin to a standard thyristor during its activation phase, where electron emission from the cathode is induced by injecting hole current through the gate. Unlike a conventional thyristor, the GTO comprises numerous narrow cathode elements intertwined with the gate electrode. This design results in a larger initially activated area and quicker conduction onset, handling high turn-on di/dt . To deactivate a GTO, the gate is reverse-biased, extracting holes from the anode, leading to a voltage drop in the p-base, ultimately halting electron injection. During deactivation, current flows through distant areas, creating high-density filaments. Quickly extinguishing these filaments is crucial to prevent device failure due to localized heating. Although applying higher negative gate voltage can aid in this, the gate-cathode junction's breakdown voltage sets a limit[4].

These GTOs are used in the current traction converter of the LE5600 series locomotive in the regulation of the firing circuit. They serve as semiconductor devices within the inverter, responsible for modulating the width of firing pulses. The firing circuit receives reference values for the stator frequency and motor voltage, processing this data to determine the necessary trigger signals for the GTOs. These signals are transformed into firing pulses and then transmitted to each of the GTOs within the inverter. The GTOs play a critical role in controlling the power supplied to the motor, enabling precise modulation of pulse width to adjust speed, current, and other operational parameters as required. Additionally, this component is responsible for managing the line-side AC-DC converter either during dynamic braking and during fully controlled rectifier mode from this block [2].

2.2.2 Insulated-Gate Bipolar Transistor

The Insulated-Gate Bipolar Transistor (IGBT) emerges as a formidable replacement candidate for the outdated GTOs in the LE5600 series locomotive. In fact, the upgrade of the semiconductor technology was made in one LE5600 by Nomad Tech, the company supporting this dissertation.

Similar to its predecessor, the IGBT functions as a three-terminal power semiconductor switch, regulating electrical energy flow. What sets the IGBT apart is its ability to bridge the speed gap present in conventional devices, such as gate turn-off thyristors (GTOs). It incorporates the finest attributes of power Bipolar Junction Transistors (BJTs) and Metal-Oxide-Semiconductor Field-Effect Transistors (MOSFETs) and like the MOSFETs the firing signal is not in current, but in voltage. This results in superior on-state characteristics, heightened switching speed, and simplified drive circuits, outclassing BJTs in efficiency. Meanwhile, IGBTs intertwine low on-state resistance akin to MOSFETs, elevating the overall performance of power devices. Their widespread commercial availability has solidified their preference for high-voltage applications, courtesy of lower conduction losses [4].

The advantages of utilizing IGBTs in converters are extensive, encompassing simplicity and modularity in converter design, streamlined gate drive, elimination of snubber circuits due to a square safe operating area, diminished switching losses, enhanced protection against over-current and short circuit faults, galvanic isolation of modules, and simplified mechanical construction of

power converters. Recognized under various names, IGBTs require meticulous design optimization to ensure reliability and superior performance across diverse power electronic systems [4]. Figure 2.5 elucidates the structure and equivalent circuit model of the IGBT.

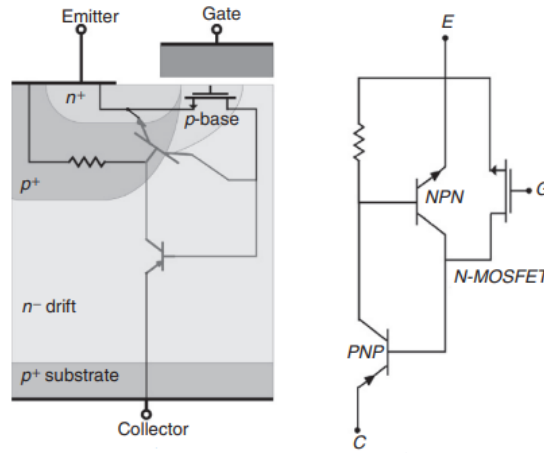


Figure 2.5: IGBT's structure and equivalent circuit model adapted from [4].

Bipolar transistors, albeit featuring lower input capacitance in comparison to IGBTs, rely on current for operation, necessitating drive circuitry capable of generating high and sustained input currents. In contrast, the IGBT's high input impedance stands as a distinct advantage over its bipolar counterpart. Nevertheless, the IGBT also contends with elevated input capacitance, demanding drive circuitry capable of swift charging and discharging of the input capacitor during transition periods. Despite this, the IGBT showcases performance akin to bipolar power transistors, particularly in maintaining low saturation voltage across the operational temperature range. To reliably facilitate output switching, the IGBT mandates a gate-emitter voltage transition ranging from -8 to 15V when used in traction control systems[4].

IGBTs come equipped with protective features, including short circuit protection and active clamping mechanisms. Short circuit protection involves the limitation of current by the IGBT when over-current or short circuit faults occur. Despite this, the fault current might surpass the nominal IGBT current significantly. Therefore, a swift turn-off of the IGBT becomes crucial immediately after a fault occurs to mitigate potential damage [4]. Additionally, active clamping is employed during the ON-OFF transition phase, monitoring the VCE (Collector-to-Emitter Voltage). This monitoring serves as a safeguard against sudden high-current cutoffs or overvoltage instances attributed to the presence of parasitic inductances. In such situations, the blocking action extends over a period to ensure that the cutoff remains within the Safe Operating Area (SOA) limits, preventing potential damage to the IGBT.

As mentioned earlier, one advantage of the IGBT is its possession of a square Safe Operating Area (SOA), signifying an idealized and uniform distribution of safe operating limits across both voltage and current axes in a graphical representation. The square-shaped SOA denotes the device's capability to uniformly withstand diverse combinations of voltage and current stresses

across its operational range. This delineates the boundary within which the IGBT can be safely operated without risking destructive failure. The Safe Operating Area serves as a crucial determinant, defining the limits of current and voltage under which the IGBT can function reliably [4]. Figure 2.6 illustrates the specific Safe Operating Area (SOA) characteristic to the IGBT, providing a visual representation of its operational limits for enhanced comprehension.

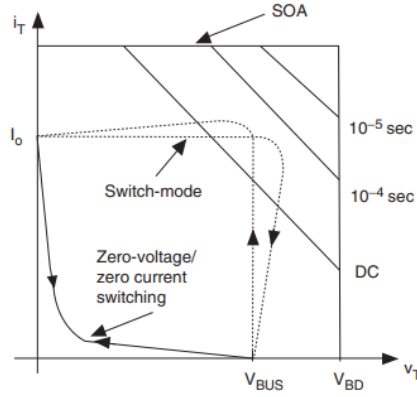


Figure 2.6: IGBT safe operating area retrieved from [4].

The Safe Operating Area (SOA) of an IGBT undergoes changes under various operating conditions. Notably, under shorter switching times, the rectangular SOA diminishes due to increased on-time duration. Thermal limitations play a crucial role in determining the SOA, with the lower limit dictated by direct current (dc) operating conditions, as an example, the DIM1200ASM45 traction IGBT, has a DC maximum current of 1200A and a peak of 2400A for 1ms[5]. The figure presented earlier also outlines the device's switching loci, delineating hard switching (dashed lines) and zero voltage or zero current switching (solid lines). In particular, the excursion is more extensive for switch-mode hard-switching applications compared to the soft-switching scenario, necessitating a broader SOA for hard-switching applications, such as the semiconductors used in a traction power converter. Presently, IGBTs are predominantly optimized for hard-switching applications. However, in soft-switching applications, optimizing the conduction losses of IGBTs may lead to a smaller SOA. For instance, adjusting the p-base doping to achieve a lower threshold voltage and subsequently a reduced forward voltage drop is feasible. Nevertheless, in hard-switching scenarios, the priority leans towards meeting SOA requirements over managing forward voltage drop and switching time. Consequently, reducing the p-base resistance to fulfill SOA requirements results in a higher threshold voltage, thereby increasing channel resistance and forward voltage drop [4].

2.3 Traction Control in Motor Systems

The driving of three-phase induction motors, such as the one utilized in this dissertation, significantly relies on the application of traction control methods to ensure efficient and stable perfor-

mance. Various control methods always present a specific range of benefits and limitations, which need thorough analysis to ensure that the chosen method for each system is the most suitable and effective. This section will conduct a review of the most common control methods, addressing their respective advantages and disadvantages.

2.3.1 Scalar Control

Scalar control method is a straightforward approach that focuses on varying the voltage applied to the motor in proportion to the load or desired speed. Consequently, it represents an easily implementable and cost-effective method. However, due to these characteristics, it is not particularly precise concerning speed control, resulting in lower efficiency in partial load situations as well.

2.3.2 V/F Control

The V/F control method, a part of scalar control, relies on maintaining a constant ratio between voltage and frequency, ensuring their proportional relationship. Consequently, an increase in voltage prompts a corresponding increase in frequency, while a decrease in voltage results in a decrease in frequency. This sustained correlation adapts to variations in motor speed as the rotor of an induction motor follows, with slip, the stator magnetic field, rotating at a synchronous speed proportional to the applied stator frequency. Under constant V/F operation in an induction motor, torque remains independent of the stator frequency, rotor current remains unaffected by the stator frequency, and the synchronous speed maintains a proportional relationship with the stator frequency. Figure 2.7 presents a brief and simple visual representation of the V/F control method [6].

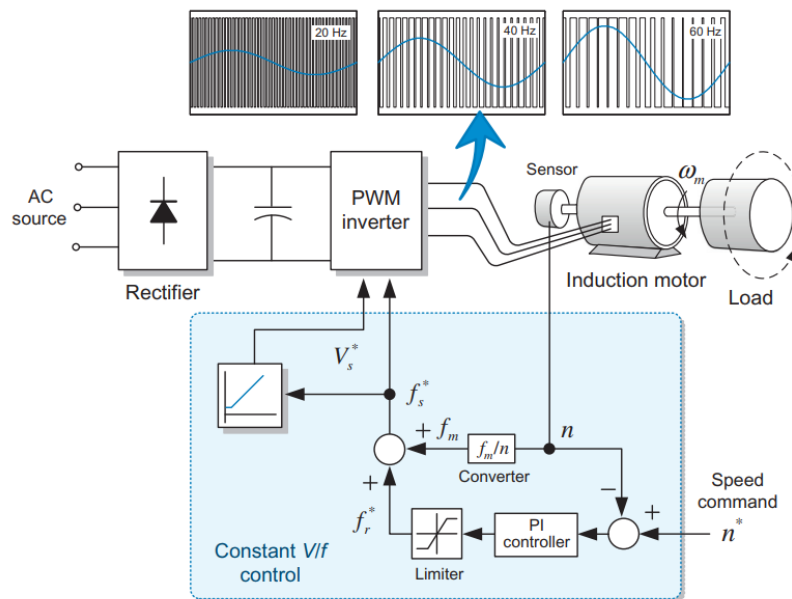


Figure 2.7: Speed control by adjusting the slip frequency with constant V/F control retrieved from [6].

2.3.3 Vector Control

The vector control method is employed in systems that demand higher precision in speed control, offering swift responses and torque variations, thereby providing greater energy efficiency compared to scalar control methods. Although it requires the utilization of high-performance microprocessors, which is not the case in scalar methods, vector control has gained more prominence and is becoming increasingly popular.

2.3.4 Field-Oriented Control

Vector Control methods, such as Field-Oriented Control (FOC), intricately manage instantaneous torque by precisely manipulating both the direction and magnitude of flux and current. The flux direction is typically regulated through the flux vector, denoted as ϕ . FOC mirrors the torque control found in separately excited DC motors [6].

In the context of induction motor systems, the interplay between two fundamental components, the stator and rotor, is pivotal. The stator, a static element, induces flux within the motor, while the rotor, upon induction, initiates rotation, thereby propelling the motor forward. Within the stator lie three-phase windings (i_{as} , i_{bs} , i_{cs}), essential for deriving the Flux-Producing Current (I_ϕ) and the Torque-Inducing Current (I_t), integral components utilized in the FOC method [6].

The transformations that happen between the moment that the three-phase currents (i_{as} , i_{bs} , i_{cs}) are fed to the motor and the derived Flux-Producing Current and the Torque-Inducing Current are known as the Clarke and Park Transformations. The Clarke transformation describes the transformation of the reference 3-phase system (a , b , c) to a 2-phase coordinate reference of, $\alpha\beta$ resulting in a set of rectangular coordinates fixed to the stator of the motor. It is important to add that the full transformation results in $\alpha\beta 0$ components, but the third component can be discarded since in this 3-phase system the sum of the three currents is 0. The α axis is aligned with the first phase (phase a) and the β axis represents the imaginary axis, 90 degrees in delay. This means that the coordinate system $\alpha\beta$ is a complex number coordinate system with $\alpha\beta$ being real numbers [7]. Depicted in figure 2.8 is Clarke Transformation's axis.

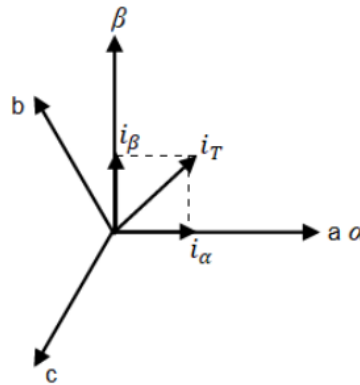


Figure 2.8: Clarke Transformation's Axis retrieved from [7].

This transformation occurs to currents, but can also happen for the voltage and flux. I_T corresponds to the sum of the currents i_a , i_b and i_c .

The magnetomotive force (MMF) produced in the stator of a 3-phase motor is the sum of the magnetomotive forces produced in each phase, being each phase spaced in a 120 electrical degrees angle. The MMF for the $\alpha\beta$ from the magnetomotive forces of the 3-phase coordinate system (a , b , c) are as shown:[7]

$$F_\alpha = F_a \cos(0) + F_b \cos\left(\frac{2\pi}{3}\right) + F_c \cos\left(\frac{4\pi}{3}\right) = F_a - \frac{1}{2}F_b - \frac{1}{2}F_c \quad (2.1)$$

$$F_\beta = F_a \sin(0) + F_b \sin\left(\frac{2\pi}{3}\right) + F_c \sin\left(\frac{4\pi}{3}\right) = 0 + \frac{\sqrt{3}}{2}F_b - \frac{\sqrt{3}}{2}F_c \quad (2.2)$$

The pulsating MMF produced for a single instance is calculated as,

$$MMF = Ni \quad (2.3)$$

From this the following equations can be developed.

$$i_\alpha = i_a \cos(0) + i_b \cos\left(\frac{2\pi}{3}\right) + i_c \cos\left(\frac{4\pi}{3}\right) \quad (2.4)$$

$$i_\beta = i_a \sin(0) + i_b \sin\left(\frac{2\pi}{3}\right) + i_c \sin\left(\frac{4\pi}{3}\right) \quad (2.5)$$

$$i_\alpha = \left(i_a - \frac{1}{2}i_b - \frac{1}{2}i_c\right) \quad (2.6)$$

$$i_\beta = \left(0 + \frac{\sqrt{3}}{2}i_b - \frac{\sqrt{3}}{2}i_c\right) \quad (2.7)$$

With these equations, the matrix to represent the Clarke's transformation can be created.

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.8)$$

And the generalized inverse matrix, useful for the inverse Clarke transformation, is given by:

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & 0 \\ -\frac{1}{3} & \frac{1}{\sqrt{3}} \\ -\frac{1}{3} & -\frac{1}{\sqrt{3}} \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (2.9)$$

The Park transformation while being similar to the Clarke transformation, it is used as an extension of the latter, since it picks up where Clarke's transformation leaves converting the $\alpha\beta$ coordinate reference into $d-q$ coordinate reference resulting in a continuous-like waveform [7]. Depicted in figure 2.9 are the Park transformation's axis:

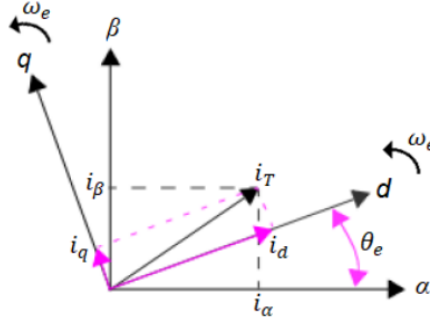


Figure 2.9: Park Transformation's Axis retrieved from [7].

The $d-q$ system rotates in synchronism with the stator's rotating magnetic field of the motor, meaning it has the same velocity ω_e . Knowing this, the rotating magnetic field can be split into two components of constant values, being the axis of the rotor d and, perpendicularly, the axis q . All the values present in the rotor (current, voltage, and flux) are transformed into similar ones that rotate at the same rate as the rotor distanced with an angle of θ_e . The matrix representation of the Park transformation that allows the transformation from the abc coordinates into $d-q$ coordinates is [7]:

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & \cos(\theta_e - \frac{2\pi}{3}) & \cos(\theta_e - \frac{4\pi}{3}) \\ \sin(\theta_e) & \sin(\theta_e - \frac{2\pi}{3}) & \sin(\theta_e - \frac{4\pi}{3}) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.10)$$

This matrix representation can also be obtained from the $\alpha\beta$ coordinate system.

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & \sin(\theta_e) \\ -\sin(\theta_e) & \cos(\theta_e) \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (2.11)$$

The respective inverted matrix, useful for the inverse Park transformation, is given by,

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & \sin(\theta_e) \\ \cos(\theta_e - \frac{2\pi}{3}) & \sin(\theta_e - \frac{2\pi}{3}) \\ \cos(\theta_e - \frac{4\pi}{3}) & \sin(\theta_e - \frac{4\pi}{3}) \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (2.12)$$

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} \cos(\theta_e) & -\sin(\theta_e) \\ \sin(\theta_e) & \cos(\theta_e) \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (2.13)$$

Moreover, the control mechanisms employed in vector control for induction motors fall into two primary categories: direct vector control (DFOC) and indirect vector control (IFOC). These methods diverge in their approaches to determining the rotor phase angle. Direct vector control

directly acquires the flux angle by detecting the rotor flux position, employing either measurement or estimation techniques [6][8]. Conversely, indirect vector control, not reliant on the flux, employs the slip angular velocity to establish the flux angle. This velocity serves to segregate the stator currents into desired values for both flux-producing and torque-producing currents [6]. The diagram in figure 2.10 illustrates the currents of an induction motor.

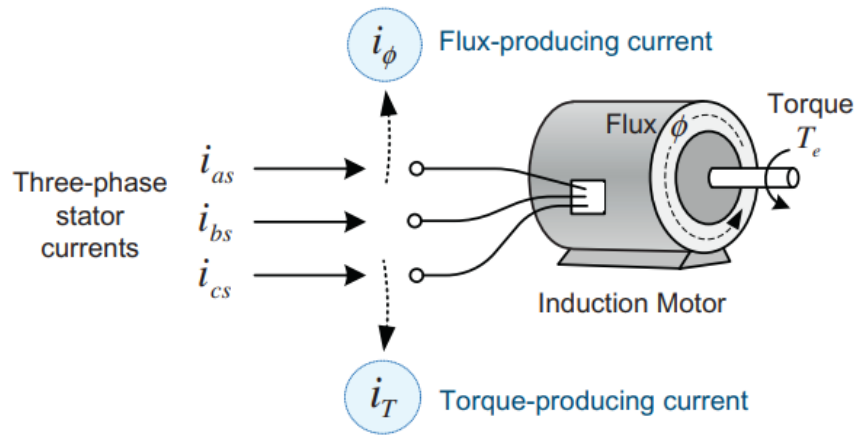


Figure 2.10: Currents of an induction motor retrieved from [6].

Nevertheless, these currents adhere to specific requirements. The steadfast maintenance of a 90-degree angular separation between the flux-producing and torque-producing currents stands as the primary directive. Ensuring autonomous manipulation of both currents — flux-producing and torque-inducing — remains imperative. Moreover, the attainability of instantaneous control over the torque-producing current reinforces the essence of this control methodology.[6]

In alignment with conventional practices, the assignment of the d-axis stator current as the flux-producing current and the q-axis stator current as the torque-producing current underscores precision. Accurately mirroring the flux-producing current demands precise alignment of the d-axis current with the actual position of the flux vector λ , elucidated in figure 2.11 [6].

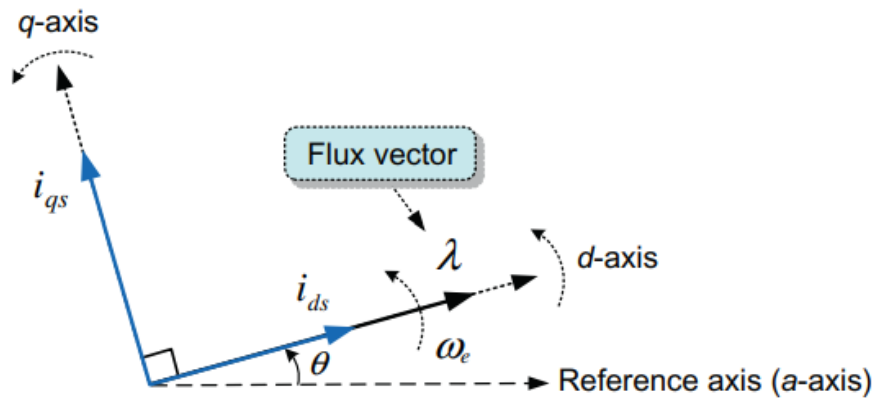


Figure 2.11: Assignment of d - q axes based on the flux vector λ retrieved from [6].

Due to the orthogonality of the d-axis stator current and q-axis current, there exists no mutual influence (cross-coupling) between i_{eqs} and i_{eds} , and vice versa. This independence facilitates separate control over the torque-producing and flux-producing currents. Additionally, the existence of a current controller enables instantaneous control over the torque-producing current [6].

The practical implementation of vector control is elucidated in figure 2.12:

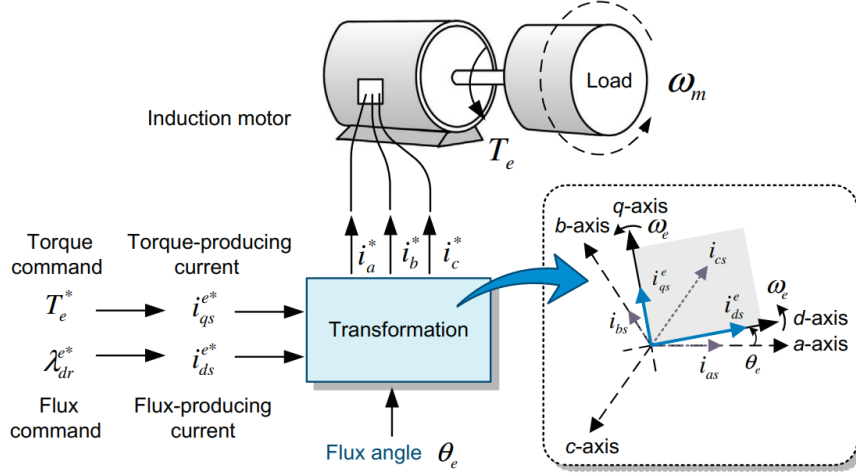


Figure 2.12: Instantaneous torque control method of an induction motor retrieved from [6].

In the conventional vector control system for an induction motor, the flux magnitude typically maintains a steady level at the rated operational value, except in instances of high-speed operations exceeding the specified speed limit. Commencing from the rated flux command λ_{eds} , the essential flux-producing current command, i_{eds} , is derived. The determination of the torque-producing current command, i_{eqs} , is predicated on the output torque command, T_e , required to drive a designated load. These designated command currents, i_{eds} and i_{eqs} , operate within the synchronous rotating reference frame. The conversion back into these d-q current commands holds pivotal significance in ascertaining the three-phase stator current commands. This transformation necessitates familiarity with the flux vector angle, commonly known as the *flux angle*. Swiftly aligning the three-phase stator currents with these current commands facilitates instantaneous control over the output torque of the induction motor [6].

2.3.5 Direct Field Oriented Control

Direct Field Oriented Control or DFOC in short is one of the two categories of Field Oriented Control mentioned previously. This approach distinguishes itself by deriving the flux angle directly from the flux. Due to this, it is neither a very common nor practical solution since this is accomplished through the utilization of the rotor flux angle, calculated via the equation 2.14, using the stationary d-q rotor flux linkages, provided they are available through detection or estimation methods [6].

$$\theta_e = \tan^{-1} \times \left(\frac{\lambda_{qr}^s}{\lambda_{dr}^s} \right) \quad (2.14)$$

2.3.6 Indirect Field Oriented Control

Indirect Field-Oriented Control (IFOC) is an alternative approach within the realm of Field-Oriented Control (FOC). Unlike Direct FOC (DFOC), IFOC does not rely on directly measuring or estimating the d-q rotor flux linkages. Consequently, these values are not employed directly in determining the rotor flux angle, thereby simplifying the implementation process.

In an induction motor, the slip (s) regulates the division of stator current into flux- and torque-producing components, enabling control over the desired flux and torque outputs [6].

Notably, in vector control, with the d-axis assigned to the rotor flux linkage position, the absence of rotor flux linkage in the q-axis is evident ($\lambda_{qr}^e = 0$). This yields the following rotor voltage equation:

$$v_{qr}^e = R_r i_{qr}^e + p \lambda_{qr}^e + (\omega_e - \omega_r) \lambda_{dr}^e = R_r i_{qr}^e + (\omega_e - \omega_r) \lambda_{dr}^e = 0 \quad (2.15)$$

Consequently, the slip angular frequency (ω_{sl}) required for vector control is derived as:

$$\omega_e - \omega_r = \omega_{sl} = -\frac{R_r i_{qr}^e}{\lambda_{dr}^e} \quad (2.16)$$

By combining ω_{sl} and i_{qr}^e , the slip angular frequency for vector control is expressed as:

$$\omega_{sl} = \frac{L_m R_r}{L_r} \frac{i_{qs}^e}{\lambda_{dr}^e} \quad (2.17)$$

The relation between the d-axis stator current (i_{ds}^e) and the rotor flux linkage (λ_{dr}^e) is established as:

$$\lambda_{dr}^e = \frac{L_m R_r}{L_r p + R_r} i_{ds}^e = \frac{L_m}{1 + T_r p} i_{ds}^e \quad (2.18)$$

$$(T_r = \frac{L_r}{R_r}) \quad (2.19)$$

Further, ω_{sl} is also represented in $d-q$ axes currents as:

$$\omega_{sl} = \left(\frac{1}{T_r} + p\right) \frac{i_{qs}^e}{i_{ds}^e} = \frac{L_m}{1 + T_r p} i_{ds}^e \quad (2.20)$$

Given constant rotor flux linkage, the slip requirement for indirect vector control is finally obtained as:

$$\omega_{sl} = \frac{1}{T_r} \frac{i_{qs}^e}{i_{ds}^e} \quad (2.21)$$

This relation underscores the need to set the slip angular frequency to specific values to divide stator currents into desired $d-q$ axes currents. Subsequently, the rotor flux angle can be determined by integrating the rotor velocity (ω_r) with the slip angular frequency [6].

$$\Theta_e = \int \omega_e dt = \int (\omega_{sl} + \omega_r) dt \quad (2.22)$$

The IFOC method, by adjusting the slip angular frequency instead of directly using rotor flux linkage, achieves the desired division of $d - q$ axes currents. However, a drawback of IFOC is its reliance on rotor speed for obtaining the rotor flux angle, typically measured using an encoder [6].

2.3.7 Direct Torque Control

Direct Torque Control (DTC) is another vector control method applied in railway systems. DTC, also known as direct self-control, refers to the direct control of flux and torque [9]. This method is simpler to implement than Field Oriented Control (FOC) but has certain drawbacks, such as high torque and flux ripple, variable switching frequency, and performance degradation at low speeds. Additionally, it requires a high sampling frequency due to the need for digital implementation of hysteresis [10].

Unlike DFOC, the DTC scheme is characterized by the absence of [11]:

1. PI regulators;
2. Coordinate transformations;
3. Current regulators;
4. PWM signal generators.

DTC controls the stator flux and torque using hysteresis comparators instead of the PI controllers found in FOC. This is achieved using a lookup table of voltage vectors. Accurate estimation of the stator flux and torque is essential for this method [12].

The stator flux can be calculated using the following equation:

$$\phi = \int_0^t (V_s - I_s \cdot R_s) dt \quad (2.23)$$

The torque can be calculated by:

$$T = \frac{3}{2} \cdot p \cdot \phi_s \cdot V_s \quad (2.24)$$

Where p are the pole pairs, ϕ is the stator flux and V_s the stator voltage. The basic DTC scheme is illustrated in Figure 2.13.

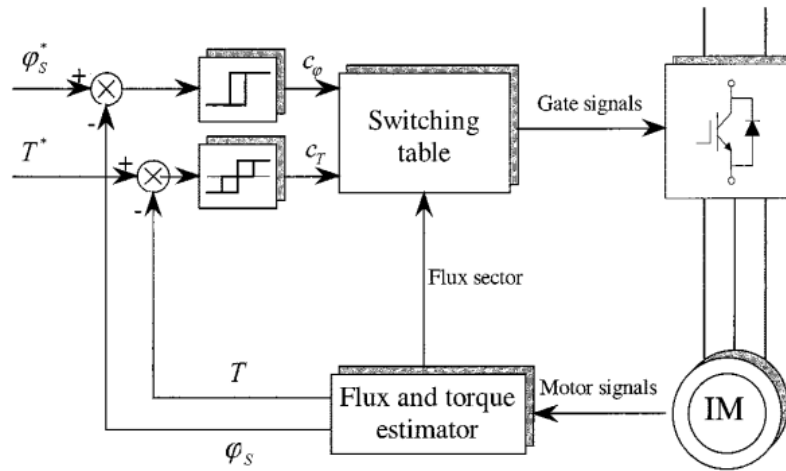


Figure 2.13: Direct Torque Control Scheme retrieved from [11].

Here, T^* represents the torque reference, T is the torque estimation, ϕ^* is the reference stator flux magnitude, ϕ is the estimated stator flux magnitude and C_T as well as C_ϕ are the outputs from the hysteresis comparators. These values are compared to create an error input into the hysteresis comparators, similar to a Proportional-Integral Controller [11]. Figure 2.14 displays the voltage vectors utilized in the switching table.

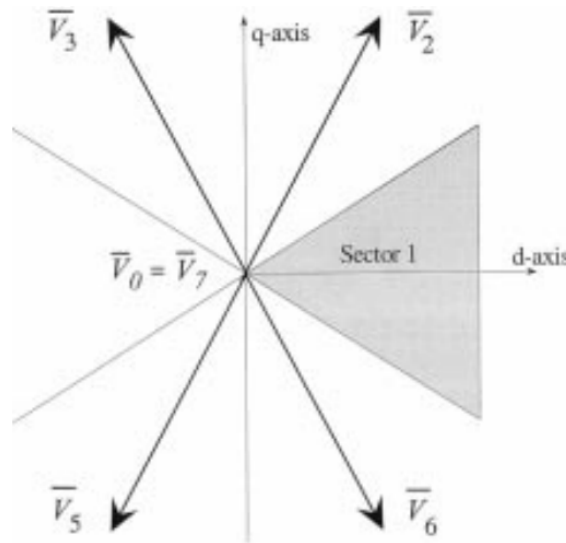


Figure 2.14: Direct Torque Control Voltage Vectors retrieved from [11].

The appropriate voltage vector is selected based on the switching table shown in Table 2.1.

Table 2.1: Direct Torque Control Voltage Vector Switching adapted from [11].

Sector		1	2	3	4	5	6
$C_\phi = -1$	$C_T = -1$	$\overline{V}_2$	$\overline{V}_3$	$\overline{V}_4$	$\overline{V}_5$	$\overline{V}_6$	$\overline{V}_1$
	$C_T = 0$	$\overline{V}_7$	$\overline{V}_0$	$\overline{V}_7$	$\overline{V}_0$	$\overline{V}_7$	$\overline{V}_0$
	$C_T = +1$	$\overline{V}_6$	$\overline{V}_1$	$\overline{V}_2$	$\overline{V}_3$	$\overline{V}_4$	$\overline{V}_5$
$C_\phi = +1$	$C_T = -1$	$\overline{V}_3$	$\overline{V}_4$	$\overline{V}_5$	$\overline{V}_6$	$\overline{V}_1$	$\overline{V}_2$
	$C_T = 0$	$\overline{V}_0$	$\overline{V}_7$	$\overline{V}_0$	$\overline{V}_7$	$\overline{V}_0$	$\overline{V}_7$
	$C_T = +1$	$\overline{V}_5$	$\overline{V}_6$	$\overline{V}_1$	$\overline{V}_2$	$\overline{V}_3$	$\overline{V}_4$

The input quantities are the stator flux sector and the outputs of the two hysteresis comparators. This approach allows a quick torque response but results in undesired ripple in current, flux, and torque due to the absence of torque and rotor speed information in the voltage vector selection algorithm [11].

One way to reduce the current and torque ripple in Direct Torque Control (DTC) schemes is by using a preview technique. This technique calculates the required variation of the stator flux vector to precisely compensate for flux and torque errors during each cycle period. To implement this principle, the control system should be capable of generating any voltage vector, which can be achieved using the space vector modulation technique.

Research has shown that subdividing the cycle period into three equal time intervals significantly reduces torque and current ripple without requiring overly complex switching tables[13]. The increased number of voltage vectors allows for the creation of more precise switching tables, where the selection of voltage vectors can be made based on the rotor speed. These switching tables can be derived from the analysis of equations that link the applied voltage vector to the corresponding torque and flux variations.

By following this operational principle, new voltage vectors can be synthesized beyond those used in the basic DTC technique [11]. This approach offers a refined control mechanism, enhancing the performance and efficiency of the DTC scheme.

2.4 Modulation Techniques

For controlling an induction motor, a variable voltage and variable frequency supply is provided by a Voltage Source Inverter (VSI). The switching pulses for the VSI can be generated using different Pulse Width Modulation (PWM) techniques [14].

In high-power traction drives, such as those used in the LE5600 series locomotives, minimizing high switching losses necessitates operating at low switching frequencies. Given that the current model uses Gate Turn-Off Thyristors (GTOs), the switching frequency is set at 300 Hz [2]. However, this intentional reduction in switching frequency results in notable current and torque ripples. These fluctuations can significantly affect mechanical transmission stress, train comfort, and adherence to standards [15].

Recent advancements in high-speed power semiconductor devices have enabled three-stage inverters to play a crucial role in variable speed AC motor drives [14]. As a result, various advanced PWM techniques have been developed to generate the switching pulses. These techniques offer significant advantages, including higher efficiency due to lower power losses in the switching devices. Additionally, PWM strategies enhance effective DC bus utilization [14].

One such advanced technique is Space Vector Modulation (SVM), which provides better performance compared to traditional Sinusoidal PWM. This section will explain and compare these PWM techniques, highlighting their benefits and applications.

2.4.1 Space Vector Modulation

Space Vector Modulation (SVM) stands as one of the most prominent modulation techniques. This approach is often utilized alongside Synchronous PWM, which will be delved into in the subsequent section. SVM's primary advantage lies in its ability to precisely position pulses, offering an additional dimension for optimizing harmonic performance gains [16].

SVM operates on the principle that a three-phase inverter comprises only eight potential switch combinations. Each of these stationary vectors corresponds to a specific fundamental angular position [16]. The zero vectors (V_0, V_7) are obtained when all the upper or all the lower switches are turned on simultaneously, allowing for a freewheeling state for the three-phase current. The active vectors ($V_1, V_2, V_3, V_4, V_5, V_6$) are achieved when some poles of the inverter are connected to the positive DC bus and the remaining poles to the negative DC bus [17]. Figure 2.15 displays the placement of the eight voltage vectors.

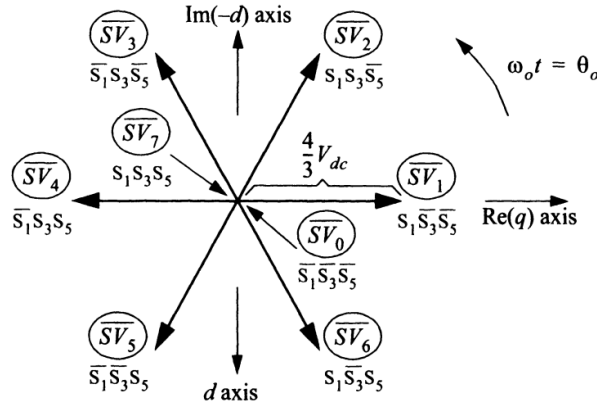


Figure 2.15: Illustration showcasing the positions of the eight possible stationary voltage vectors within the d-q (Re-Im) plane retrieved from [16].

The magnitude of each of these six active vectors can be represented as:

$$V_m = \frac{4}{3} V_{dc} \quad (2.25)$$

Once the stationary vectors are identified, an arbitrary target output voltage vector $\bar{V}_o^*$ can be created at any given moment by summing up (or averaging) a series of these space vectors within a single switching period $\Delta T/2$. This technique offers a nuanced control mechanism, allowing for precise manipulation of the output voltage vectors for optimal motor performance and efficiency [16].

Space Vector Modulation is also widely used with the control method Direct Torque Control (DTC) due to the ripples in the hysteresis band waves when not utilizing SVM. This makes SVM very useful in tandem with DTC to improve its accuracy [18].

2.4.1.1 Disadvantages of Space Vector Modulation

- It involves high cost to implement and its duty cycle parameters are not calculated, compared to other techniques [19].
- Its output voltage contains high total harmonic distortion (THD) and dv/dt , resulting in a large LC filter size to be installed in the inverter output [19].
- SVPWM is complex and computationally intensive, therefore, the PWM is somewhat limited. However, this complexity can be simplified by using a lookup table [19].

2.4.1.2 Advantages of Space Vector Modulation

- It can increase inverter voltage output by up to 15% more than Sinusoidal PWM[17][19].
- SVPWM is a common method applied in AC drive systems [19].
- Within the sampling period, the modulation index can be varied within direct control of inverter current [19].

2.4.2 Sinusoidal PWM

Sinusoidal Pulse Width Modulation (SPWM) is achieved by comparing a high-frequency triangular signal with a low-frequency sinusoidal reference signal, known as the modulation wave. The ratio between the carrier signal amplitude and the sinusoidal reference signal is called the modulation index, which controls the amplitude of the output voltage. In the SPWM technique, multiple output pulses are generated in each half-cycle, and the width of each pulse varies in its middle[20].

This technique consists of two types of switching schemes: Unipolar and Bipolar. In unipolar switching schemes, pulses are generated by comparing carrier signals with two low-frequency reference signals. In bipolar switching schemes, pulses are produced by comparing a high-frequency carrier signal with a single reference signal [20].

In three-phase SPWM, a triangular voltage waveform (VCR) is compared to three sinusoidal control voltages (V_A , V_B , and V_C), which are 120 degrees apart [21]. Figure 2.16 displays Sinusoidal PWM for a two-level Inverter.

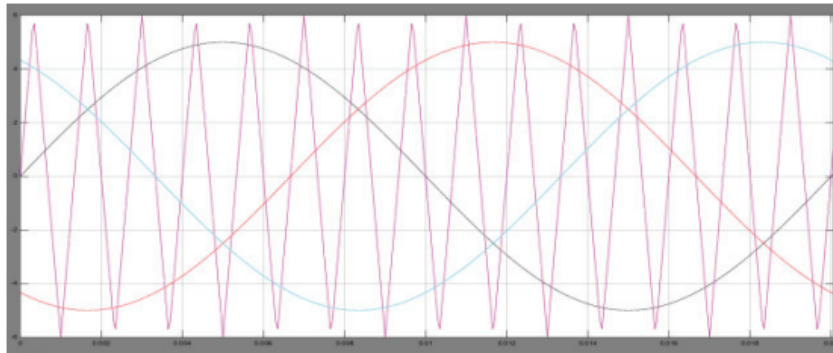


Figure 2.16: Sinusoidal Pulse Width Modulation for a two-level Inverter retrieved from [21].

2.4.2.1 Disadvantages of Sinusoidal PWM

- SPWM technique has high switching losses and requires high input voltage to be supplied from the same DC power supply [19].
- Poor sinusoidal shape of voltage and current control due to high harmonic pollution present in the voltage and current output waveforms [19].

2.4.2.2 Advantages of Sinusoidal PWM

- SPWM is known for having fixed amplitude pulses with varying duty cycles [19].
- The width of a pulse varies depending on the amplitude of the sine wave measured at its center in the SPWM. Gate pulses of SPWM are created by comparing two signals: the reference signal (sinusoidal) and the carrier signal (triangular) [19].
- SPWM is also recognized for its wave shaping capabilities in pulse width modulation [19].

Theoretical Fundamentals

3.1 Current power configuration

The diagram illustrates a closed-loop speed control system for a 50 Hz AC catenary. The system components and their interconnections are as follows:

- AC Catenary** and **50 Hz Transformer** provide input to the **PWM Rectifier**.
- The **PWM Rectifier** is connected to a **DC link** (represented by a capacitor symbol).
- The **DC link** feeds the **SVPWM Inverter**.
- The **SVPWM Inverter** drives the **Motor (M)**.
- The **Motor (M)** outputs **Speed** and provides feedback to the **Speed Controller**.
- The **Speed Controller** outputs the **I_q Reference** and **I_d Reference**.
- The **I_q Reference** and **I_d Reference** are inputs to the **Current Controller**.
- The **Current Controller** outputs **I_q** and **I_d** to the **Current Transformation** block.
- The **Current Transformation** block also receives **Θ** (Theta) from the **Theta (Θ) Estimation** block.
- The **Current Transformation** block outputs **I_a**, **I_b**, and **I_c** to the **SVPWM Inverter**.
- The **Theta (Θ) Estimation** block receives **Speed** as input.
- The **Flux Command** block receives **Speed** as input.

Several loops can be found inside this controller that make up the logic for its control. Each

loop has a specific goal that must be achieved, and they are the flux angle estimation, speed command calculation, torque command calculation, and current control. The output of these control loops are modulation signals that will correspond to specific PWM firing signals for the semiconductor devices. Ultimately, the primary objective of this control structure is to be able to effectively drive the induction motor in an efficient and safe manner. Section 3.1.1 will delve into the characteristics of IMs to gain a better understanding of the requirements for motor control.

3.1.1 Induction Motor parameters

The table 3.1 that follows displays the motor parameters that were provided by CP.

Table 3.1: Induction Motor Parameters.

Parameters	Value
Nominal Voltage (V)	2080
Nominal Current (A)	510
Nominal Power (kW)	1428
Power Factor	0.81
Nominal Speed (rpm)	1490
Nominal Frequency (Hz)	50

Other parameters such as the equivalent circuit characteristics were not provided by CP, making it a useful project in the future to verify these parameters through testing.

3.1.2 Induction Motor model

To gain a comprehensive understanding of the electrical behavior of the motor, the T model representation is commonly employed, providing insights into its steady-state scenario. As depicted in Figure 3.2, the motor can be simplified to a circuit comprising a series of resistances and inductances. The values of these resistances and inductances are determined through measurements and estimations.

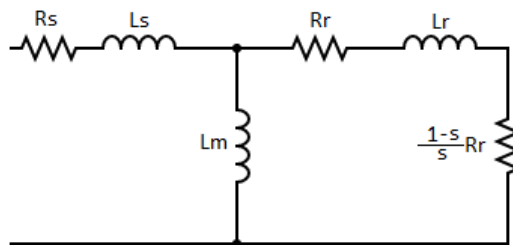


Figure 3.2: T model of an induction motor.

The letter s represents the slip that is calculated by the equation 3.1.

$$s = \frac{n_{sync} - n_{rotor}}{n_{sync}} \quad (3.1)$$

Where n_{rotor} is the rotor speed and n_{sync} is the synchronous speed defined by the equation 3.3

$$n_{sync} = \frac{60f_s}{p} \quad (3.2)$$

Where f_s is the stator frequency and p is the number of pole pairs.

3.1.2.1 Parameter Identification

For the identification of the IM motor parameters such as a squirrel cage, which is the type found in the LE5600, 3 steps are required. The first, is the measurement of the value of the stator resistance. This is done by measuring the resistance between two of the terminals. Since the motor from the laboratory at *Nomad Tech* has a star configuration, by dividing the value measured by two, the stator resistance (R_s) is obtained.

Afterward, there are two tests that are performed: The no-load test and the locked rotor test. The no-load test, involves applying a balanced rated voltage to the stator winding. This test is conducted to gather information about the magnetizing inductance L_m . The motor operates at almost synchronous speed during this test, resulting in a very small slip value ($s \approx 0$). Consequently, with the synchronous speed and the value of s being very small, it leads to a very large value for $\frac{1-s}{s}R_r$. As a result, the components including L_r and R_r are removed due to their significant values. During this test, the motor consumes minimal power, primarily attributed to core losses, friction, and winding losses. The provided voltage establishes the necessary flux I_d current to create the appropriate magnetic field. Ultimately, in the no-load test, all current flows through R_s and L_m [22].

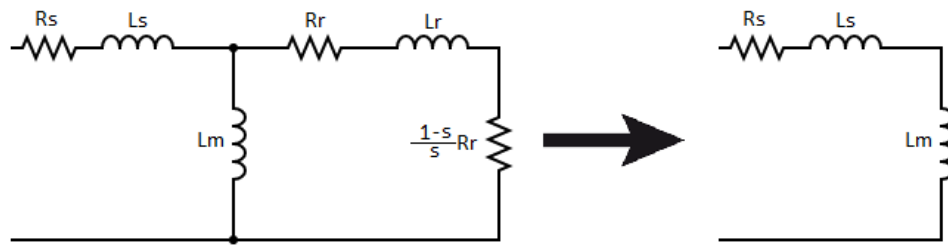


Figure 3.3: Equivalent model of no-load test.

Knowing this, L_m is calculated:

$$L_m = \frac{V_s}{2\pi f_s I} \quad (3.3)$$

The locked rotor test is essential for motor assessment. Similar, to the short circuit test on a transformer, a lower voltage is applied for a short duration to avoid overheating and motor damage. With the rotor immobilized, a slip factor of $s=1$ reduces the resistance $\frac{1-s}{s}R_r$ to 0. This test

reveals leakage impedances and rotor resistance, with the magnetizing inductance L_m far exceeding leakage inductances L_{ls} and L_{lr} . Typically, L_{ls} and L_{lr} should range from 2% to 10% of L_m . Given the rotor's stationary state ($s=1$), the equivalent circuit is simplified accordingly [22].

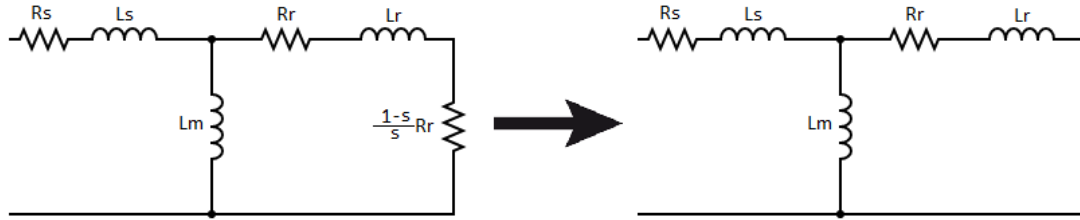


Figure 3.4: Equivalent model of locked rotor test.

3.2 Flux angle

As detailed in subsection 2.3.4, accurately determining the d - q current commands requires the estimation of the flux vector angle. In this control strategy, the Indirect Field Oriented Control method will be applied, using the encoders already equipped in the locomotive. Referring back to equation 2.20, to compute the slip, the q current command is divided by the d current command and then multiplied by the sum of the number of pole pairs (p) and the time constant of the rotor circuit (Tr). Once the slip is obtained, it is added to the rotor speed in radians per second and integrate the resulting sum. It's crucial to limit this integration, so that the value of θ remains within the range of 0 to 2π .

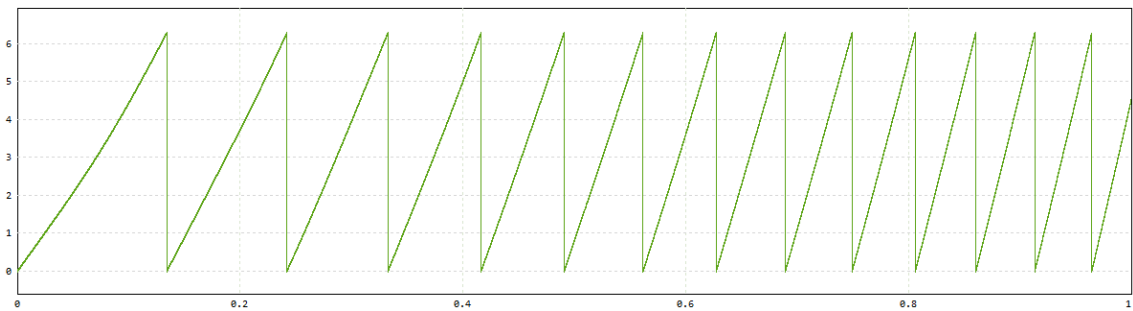


Figure 3.5: Typical θ curve.

As perceived from figure 3.5 and as expected from the equation 2.22, as the speed of the rotor is increased, the frequency of the theta angle steadily increases until reaching close to the nominal frequency. The slip value can vary depending on the speed of the motor and can even be 0 if the motor is in generator mode and the electrical angle is the same as the rotor angle when it is not producing any torque.

3.3 Flux and Torque Commands

To effectively operate the locomotive, the driver must convey information to the motor regarding the desired behavior. This involves providing reference values for speed to the controller, which then translates these into the d-q currents supplied to the motor.

These reference values can be obtained directly from the motor parameters, such as the reference speed, but references such as the I_d current reference need to be calculated. The I_d current reference is obtained using the equation 3.4:

$$I_{d_ref} = \frac{\lambda_{dr}}{L_m} \quad (3.4)$$

The value of the rotor flux linkage can be retrieved from the equation 2.18.

The I_q current reference is defined by the difference between the reference speed defined by the locomotive driver and the current motor speed. This reference is used by the PI controller to output the correct current signal in order to achieve the reference inserted.

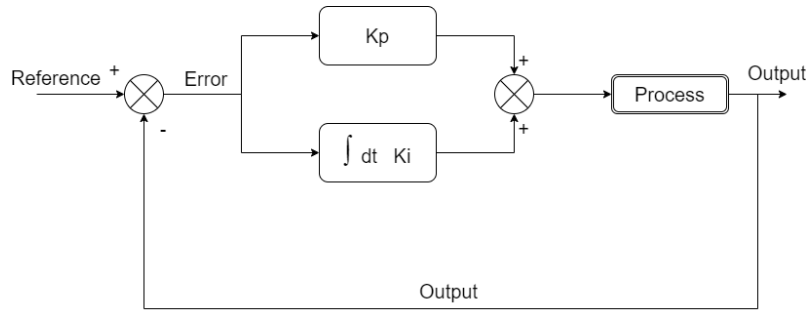


Figure 3.6: Functioning of a parallel PI controller.

In figure 3.6 it is possible to see the subtraction mentioned earlier, where the reference is the speed defined by the locomotive driver, limited by the maximum speed of the motor and the error being the difference between the current motor speed and the reference value. The resulting value is then passed through each of the controller segments, the P segment representing the proportional controller and the I segment that is the integral controller. The PI controller shown is a parallel PI controller, but there is also a series PI controller, where instead of the Proportional and the Integral parts receiving the same input, the output of the Proportional controller is used as input of the Integral controller.

3.4 Current Control

The IM is fed by the three-phase currents I_a , I_b and I_c and through the Clarke and Park transformations, these currents are converted into their I_d and I_q counterparts, with a DC-like waveform compatible with the usage of PI controllers. Upon being converted, they will be fed into a set of PI controllers.

In these controllers, they will be subtracted from the references, to create a new current command.

Although there are 2 PI controllers being used, one for each current, their gains are the same. That cannot be said for the PI controller used for the torque inducing current, since its gains are larger than the ones found inside the current controller. This current command is then passed through a modulation block such as Space Vector PWM, which outputs to the IGBTs.

As mentioned, the PI controllers compare the output of the system with the reference imposed in order to adjust the values dynamically to match the reference. These controllers are Proportional Integral (PI) controllers and represent two distinct control processes.

3.4.1 Proportional

Proportional control focuses on multiplying the error with the gain of the controller. The error as explained beforehand can be calculated as follows:

$$err = ref - out \quad (3.5)$$

And the Proportional block output is obtained by equation 3.6.

$$out = err * KP \quad (3.6)$$

Where KP is the Proportional block's gain. This block, as it stands with no other modules, is not very effective in achieving the reference value, since as the error decreases, it will stabilize without reaching the reference, as shown by figure 3.7.

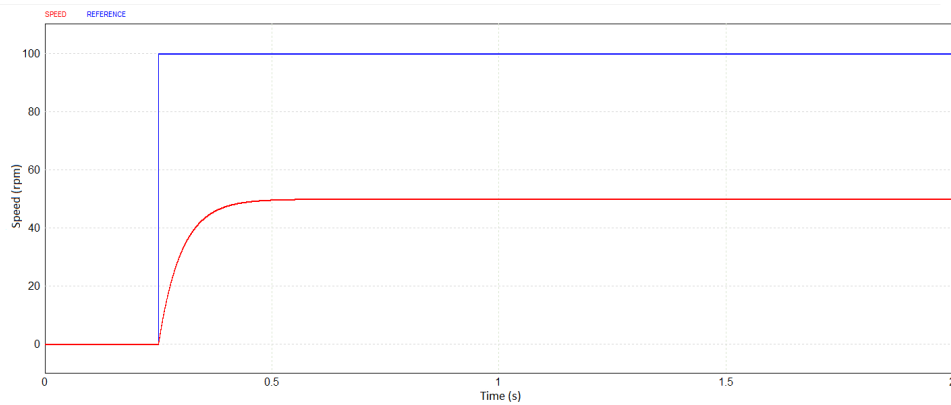


Figure 3.7: Curve of a P Block.

3.4.2 Integral

Integral control adds the capability to continuously reduce the error by comparing it with the error of the previous iteration and integrating this difference over time. By making this comparison, the controller is able to adjust the response based on the accumulation of error in steady-state and eliminate it. The calculations for the integral controller can be done in several ways, but it will only be discussed the Midpoint method [23], and it is obtained as follows.

$$y_{n+1} = y_n + hf\left(t_n + \frac{h}{2}, \frac{1}{2}(y_n + y_{n+1})\right) \quad (3.7)$$

Applying this formula and its variables to this specific controller, equation 3.8 is obtained

$$out_n = out_{n-1} + \frac{1}{2} * KI * T * (err_n + err_{n-1}) \quad (3.8)$$

Where T is the sampling period of the controller. The output of this controller can be shown in figure 3.8

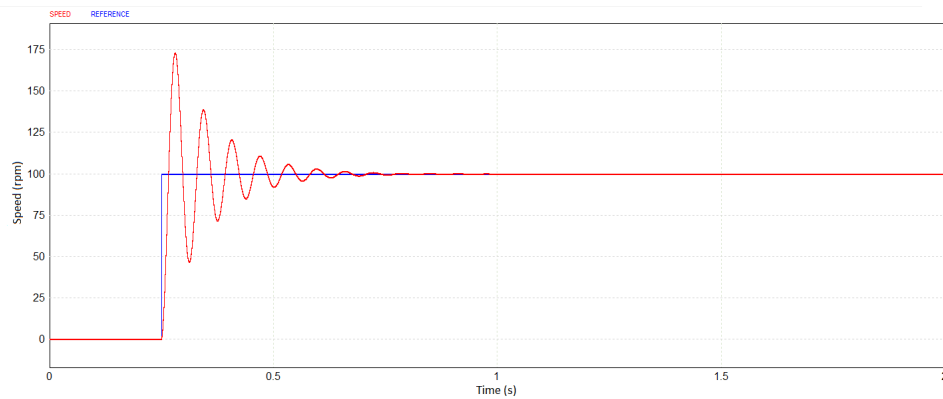


Figure 3.8: Curve of an I Block.

As shown in the picture, the error is continuously reduced over time, stabilizing at the reference value where the Integrator block will stop outputting.

3.4.3 Proportional Integral

The two controllers mentioned are often then used as a single unit, working in tandem to eliminate any steady state error that exists. The values of the KP and KI are adjusted in order to be able to correctly follow the reference given. By adjusting these values, there are some aspects that can be expected to occur. Upon reaching values that can provide enough stability to the system, these tweaks can then maximize the efficiency of the system.

Table 3.2 displays the effects of an increase in the KP and KI gains.

Table 3.2: Effects of an increase in the KP and KI gains.

PI Gain	Overshoot Percentage	Settling Time	Steady-State Error
KP	Increased	Small Impact	Decreases
KI	Increases	Increases	No Steady-State Error

Figure 3.9 illustrates the curve of a P block with an increased K_P in relation to the graph shown in figure 3.7

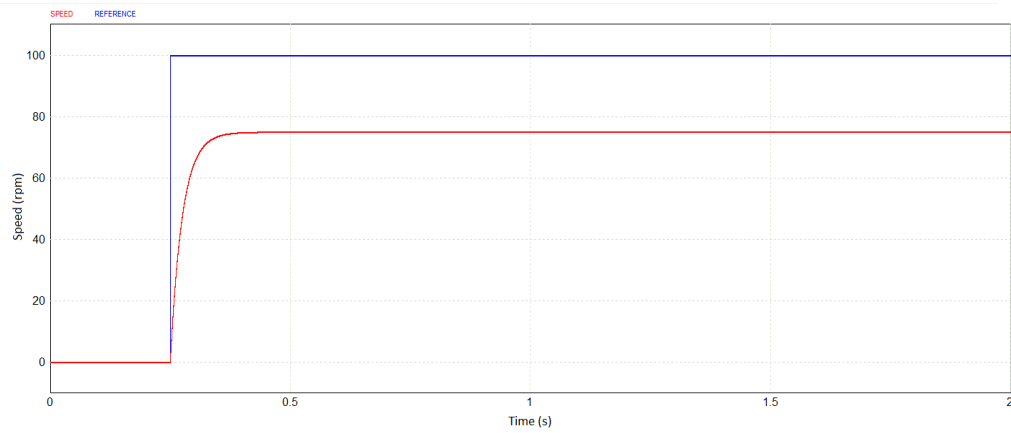


Figure 3.9: Curve of P Block from 3.7 but with an increased K_P .

The curve for the I Block shown in figure 3.8 was already with a large K_I to demonstrate the rippling effect and the settling time to reach the reference, so figure 3.10 will be with a smaller value of K_I to show the effects opposite to the ones displayed in table 4.2.

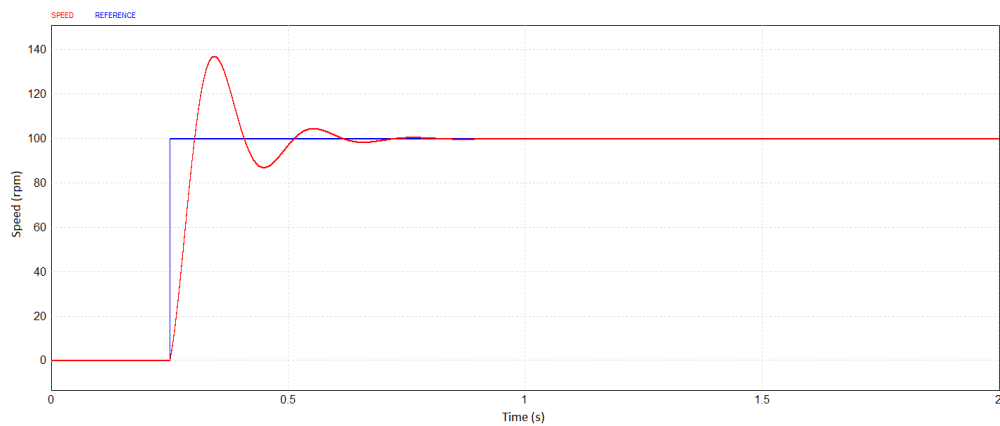


Figure 3.10: Curve of I Block from 3.8 but with a decreased K_I .

When the values of both the K_P and K_I are calibrated to the system, the curve follows the reference much closer and with less overshoot, as shown in 3.11.

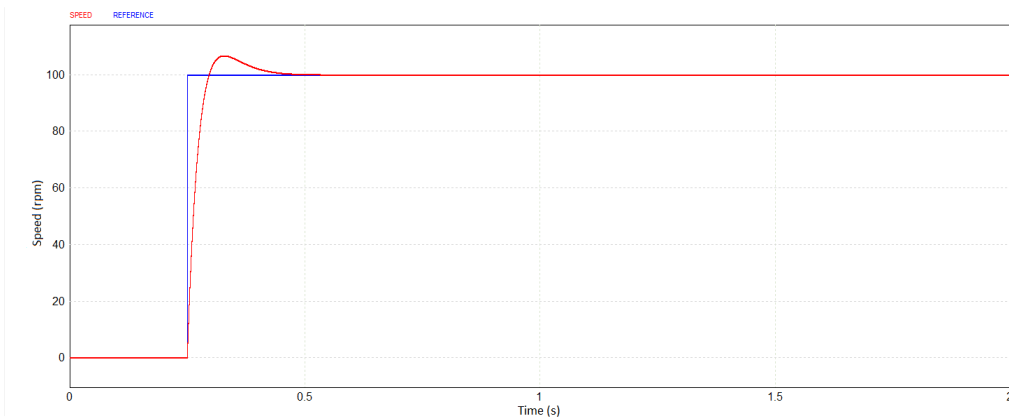


Figure 3.11: Curve of a PI block with calibrated gains.

3.5 Filtering

One thing to keep in mind that is often unaccounted for in the simulation environment is the presence of noise in the different measurements taken in a closed loop system. These errors lead to instability in the system, since it is propagated throughout the system, which then accumulates and may lead to failures. For that reason, the use of filters is required in order to eliminate these uncertainties.

There are four common types of filters: low-pass, high-pass, band-pass and band-stop filters. The first type focuses on attenuating high frequencies. When the frequency surpasses a predefined value for the cut-off, the filter's gain will decrease in order to lower the amplitude of the signal.

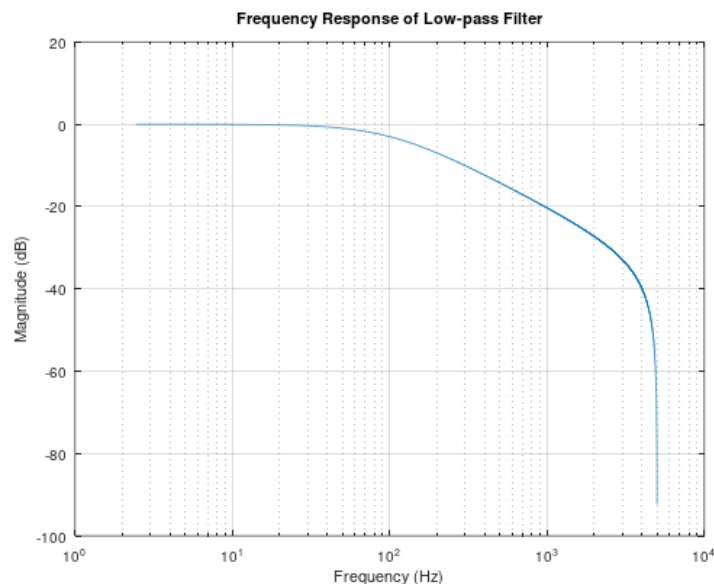


Figure 3.12: Magnitude to frequency graph for a first order low pass filter.

The second filter, functions in an opposite way as to the low-pass. This is because it will focus on

reducing the amplitude of the signal when encountering interference with low frequency. Facing this, the filter will reduce the amplitude of the signal until reaching and surpassing a predefined cut-off frequency.

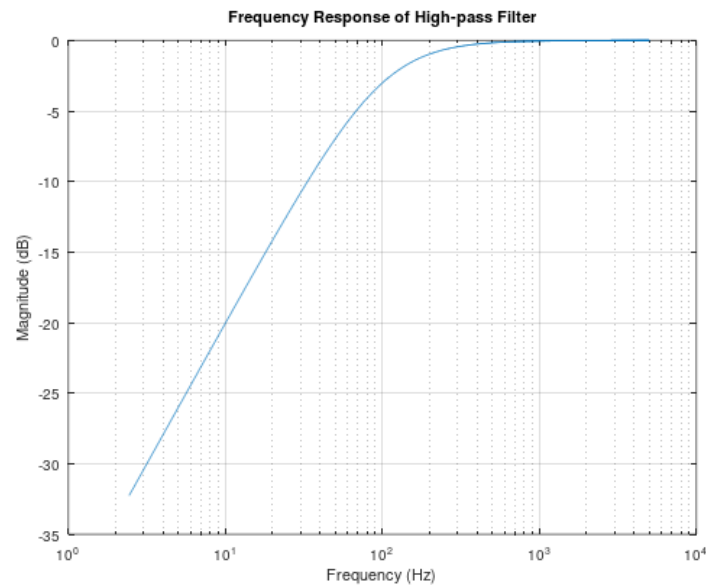


Figure 3.13: Magnitude to frequency graph for a first order high pass filter.

Thirdly, the band-pass filter is a mixture of the first two filters. This one attenuates any frequency outside a range of two cut-off frequencies, filtering any high frequency changes as well as low frequency changes.

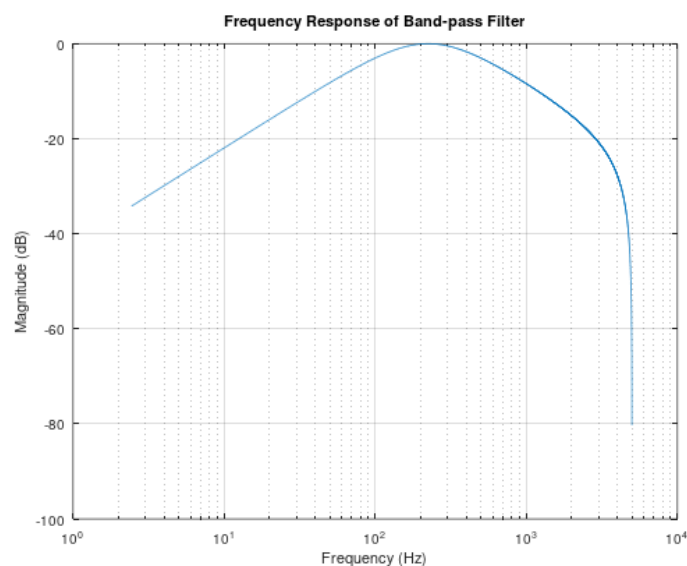


Figure 3.14: Magnitude to frequency graph for a first order band pass filter.

Finally, the band-stop filter is the opposite of the band-pass filter. This one attenuates any frequency inside a range of two cut-off frequencies, filtering, like the band-pass filter, any high

frequency changes as well as low frequency changes.

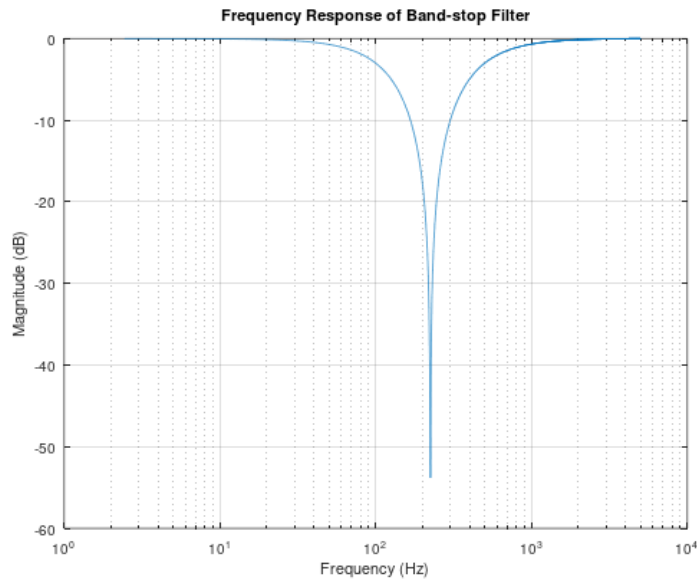


Figure 3.15: Magnitude to frequency graph for a first order band pass filter.

In this project, accurate speed measurement is essential as it is prone to high-frequency noise, which can significantly impact the performance of the control system. To address this issue, it is employed a low-pass filter to mitigate the noise and ensure precise speed readings. Specifically, the Infinite Impulse Response (IIR) filter is used. The main advantage of using an IIR filter in this context is its ability to effectively suppress high-frequency noise while preserving the integrity of the speed signal. By implementing the IIR filter, it is ensured that the speed reading is smooth and accurate, providing a reliable basis for the control algorithms to operate effectively. This approach enhances the overall stability and responsiveness of the motor control system, ensuring that it performs optimally under various operating conditions.

The Infinite Impulse Response (IIR) filter is a low-pass filter that operates by utilizing both current and past input and output values. It applies coefficients determined by the filter's order, sampling frequency, and cut-off frequency to achieve the desired noise attenuation.

The number of coefficients depends on the order of the filter. This means that the filter attenuates frequencies outside the passband more rapidly and the range of frequencies between the passband and stopband becomes narrower as the order increases. However, higher-order filters require more computation and memory compared to lower-order filters. This increased complexity can impact the implementation of the filter, especially in real-time applications such as motor control.

For this application, a first-order filter should suffice to filter any unwanted high frequency noise.

The coefficients needed to calculate the output of an IIR filter are the a and b coefficients. To calculate these, it is first needed the value of α .

$$\theta_c = 2\pi * \frac{f_c}{f_s} \quad (3.9)$$

$$K = \tan(\theta_C/2) \quad (3.10)$$

$$\alpha = 1 + K \quad (3.11)$$

Where f_C is the cut-off frequency and f_S is the Nyquist frequency, which is half the sampling frequency used.

After obtaining the value of α the coefficients can start being calculated. Since it is of order 1, only the values of index 0 and 1 of a and b will be calculated

$$a_0 = 1 \quad (3.12)$$

$$a_1 = \frac{-(1-K)}{\alpha} \quad (3.13)$$

$$b_0 = \frac{K}{\alpha} \quad (3.14)$$

$$b_1 = \frac{K}{\alpha} \quad (3.15)$$

After calculating the required coefficients, the output equation can be easily obtained through equation 3.16

$$y(n) = (b_0 * u(n) + b_1 * u(n-1)) - a_1 * y(n-1) \quad (3.16)$$

3.6 Pulse Width Modulation

As mentioned in the literature review of this document, in order to efficiently generate the correct duty cycles for the correct functioning of the IGBTs, modulation techniques have to be used. For this application, the contending methods are the Sinusoidal Pulse Width Modulation and the Space Vector modulation.

The Sinusoidal PWM technique is described by the comparison of the AC voltage reference with a high-frequency triangular wave in order to determine the correct switching states. This method is much simpler to implement than SVM, but yields less efficient results than the latter.

Space Vector Modulation on the other hand uses the $\alpha - \beta$ voltages to, through the usage of a space vector representation of the voltage vectors, determine which IGBT pair to turn on and the respective duty cycles.

3.6.1 Space Vector Modulation

In this project, SVM was the chosen method for the pulse width modulation.

As mentioned, this method uses the $\alpha - \beta$ voltages outputted by the current control loop to calculate the respective duty cycles of the IGBT pairs. These voltages are represented as vectors inside a unit circle divided in 6 parts, as shown in figure 3.16.

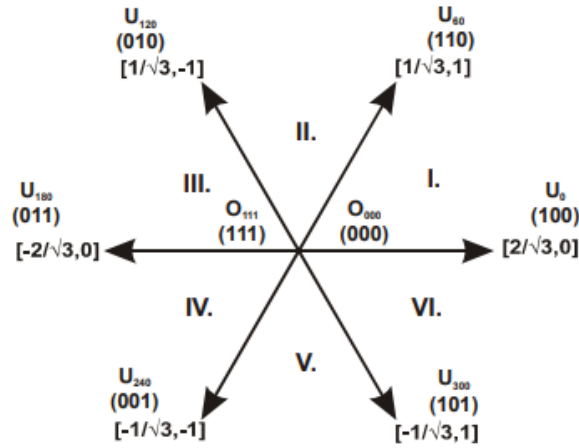


Figure 3.16: Basic space vectors of SVM retrieved from [24].

By finding the resulting voltage from the $\alpha - \beta$ components and its value of theta, the section of the hexagon in which it is located can be determined. Each vertex of the shape represents one or more switches to be turned on or off. This switching is what results in the actual functioning of the inverter, converting the DC current into AC current. During the functioning of this modulation technique, the duty cycles for the switches present in that section are obtained. But it is important to mention the existence of null vectors. Null vectors represent one of the following cases, when either all or none of the switches are turned on. After discovering the sector in which the vector is located, the respective projections are calculated to represent the voltage. These projections represent the duty cycles for each IGBT pair.

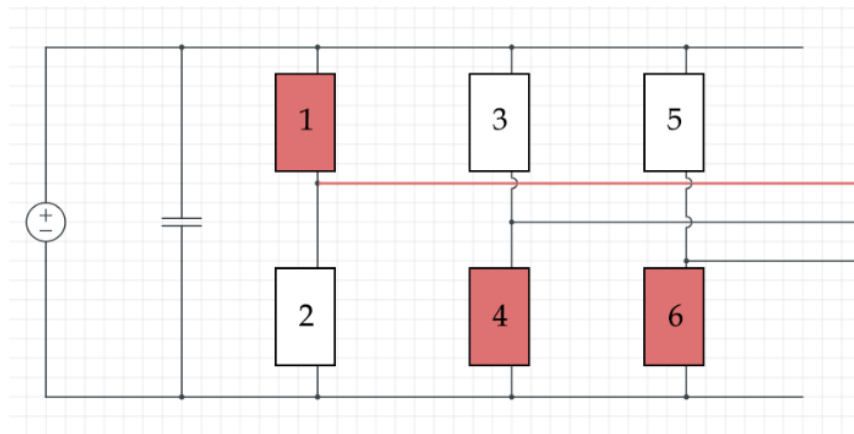


Figure 3.17: Visual representation of the IGBTs in the 1 0 0 configuration.

The calculation of the projections is mostly constant for all sectors except for sector 2 and 5, since these are centered around the beta axis. In the case of these two sectors, they must be divided into half-sectors, one on each side of the alpha axis, to correctly obtain the projections. Figure 3.18 displays the vector projections regarding sector 1.

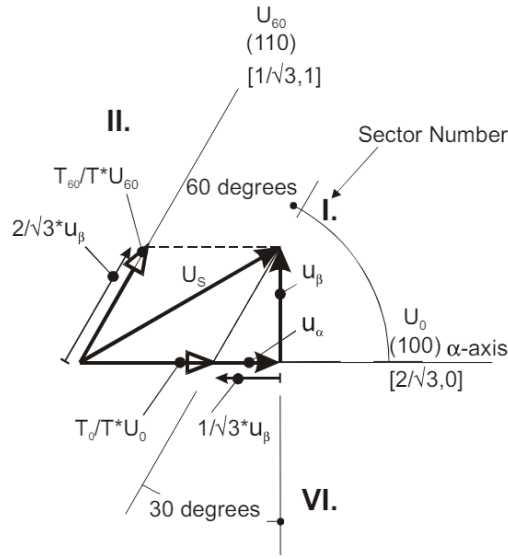


Figure 3.18: Vector projections regarding sector 1 retrieved from [24].

For sectors 1,3,4,6 the equations are similar to the equations present in the 1st sector with some minor changes in the signs of the $\alpha - \beta$ voltages. The equations in radians are as follows:

$$T = T_{\frac{\pi}{3}} + T_0 + T_{null} \quad (3.17)$$

$$U_S = \frac{T_{\frac{\pi}{3}}}{T} \cdot U_{\frac{\pi}{3}} + \frac{T_0}{T} \cdot U_0 \quad (3.18)$$

Where $T_{\pi/3}$ and T_0 represent the duty-cycle ratios determining the application of basic space vectors $U_{\pi/3}$ and U_0 within time period T . T_{null} denotes the duration for which the null vectors O000 and O111 are applied. These duty-cycle ratios are computed using the following equations [24]:

$$u_\beta = \frac{T_{\frac{\pi}{3}}}{T} \cdot |U_{\frac{\pi}{3}}| \cdot \sin \frac{\pi}{3} \quad (3.19)$$

$$u_\alpha = \frac{T_0}{T} \cdot |U_0| + \frac{u_\beta}{\tan \frac{\pi}{3}} \quad (3.20)$$

With the normalized magnitudes of basic space vectors set at $|U_{\frac{\pi}{3}}| = |U_0| = \frac{2}{\sqrt{3}}$ and by substituting the trigonometric expressions $\sin(\frac{\pi}{3})$ and $\tan(\frac{\pi}{3})$ with their respective quantities of $\frac{2}{\sqrt{3}}$ and $\sqrt{3}$ [24].

$$\frac{T_{\frac{\pi}{3}}}{T} = u_\beta \quad (3.21)$$

$$\frac{T_0}{T} = \frac{1}{2} \cdot (\sqrt{3} \cdot u_\alpha - u_\beta) \quad (3.22)$$

For sectors 2 and 5, the calculations become more complicated as they are split in half by the beta axis. In sector 2, the reference stator voltage vector U_S is generated by the appropriate duty-cycle ratios of the basic switching states $U_{\frac{\pi}{3}}$ and $U_{\frac{2\pi}{3}}$ [24].

$$T = T_{\frac{2\pi}{3}} + T_{\frac{\pi}{3}} + T_{null} \quad (3.23)$$

$$U_S = \frac{T_{\frac{2\pi}{3}}}{T} \cdot U_{\frac{2\pi}{3}} + \frac{T_{\frac{\pi}{3}}}{T} \cdot U_{\frac{\pi}{3}} \quad (3.24)$$

The next resulting duty cycles originate from the auxiliary components "A" and "B" as it is shown in figure 3.19 [24].

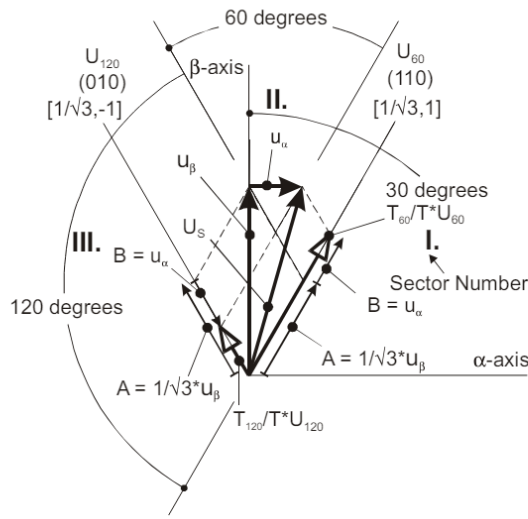


Figure 3.19: Vector projections regarding sector 2 retrieved from [24].

$$\frac{\sin \frac{\pi}{6}}{\sin \frac{2\pi}{3}} = \frac{A}{u_{\beta}} \quad (3.25)$$

$$\frac{\sin \frac{\pi}{3}}{\sin \frac{\pi}{3}} = \frac{B}{u_{\alpha}} \quad (3.26)$$

By restructuring their equations, the auxiliary time-duration components of "A" and "B" can be calculated. This is accomplished by substituting the trigonometric terms $\sin(\frac{\pi}{6})$, $\sin(\frac{\pi}{3})$, and $\tan(\frac{\pi}{6})$ with their numerical equivalents $\frac{1}{2}$, $\frac{\sqrt{3}}{2}$, and $\frac{1}{\sqrt{3}}$, respectively [24].

$$A = \frac{1}{\sqrt{3}} * u_{\beta} \quad (3.27)$$

$$B = u_{\alpha} \quad (3.28)$$

The auxiliary components can then be utilized to represent the resulting duty-cycles.

$$\frac{T_{\frac{2\pi}{3}}}{T} * |U_{\frac{2\pi}{3}}| = A - B \quad (3.29)$$

$$\frac{T_{\frac{\pi}{3}}}{T} * |U_{\frac{\pi}{3}}| = A + B \quad (3.30)$$

Utilizing these equations and considering the normalized magnitudes of basic space vectors as $|U_{\frac{\pi}{3}}| = |U_{\frac{2\pi}{3}}| = \frac{2}{\sqrt{3}}$, the equations describing the unknown duty-cycle ratios of basic space vectors $\frac{T_{\frac{\pi}{3}}}{T}$ and $\frac{T_{\frac{2\pi}{3}}}{T}$ can be expressed as follows [24]:

$$\frac{T_{\frac{2\pi}{3}}}{T} = \frac{1}{2} * (u_{\beta} - \sqrt{3} * u_{\alpha}) \quad (3.31)$$

$$\frac{T_{\frac{\pi}{3}}}{T} = \frac{1}{2} * (u_{\beta} + \sqrt{3} * u_{\alpha}) \quad (3.32)$$

After extensive research and analysis, the equations were simplified to a form that allows using the same equations for each case by merely switching the polarity [24].

$$X = u_{\beta} \quad (3.33)$$

$$Y = \frac{1}{2} * (u_{\beta} + \sqrt{3} * u_{\alpha}) \quad (3.34)$$

$$Z = \frac{1}{2} * (u_{\beta} - \sqrt{3} * u_{\alpha}) \quad (3.35)$$

With these equations, a table can be created that may easily obtain the correct projections for each sector [24].

Table 3.3: Expressions for t_1 and t_2 for all sectors adapted from [24].

Duty Cycle Ratios	$U_0, U_{\pi/3}$	$U_{\pi/3}, U_{2\pi/3}$	$U_{2\pi/3}, U_{\pi}$	$U_{\pi}, U_{4\pi/3}$	$U_{4\pi/3}, U_{5\pi/3}$	$U_{5\pi/3}, U_{2\pi}$
DC_1	X	Y	-Y	Z	-Z	-X
DC_2	-Z	Z	X	-X	-Y	Y

Finally, upon calculating the two projections, it is essential to convert them into the duty cycles to be fed into the IGBTs. Again, simplified equations were created to easily interchange for each sector to obtain the desired duty cycle value[24].

$$t_1 = \frac{T - DC_1 - DC_2}{2} \quad (3.36)$$

$$t_2 = t_1 + DC_1 \quad (3.37)$$

$$t_3 = t_2 + DC_2 \quad (3.38)$$

Table 3.4: Expressions for the duty cycles for the respective phases adapted from [24].

Motor Phase	$U_0, U_{\pi/3}$	$U_{\pi/3}, U_{2\pi/3}$	$U_{2\pi/3}, U_{\pi}$	$U_{\pi}, U_{4\pi/3}$	$U_{4\pi/3}, U_{5\pi/3}$	$U_{5\pi/3}, U_{2\pi}$
pwm_a	t_3	t_2	t_1	t_1	t_2	t_3
pwm_b	t_2	t_3	t_3	t_2	t_1	t_1
pwm_c	t_1	t_1	t_2	t_3	t_3	t_2

This results in the waveform in image 3.20 that is characteristic of Space Vector Modulation.

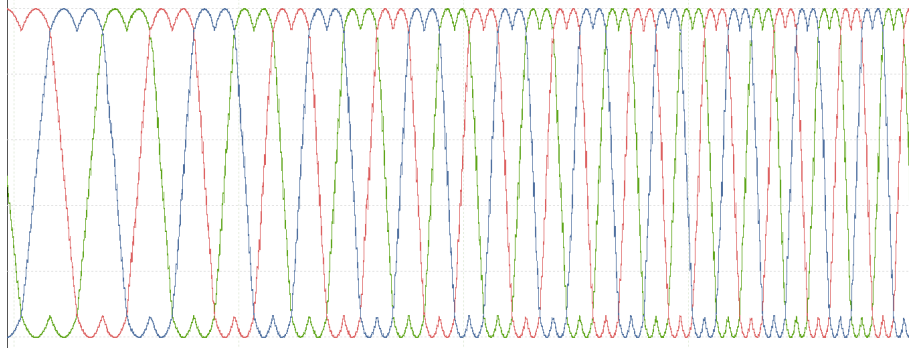


Figure 3.20: Space Vector Modulation 3-phase Waveform.

3.7 Torque Estimation

In this dissertation, the focus is to display a Traction Control Unit that utilizes a speed reference for its control. However, in most cases another method of control is used: torque control. Instead of defining a desired speed for the train to function, a torque value is specified and the locomotive motors must follow. This also requires that the torque value is estimated or measured. Due to the unavailability of measuring equipment for torque, the method for estimating this value will be demonstrated.

The torque created by an induction motor can be calculated through:

$$T = \frac{3}{2} * p * \phi_s * V_s \quad (3.39)$$

Where p are the pole pairs, ϕ is the stator flux and V_s the stator voltage. The value of stator flux and stator voltage can be calculated with the use of the $\alpha - \beta$ components that are outputted from the current controller into the Space Vector Modulation. The stator flux is obtained using the equation:

$$\phi = \int_0^t (V_s - I_s * R_s) dt \quad (3.40)$$

To obtain the stator flux, the stator voltage is required. Since there are no voltage probes to measure

this voltage directly, it can be derived by:

$$V_s = \sqrt{V_\alpha^2 + V_\beta^2} * V_{in} \quad (3.41)$$

Where V_α and V_β are the modulation indexes that are used for the Space Vector Modulation, and V_{in} is the input voltage.

After establishing the methods and theoretical foundations, the next step is to proceed to the simulation phase of this dissertation, where these concepts will be implemented in a simulated environment.

Chapter 4

Computational analysis

In this chapter, the results from the simulation work for this project will be demonstrated. With the aid of the simulation program PSIM it was possible to demonstrate the control loops in action and to later convert it into programmable code that could be exported to the physical test bench. Due to the unavailability of the motor parameters in the LE5600 locomotive, these simulations will be focused on the test bench present in the laboratory of Nomad Tech. Figure 4.1 displays the Simulation Environment used for this chapter.

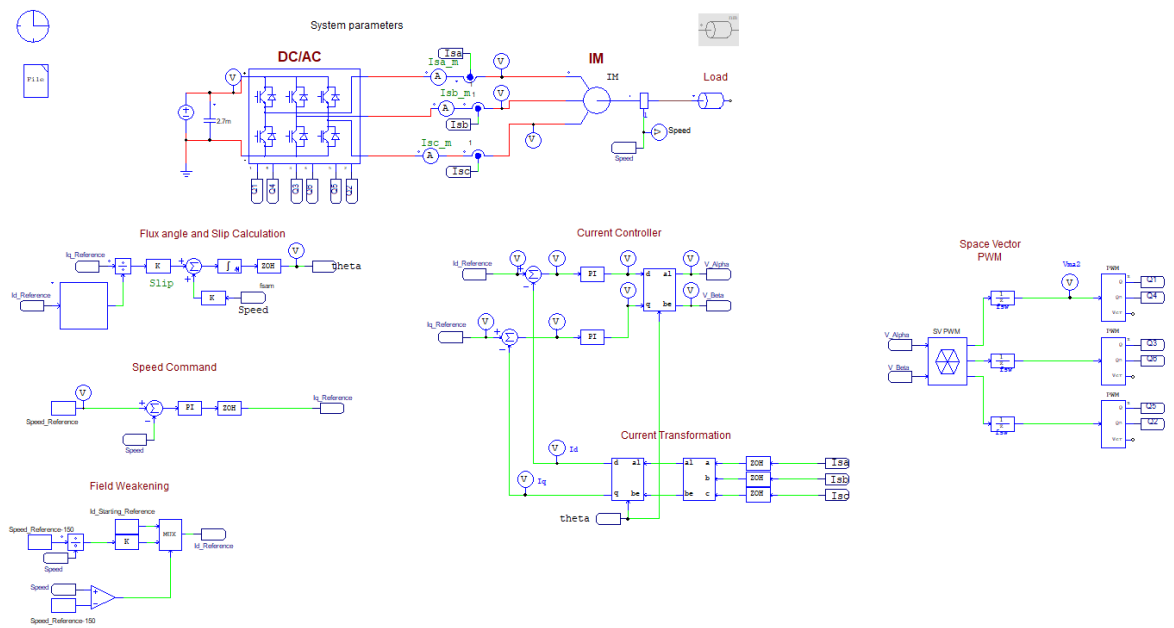


Figure 4.1: Simulation Environment used.

4.1 DC to 3-phase AC Inverter

To effectively control a 3-phase induction motor, precise control over the input current is essential. The power flow within a locomotive begins with the catenary line, which supplies single-phase AC current. In this simulation, a DC power source was employed, which is equivalent to an uncontrolled rectifier, converting the single-phase AC into DC.

The core component for converting this DC power, provided in this case from a 3-phase rectifier into AC for the motor, is the DC/AC inverter. This is accomplished by controlling the switching patterns of the IGBTs allowing for the modulation of the DC current into a sinusoidal AC waveform.

The principle behind this modulation is PWM. By varying the duration of the on and off states of the IGBTs, the inverter can control the voltage supplied to the motor. If the switching is executed with precise timing, the resulting waveform closely approximates a sinusoidal AC current.

For generating 3-phase AC current, the inverter must include three pairs of IGBTs, one pair for each phase. Each pair of IGBTs is responsible for producing one of the three phases of AC power. By coordinating the switching of these pairs, the inverter outputs the necessary 3-phase AC current required to drive the induction motor.

4.2 Motor parameters

The test bench provided by Nomad Tech is composed of a 3-phase induction motor, whose parameters were obtained by analyzing [25] and [26] in table 4.1.

Table 4.1: Induction Motor Parameters.

Parameters	Value
Stator Resistance, R_s (Ω)	0.01875
Stator Leakage, L_s (H)	0.0061
Rotor Resistance, R_r (Ω)	0.0267
Rotor Leakage, L_r (H)	0.0061
Magnetizing Leakage, L_m (H)	0.0157
Number of Poles, P	6
Moment of Inertia (kg.m^2)	0.002

Due to lack of equipment, these parameters were unfortunately not able to be confirmed through the parameter identification tests such as the previously mentioned no-load test and locked rotor tests.

4.3 Flux Angle and Slip Calculation

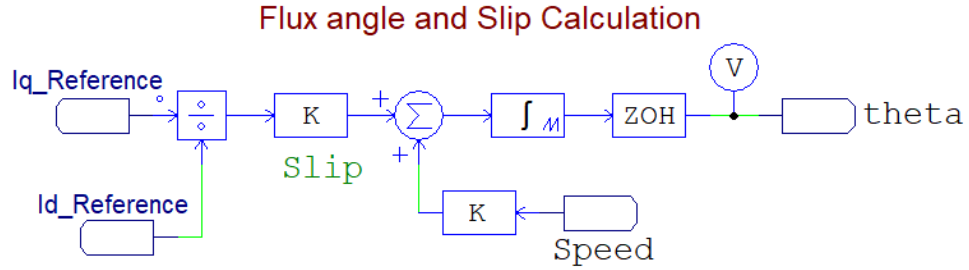


Figure 4.2: PSIM model for obtaining Theta.

As shown in figure 4.2, the calculations behind the value of theta revolve around the integration of the sum of the slip value and the calculation involving the value of speed.

The calculation of the slip value is as follows:

$$T_r = \frac{L_r}{R_r} = 0.2285 \quad (4.1)$$

$$w_{sl} = \left(\frac{1}{T_r} + p \right) \frac{I_{q_{ref}}}{I_{d_{ref}}} \quad (4.2)$$

$$w_{sl} = 7.377 \frac{I_{q_{ref}}}{I_{d_{ref}}} \quad (4.3)$$

The speed value must be converted to Radians per second in order to be used:

$$Speed = RPM \times \left(\frac{2\pi}{60} \right) \quad (4.4)$$

Finally, after adding these 2 values, they are inputted into an Integration block limited between $-\pi$ and π generating a curve like the one shown in figure 4.3.

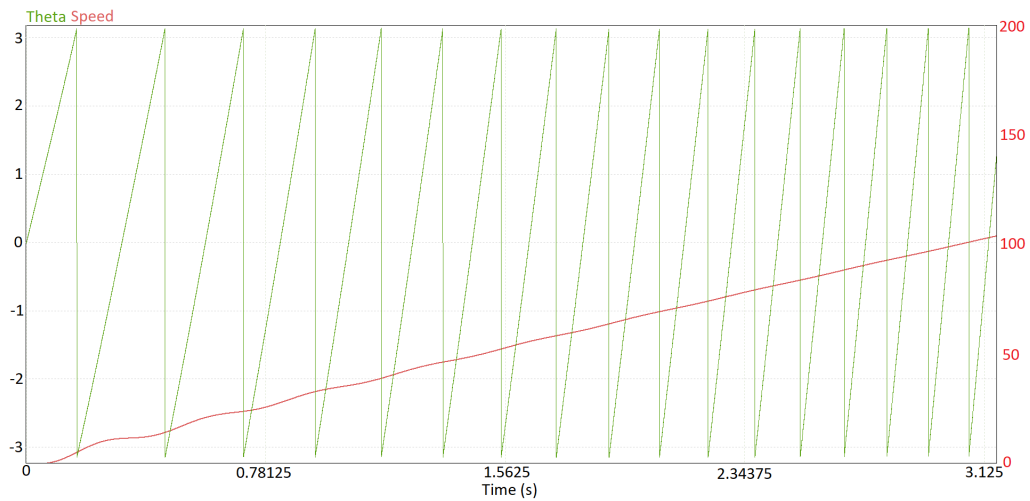


Figure 4.3: Theta curve in relation with the speed curve.

The theta value is fed into the rest of the control scheme, after a Zero-Order-Hold (ZOH) block in order to hold the sampled value until the next cycle. This is achieved by creating a delay to make it resemble more the real case scenario, and allows the theta to be on par with the sampling frequency of the rest of the circuit.

4.4 Current Controller

As explained beforehand, the current controller is the centerpiece of the motor controller, since it is what controls the currents in order for them to match the references generated by the other controllers. Figure 4.4 is the model of the Current Controller on PSIM.

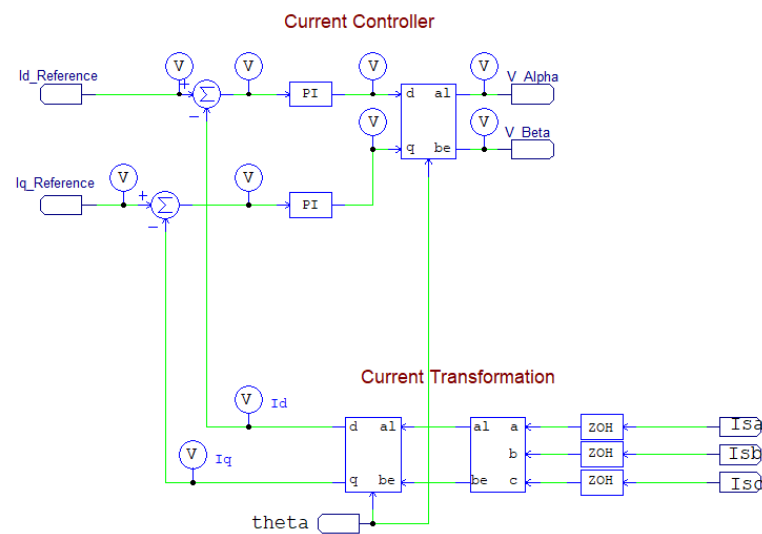


Figure 4.4: PSIM model for the Current Controller.

4.4.1 Current Transformations

Figure 4.4 displays all the transformations that occur inside the Current Controller. Inside this loop is where the Clarke and Park transformations, previously presented in Chapter 2, Section 2.3.4, can be found. After retrieving the current values, the Clarke Transform is applied to convert the 3-phase currents into their α - β counterparts. This is needed for the next transformation, the Park Transform, to convert them into the d - q rotational frame with the usage of the previously generated Theta. After these two transformations, the DC valued currents remain instead of AC, making them usable in the current controller inside the Proportional Integral blocks. In the output of the controller, it is also visible an Inverse Park Transform block. This is a needed transformation in order for the values exiting this controller to be usable in the modulation block that will be detailed in a later section.

4.4.2 Proportional Integral Blocks

Although it is possible to calculate the KP and KI gains for the motor used in this simulation using the motor parameters found in table 4.1, it was opted to determine these gains experimentally. The reason behind this is the fact that although the motor parameters provided seem to be close to the real motor parameters, unfortunately the real parameters are unknown. For this reason, even though the gains could be calculated, it was preferred to obtain them directly through experimentation since that is a step that will have to be repeated in the implementation phase.

The Current Controller PI block gains must address two potential problems: having a significantly lower time constant than the Speed Controller, and minimizing overshoot. The time constant must be much lower than the Speed Controller to avoid conflict between these two loops. In cases where the Current Controller has a high time constant, it can lead to both controllers integrating error between them instead of integrating error based on the motor behavior. If the Speed Controller demands higher values of I_q , but the Current Controller does not provide them quickly, it will cause the Speed Controller to request even more current. If this happens, it can cause an overshoot when the Current Controller catches up to the Speed Controller's requests. This behavior can also be mitigated by utilizing Anti-Windup, which limits the integration and prevents excessive oscillations. By reducing the Current Controller's time constant, that risk is eliminated by promptly meeting the current requests made by the Speed Controller, thus avoiding the double error integration issue.

In terms of overshoot, if it is too high or not corrected quickly enough, it can lead to issues such as improper magnetization or undesired motor behavior, and in worse situations, damage to the equipment.

After dealing with these concerns, the PI gains can start being calibrated. One way of achieving this is to start by increasing the value of KP, until it reaches a point of steady-state error, and the KI is then slowly increased to compensate for the rest of the error. But because it is needed to have a low Time Constant, calculated by:

$$TC = \frac{KP}{KI} \quad (4.5)$$

The KI value must at least be higher than the KP value in order to have a usable Time Constant, and after some experimentation, the values for KP and KI were defined as:

Table 4.2: Current Controller KP and KI gains.

PI Gain	Value
KP	0.012890
KI	0.2555
TC	0.05045

4.4.3 Controller Performance

This subsection will be dedicated to demonstrating the performance of the current controller.

4.4.3.1 Constant Reference

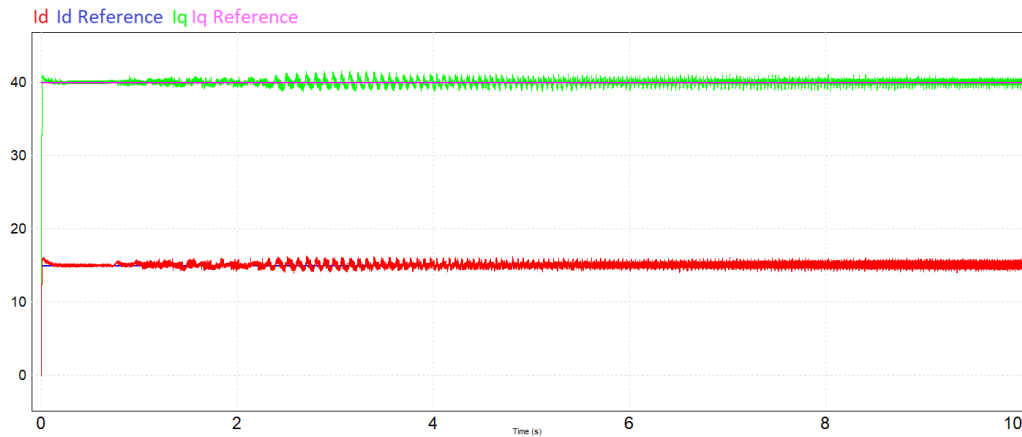


Figure 4.5: d-q Currents Following Reference.

From the graph in Figure 4.5, it is evident that the d-q currents follow their respective references with minimal deviation. The initial overshoot in both the I_d and I_q currents is minimal, which is the desired behavior of the current controller. The accurate tracking of I_d and I_q currents demonstrates the effectiveness of the tuned PI gains in maintaining the desired performance. This accurate control is important for ensuring that the motor operates efficiently and reliably under various load conditions.

4.4.3.2 Step response

One key test performed was the step response test. In this test, the controller was subjected to varying step changes in the I_q and I_d references to evaluate its dynamic response. The objective was

to observe how well the controller could follow these reference changes with minimal overshoot, undershoot, and a fast response time. This test helps in understanding the controller's behavior in handling sudden changes in current demands, which is necessary for maintaining the motor's performance and stability.

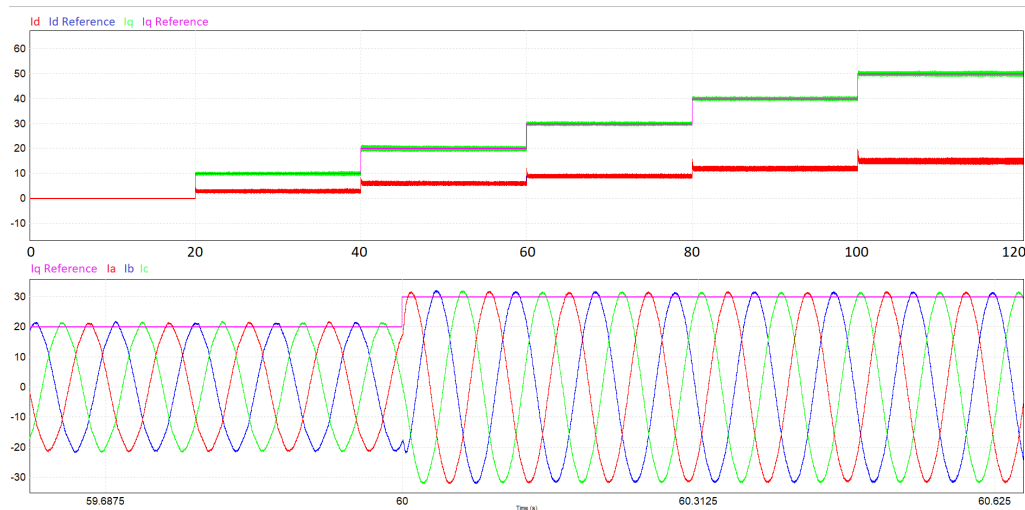


Figure 4.6: Step Response test.

The graph in Figure 4.6 illustrates the results of the step response test conducted in PSIM. As depicted, especially in the zoomed version showing the 3-phase currents, the third step response is instant and without deviation. The controller effectively follows the provided references with minimal overshoot and quick response time. These results indicate that the controller is well-tuned and capable of managing current changes efficiently.

Although the exact gains used in the simulation might need further adjustment during physical implementation, these simulation results provide a solid foundation and valuable insights into the controller's performance. The step response test allowed us to identify any potential issues and make necessary adjustments to the controller's parameters, ensuring a smoother transition to real-world testing.

By conducting this simulation, it was ensured that the current controller was not only stable and responsive, but also robust enough to handle even in unexpected operational conditions. This thorough simulation process was a significant step in the overall design and testing workflow, providing confidence in the controller's capability before moving to physical tests.

4.5 Speed Controller

To ensure the proper functioning of the current controller, it is essential to provide accurate references for I_q , the torque-inducing current. This is where the speed controller plays a crucial role.

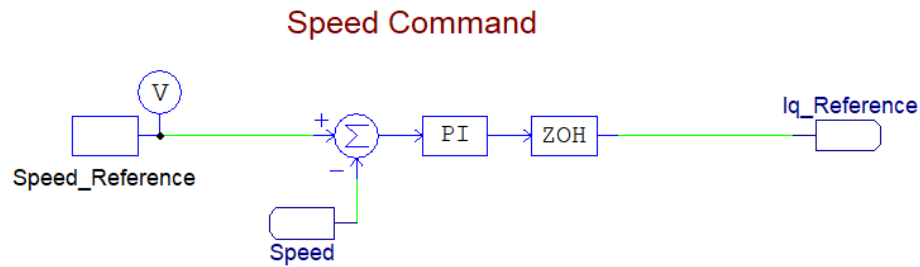


Figure 4.7: PSIM model for the Speed Controller.

The speed controller, depicted in Figure 4.7, is simpler than the current controller as it only requires a single PI block instead of two. However, despite its simplicity, the speed controller's tuning is critical. Unlike the current controller, which needs to respond accurately to references, the speed controller must generate smooth and reasonable I_q references. Poor adjustment of this controller can result in incorrect I_q references, leading to unintended motor behavior or even damage to the equipment.

The tuning of the speed controller's gains, K_p and K_i , was determined through experimentation, similar to the process for the current controller. However, the key difference lies in the time constant. As mentioned earlier, the time constant for the speed controller must be higher than that of the current controller to avoid conflicting behavior. This higher time constant, which results in a slower response, is beneficial in applications such as locomotive control. A slower controller results in gentler starts and smoother speed transitions, enhancing passenger comfort. If the controller responded too quickly, it could cause abrupt starts and excessive torque, leading to an uncomfortable experience.

4.5.1 Speed Controller Performance

This subsection will be dedicated to demonstrating the performance of the speed controller.

4.5.1.1 Fixed Reference

After tuning the PI gains, the speed curve obtained with a fixed reference of 500 RPM is shown in figure 4.8.

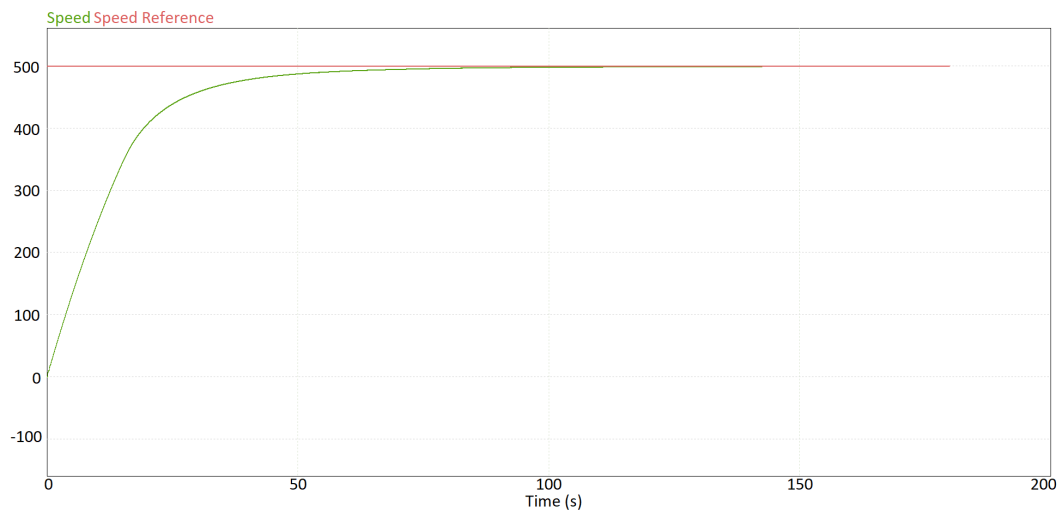


Figure 4.8: Speed Curve with 500 rpm reference.

As illustrated in the graph of Figure 4.8, the speed curve is very smooth with no overshoot. The motor begins to reduce its acceleration as it approaches the reference speed, ensuring a gentle transition to the desired speed. This behavior validates the effectiveness of the speed controller in maintaining a stable and comfortable operation.

4.5.1.2 Step Response

This test involved subjecting the speed controller to varying step changes in the speed reference. The primary objective was to evaluate the controller's dynamic response to sudden changes in speed demand. The main focus is to see how well the controller can follow these reference changes with minimal overshoot, undershoot, and to monitor the response time. This test is crucial for understanding the controller's ability to manage rapid changes in speed, which is needed for maintaining the overall stability and performance of the motor.

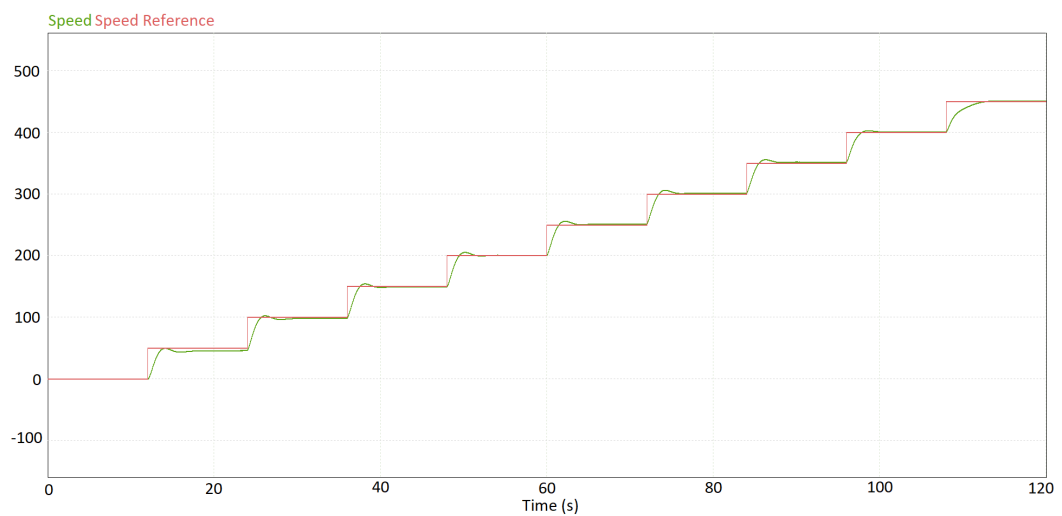


Figure 4.9: Step Response test.

The graph in Figure 4.9 demonstrates the results of the step response test conducted in PSIM. As shown, the speed controller accurately follows the reference speed changes with minimal overshoot and a prompt response time. These results confirm that the controller is well-tuned and capable of handling sudden speed changes effectively.

4.5.1.3 Steady-State Test

This test measured the steady-state error to ensure the controller maintains the desired speed with minimal deviation under constant conditions. By running the motor at a fixed speed for an extended period, the accuracy and stability of the speed control was noticeable. The steady-state test helps in verifying that the speed controller can hold a consistent speed without significant drift or fluctuation, ensuring reliable long-term operation.

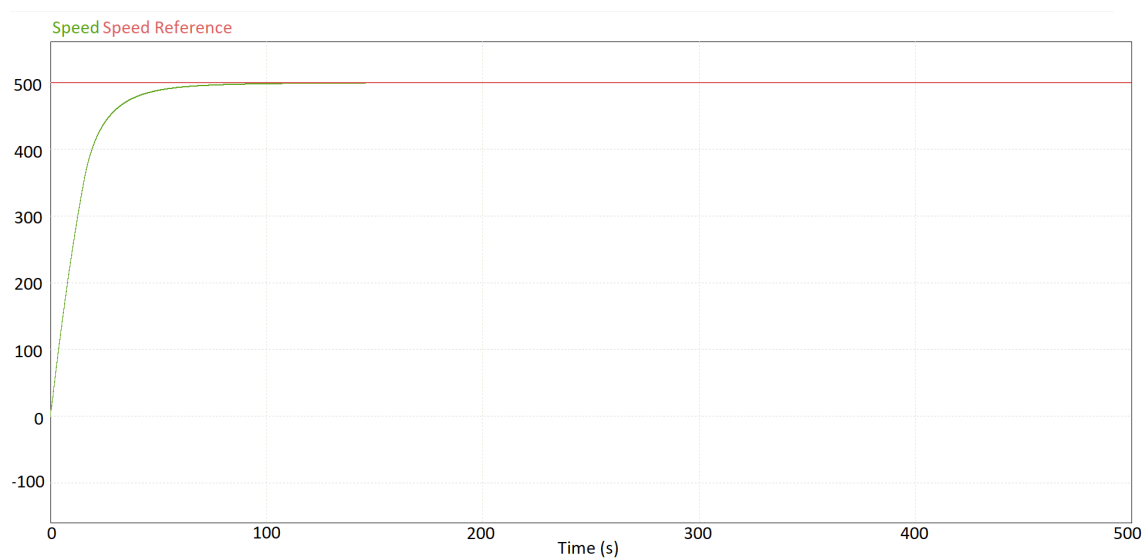


Figure 4.10: Steady-State Test.

The graph in Figure 4.10 illustrates the speed controller's performance during the steady-state test. The controller maintains the desired speed accurately with no noticeable deviation over time. The smooth acceleration and absence of overshoot indicate that the PI gains are well-adjusted, providing stable and precise speed control.

By conducting these tests in the simulation environment, it was ensured that the speed controller was both responsive and stable. The results from the step response test and the steady-state test provided confidence in the controller's ability to handle various operating conditions effectively. These thorough simulation tests were a critical part of the design and testing workflow, laying the groundwork for successful physical implementation.

4.6 Field Weakening

In this experimentation, it was found that field weakening provided significant benefits even at speeds lower than the rated speed, improving overall control.

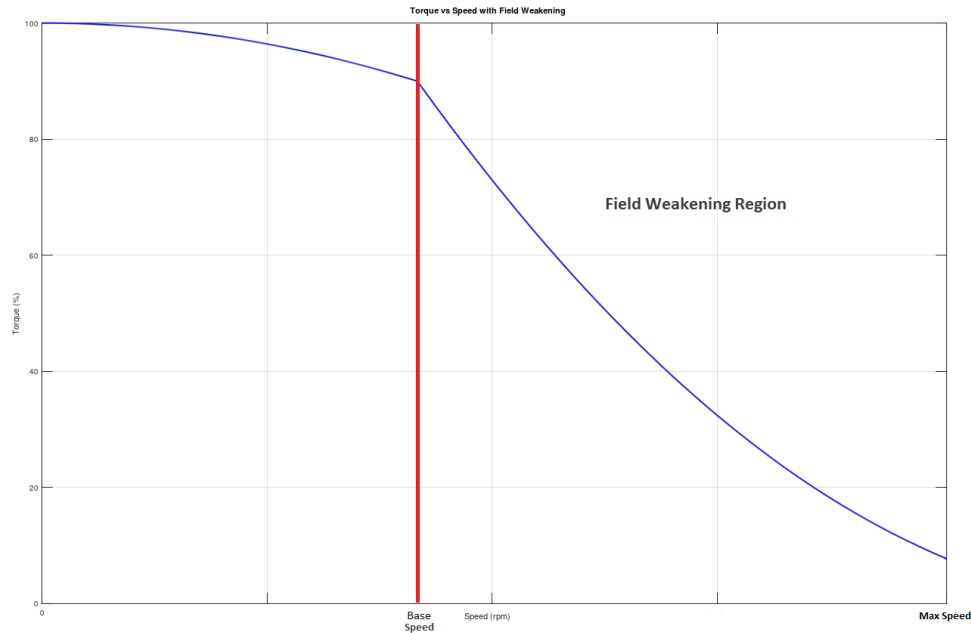


Figure 4.11: Torque versus Speed with Field Weakening.

In induction motors that are used in high-speed applications, like in this case, the objective is to maximize torque and fully utilize the machine and converter capabilities. However, at high speeds, the back electromotive force (EMF) restricts the flow of stator currents, resulting in a decrease in torque. To achieve higher speeds, it is essential to reduce the flux level [27]. In the case of the simulation environment the back EMF was not restricting the flow of the stator currents; however, it was found that by enabling Field Weakening at a lower speed than intended, the torque reduction created by it would be beneficial. In these tests, Field Weakening was set to begin around the 350 rpm mark. This early application of Field Weakening helped achieve a smoother acceleration curve with less overshoot, which is important in applications such as locomotives, where smooth and gradual acceleration is preferred for passenger comfort and equipment longevity.

The field weakening technique was tuned to ensure optimal performance and stability. The initial magnetizing current I_d was gradually reduced as the speed increased past the predefined set point of 350 rpm. This gradual reduction was essential to prevent instability and maintain control over the motor currents.

The equation used for the reference I_d at speeds above 350 rpm was:

$$I_{dRef} = I_{dBaseRef} \times \frac{Speed_{Ref}}{Speed} \quad (4.6)$$

This formula allowed us to adjust the I_d reference smoothly and effectively, avoiding the rapid demagnetization that could lead to instability and in a real life scenario possibly motor damage.

4.7 Code Implementation

To facilitate the transition from simulation to real-world application, PSIM's coding capabilities were utilized by beginning the implementation process within the simulation environment. This approach was necessary as it enabled us to develop and test the code thoroughly without risking damage to physical components.

Initially, the focus was on converting logical and functional blocks into code. This process involved replacing most elements except for essential components like wires and basic blocks, such as proportional, multiplication, and division blocks. Due to the modular nature of these functions, this step was relatively straightforward and allowed us to reuse code efficiently.

Next, each controller was compartmentalized by creating dedicated blocks for Theta estimation, Current Control, Speed Control, and Field Weakening. The coding for the Space Vector Block, however, would be addressed later during the physical implementation phase.

The final step involved consolidating all the controllers into a single comprehensive C block. This step posed a significant challenge due to the need to determine the order of execution for each controller. After analyzing the flow of data and the sequence of operations, it was established the execution order as depicted in Figure 4.12.

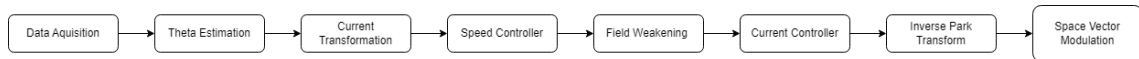


Figure 4.12: Controller functions execution sequence.

Figure 4.12 outlines the critical steps in this control sequence. Initially, all necessary values are gathered, including the three-phase currents (I_a , I_b , and I_c) with the use of current probes and the motor's rotational speed employing an encoder. Using these values, other essential parameters of the motor can be estimated.

The next step involves calculating the Theta angle, which is crucial for the Clarke and Park transforms transformation process. The Theta value is derived from the motor's speed data, along with other constant or referenced values. Once the Theta value is determined, the Current Transformation stage begins, converting the three-phase currents into the rotational reference frame components, I_d and I_q , using the Park and Clarke transformations.

With the I_d and I_q values obtained, their respective reference values are then generated. These references are compared within the Speed Controller and the Field Weakening block, and the results of these comparisons are processed by the Current Controller before being fed into the Space Vector Modulation.

To bring the simulation environment closer to the real life scenario, it was implemented a square-wave generator to simulate the microprocessor control loop interruptions, operating at a

frequency of 10 kHz. This step was important for synchronizing the controller operations and ensuring accurate simulation results.

Upon completion and verification of the code, a fully functional simulation environment was achieved, as illustrated in Figure 4.13.

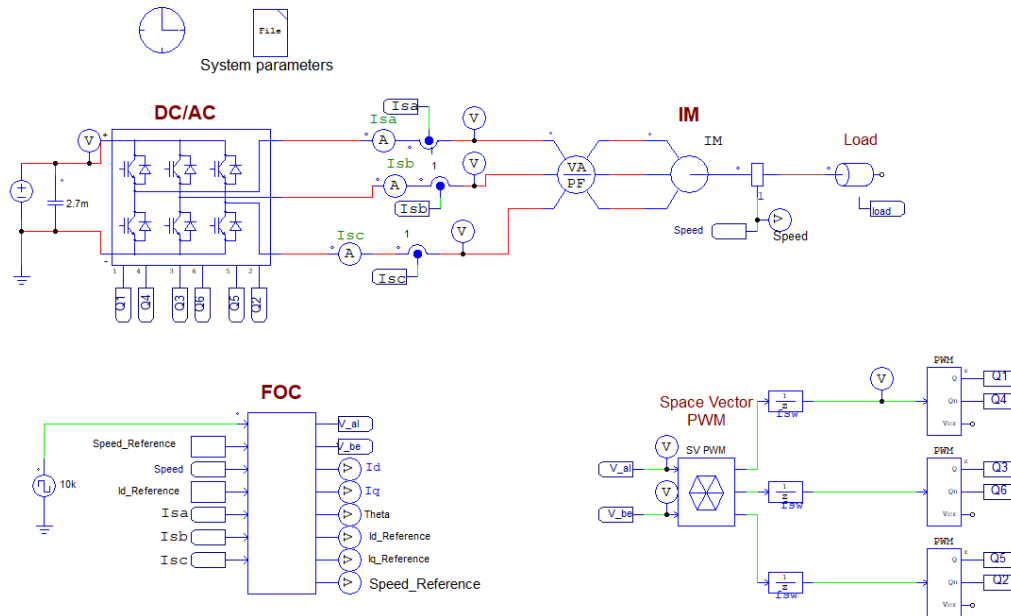


Figure 4.13: Final Simulation Environment.

This simulation environment served as a robust platform for further testing and refining the control algorithms, providing a solid foundation for subsequent physical implementation. Through this approach, potential issues were able to be minimized and ensure a smooth transition from simulation to real-world application.

Chapter 5

Controller Implementation

As mentioned in the previous chapter, the implementation of a traction control unit on the test bench provided by Nomad Tech will be demonstrated.

5.1 The Test Bench

The physical setup is very similar to the one presented in the simulations in figure 4.1. For a source, it was used a variac that could provide a wide range of voltages from 0 up to 400V, which was ideal for the variety of studies performed. This rectifier DC source is then connected to the 3-phase inverter, connected by a DC capacitor bank that is short-circuited to the source. Connected to the IGBTs is a Driver that receives the PWM signal generated by the main control unit. At the output of the IGBT pairs are the 3-phase power cables connected to the three phases of the motor, completing the circuit.

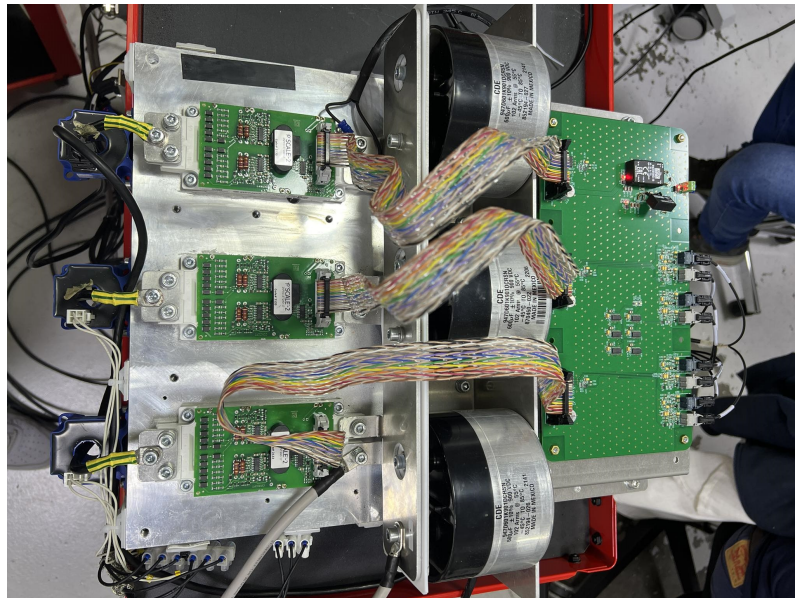


Figure 5.1: 3-phase inverter physical setup.

The motor is a 3-phase motor with a nominal voltage of 380V and a rated speed of 1000 RPM with a power factor of 0.87. This motor is coupled to a DC motor whose purpose is to generate torque to simulate the behavior of a train during different stages.



Figure 5.2: 3-phase motor from the test bench.

5.2 Generating a PWM signal

In the previous chapter, it was explained that the C Block supported the preparation for the physical implementation by coding the control scheme within PSIM. This preliminary step significantly accelerated the process of translating the control algorithms into functional code, allowing for a rapid and relatively straightforward implementation with minimal modifications to the original structure.

The initial task was to generate a PWM signal. Since the Space Vector Modulation had already been successfully coded and tested, this part of the implementation was relatively effortless to transfer and validate. To verify the SVM operation, a fixed Theta value was generated that corresponded to a manually controlled frequency during the study. The SVM output was scaled to an amplitude value ranging from 0 to 1. This approach is similar to the control method known as Voltage over Frequency (V/F) control, which is a commonly used and straightforward method for implementing control sequences.

V/F control ensures that the voltage applied to the motor is proportional to the frequency, maintaining a constant ratio, which is essential for stable motor operation. This method is particularly advantageous for its simplicity and ease of implementation, making it a suitable choice for the initial phases of testing and validation.

The graph illustrated in figure 5.3 illustrates the 3-phase currents being input into the motor, demonstrating the effective generation and control of the PWM signals.

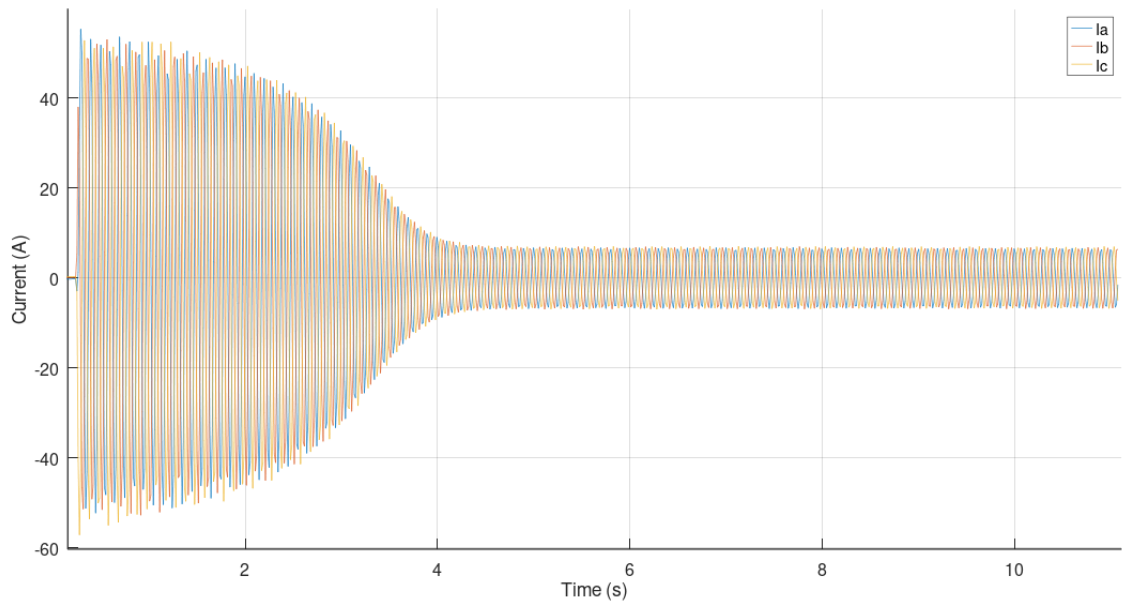


Figure 5.3: 3-phase currents of a 0.25 amplitude modulation index and 10 Hz frequency.

With this setup, it was possible to visibly observe the characteristic curve of Space Vector Modulation by measuring the voltage of the gate resistance of one of the 3 IGBTs.

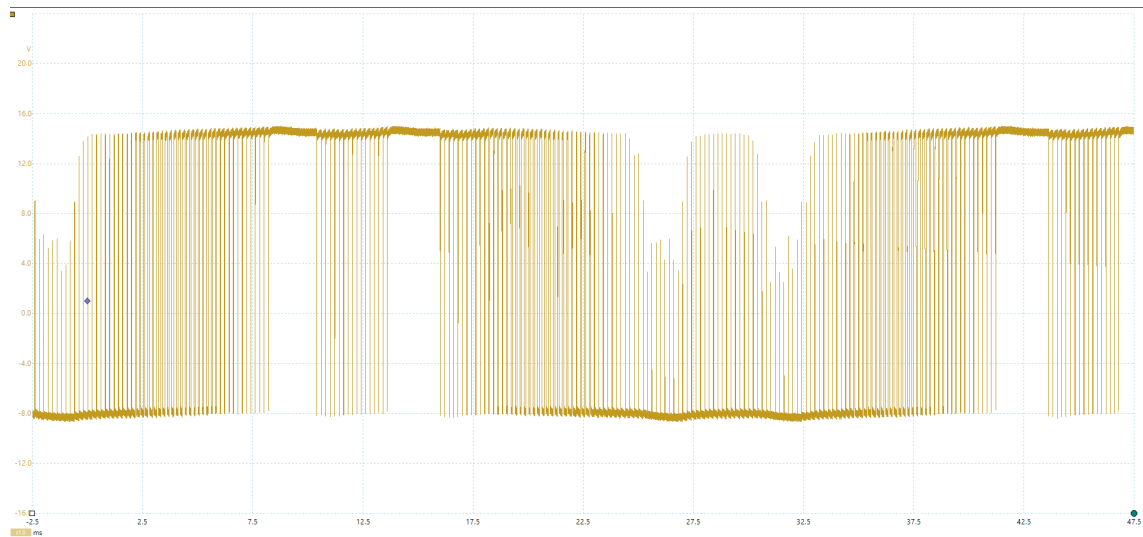


Figure 5.4: Space Vector Modulation Wave.

5.3 Theta Estimation

In order to accurately estimate the rotor position (Theta) and speed, the encoder that was already installed in the test bench was used. The encoder provides real-time readings of the motor's

rotational speed, which is essential for precise control. However, these readings were coupled with a significant amount of noise, which could adversely affect the control performance.

To mitigate this issue, it was implemented a 1st order low-pass filter to smooth out the noisy speed readings. This filtering process is crucial for ensuring that the control system receives stable and accurate speed information, thereby enhancing the overall performance and stability of the motor control.

The following image demonstrates the speed readings with and without the application of the low-pass filter, highlighting the improvement in signal clarity achieved through filtering.

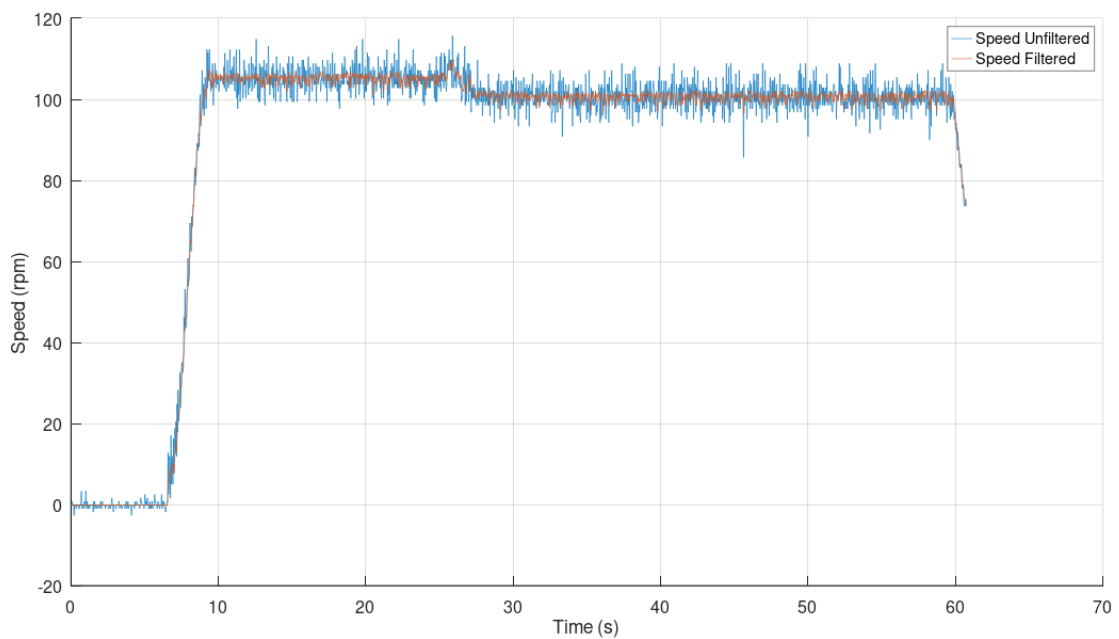


Figure 5.5: Speed reading with and without 1st order filter.

After applying the low-pass filter, the speed value becomes much more readable and suitable for an accurate estimation of Theta. To verify the correctness of the Theta estimation, it is needed to verify the alignment. It is important for the system's stability that the Theta angle is correctly aligned with one of the motor's input current phases. Proper alignment is achieved when the Theta value reaches zero simultaneously with one of the current phases also being zero.

The following image shows the Theta value aligned with one of the current phases, demonstrating the accuracy of the Theta estimation.

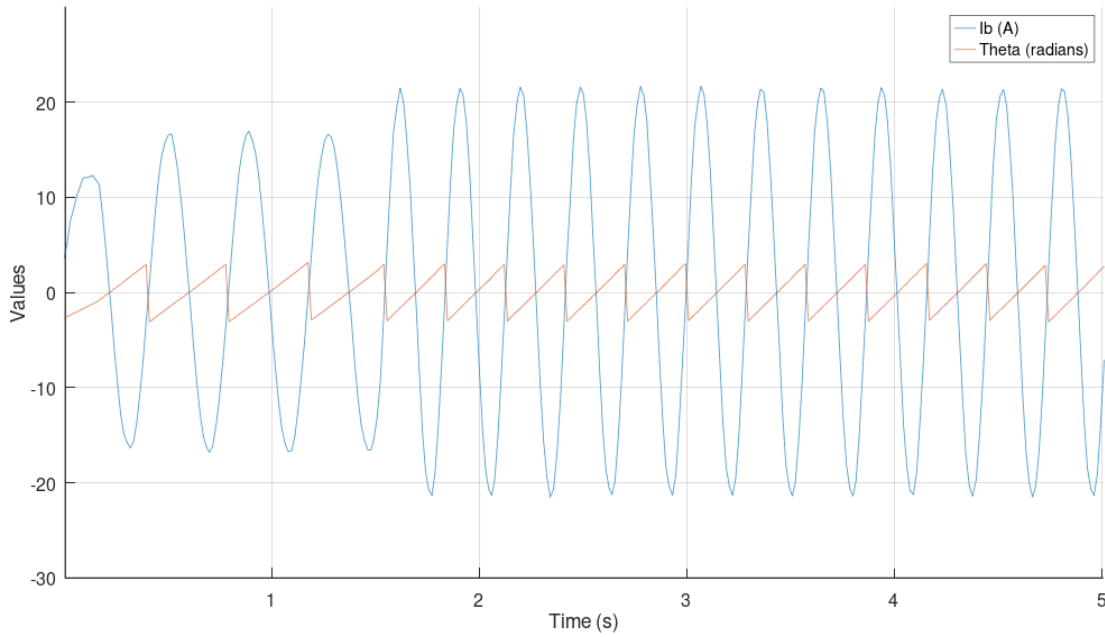


Figure 5.6: Current phase B aligned with Theta.

5.4 Clarke and Park transformations

Upon obtaining the correct Theta estimation, the Clarke and Park transformations can now be performed. By following the formulas already addressed in chapter 2, 3 and 4 it becomes possible to correctly obtain the α , β and d , q currents utilizing the Theta obtained earlier. These are the currents that are then used in the current controller in order to match them with their reference. One of the currents to take notice is the magnetizing current I_d . Unlike the torque inducing current reference, I_q reference, that is obtained through the speed controller, the I_d reference has to be estimated.

5.5 Current Controller

After obtaining the current transformations, the most important controller can be addressed: the current controller. This controller compares the current values with their references to correct and match them. Without proper fine-tuning, imbalances in the motor's functioning can occur, leading to inefficiency and instability.

Initial guidance for the gain values (K_p and K_i) was obtained from simulations, providing a solid starting point for adjusting the controller. However, despite being programmed to closely mimic real-life scenarios, these simulation values still required further adjustment. The gains were calibrated by analyzing the controller's response to fixed reference values.

One of the most important aspects of PI block calibration is its time constant. The time constant determines how quickly the integration of the PI block will accumulate error. A smaller time

constant results in faster error accumulation, thereby impacting the PI block's output more quickly. To avoid persistent errors between controllers, the time constant for the current controller must be significantly smaller than that of the speed controller. Ensuring a faster current controller than the speed controller allows for stable and safe motor control, reducing the likelihood of high current changes and deviations.

After properly regulating the gains, the current controller's response to given I_d and I_q references is as follows:

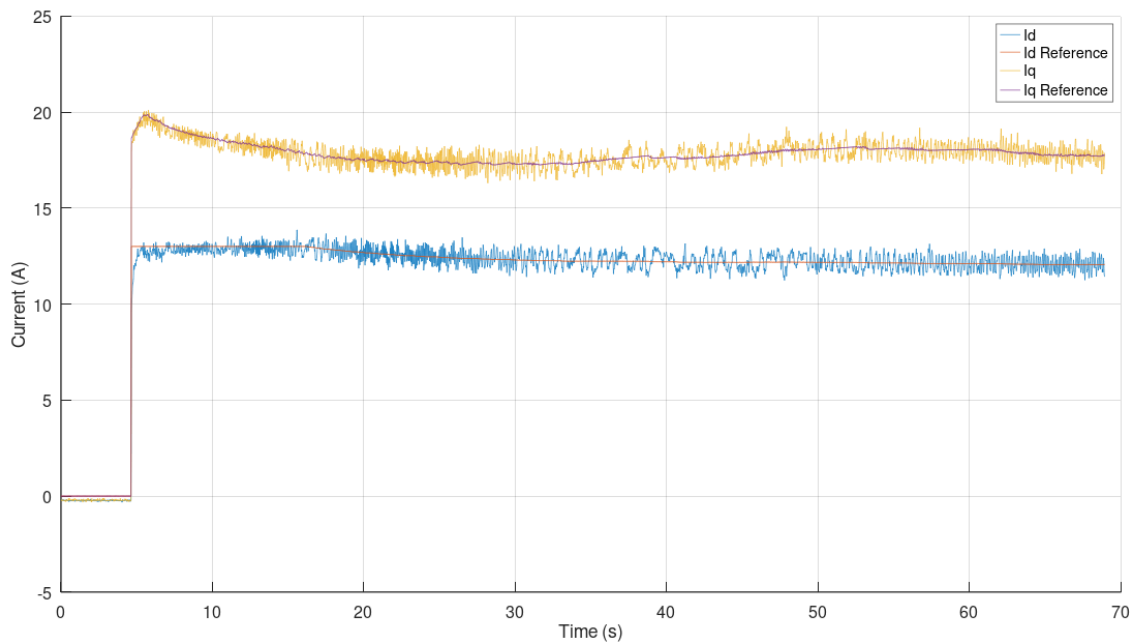


Figure 5.7: Current Control curve in the test setup.

5.5.1 Tests Performed and Implementation Notes

To verify the correctness and stability of the current controller, a step response test was conducted. This test aims to evaluate the dynamic response of the current controller to sudden changes in current reference values. By applying a step change to the I_d and I_q references, it is possible to observe the controller's response, measure the rise time, settling time, and any overshoot or undershoot in the current response.

The PI gains utilized can be found in table 5.1.

Table 5.1: Test bench Current Controller KP and KI gains.

PI Gain	Value
KP	0.013
KI	0.08
TC	0.1625

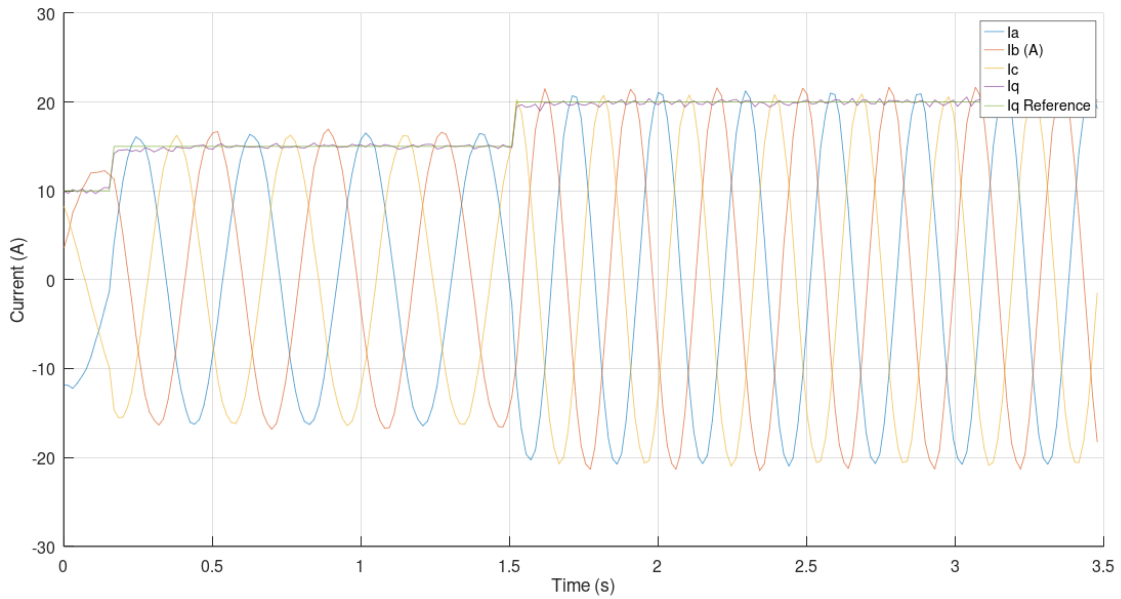


Figure 5.8: Step Response test for the Current Controller.

As it is shown in the graph in figure 5.8 the current controller, after correctly implementing the PI gains, has a very fast and precise response with very minimal to practically no overshoot whatsoever, which is very important to avoid damage to equipment due to current spikes and causing unwanted behavior in the motor such as overacceleration or excessive deceleration.

For data acquisition, current probes were employed to measure the 3-phase current input to the motor in amperes. The analysis of this data required the use of various software tools. STMCube-Monitor allowed us to view and adjust any variable in the inverter controller's code in real-time, providing crucial insights into the system's behavior under test conditions. PicoScope enabled us to visualize current and voltage values at different points within the physical setup, offering a detailed view of the electrical characteristics. Additionally, Octave was utilized to generate graphs, facilitating a comprehensive analysis of the test data.

Ensuring safety during the testing and verification process was paramount, given the high voltages and currents involved. Protections were integrated into both the software and the physical test bench to reduce the risk of equipment damage and enhance safety. The implementation of Finite State Machines (FSMs) was very important in this regard. FSMs ensured that tests only commenced when specific conditions, such as input voltage limits, were met. Moreover, they were designed to halt the tests immediately if parameters like the 3-phase currents exceeded predefined thresholds.

5.6 Speed Controller

The speed controller is responsible for generating a reference I_q from the error between the current speed of the motor and the desired speed. This controller plays a critical role in maintaining the

imposed speed control, ensuring that the motor operates at the set speed despite load variations and other disturbances. Similar to the current controller, the speed controller utilizes a Proportional-Integral block to generate the reference I_q based on the speed error.

For effective implementation, it is essential to finely tune the gains of the speed controller, specifically the proportional gain (K_p) and the integral gain (K_i). Initial gain values were derived from simulations, providing a starting point for tuning. Nevertheless, real-world conditions often differ from simulated environments, demanding further adjustments for the gain values to achieve optimal performance.

The tuning process involved two main steps:

Gain Calibration: The K_p and K_i values were adjusted to minimize the speed error and ensure a stable response. This process involved iterative testing and analysis to find the optimal gain settings that provide the best transient and steady-state performance.

Time Constant Coordination: It was crucial to ensure that the time constant of the speed controller is appropriately larger than that of the current controller. This coordination prevents persistent errors between the controllers and ensures system stability. A faster current controller relative to the speed controller allows for smooth and stable motor control, reducing the likelihood of high current changes and deviations.

5.6.1 Tests Conducted

To validate the performance of the speed controller, several tests were conducted:

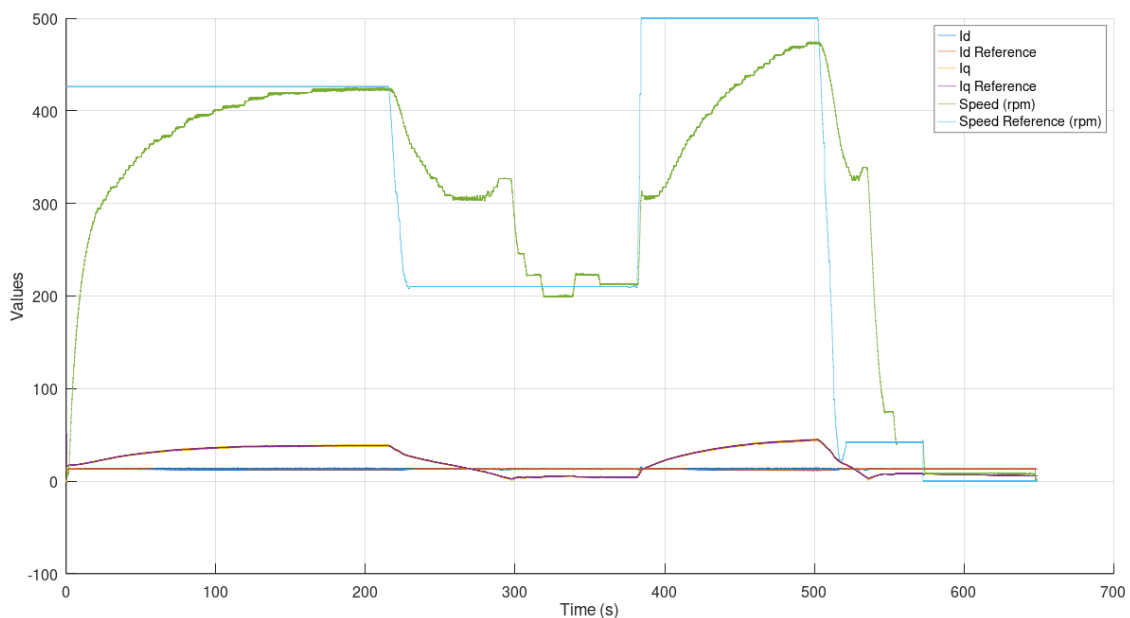


Figure 5.9: Step Response test.

As observed in the graph in Figure 5.9, the speed controller follows the reference values, whether they are high or low. However, there are noticeable anomalies around the 300 rpm mark

during the first descent and around the 340 rpm mark during the second. These anomalies appear as sudden increases in speed and can be explained by the characteristics of the speed controller's time constant.

When the speed reference is changed, the proportional part of the controller responds almost immediately, pulling the velocity towards the new reference value. The sudden increases observed are due to the integral part of the controller beginning its action. Before the speed reference change, the integral component was continuously integrating the error from the previous reference. When the reference changes, the integral part still has some accumulated error to integrate before it can start adjusting to the new reference. This residual integration from the previous reference causes a brief increase in speed.

Once the integral component finishes processing the residual error, it begins to integrate the error based on the new speed reference. This results in the speed controller making the necessary adjustments to reach and stabilize at the new reference speed.

Steady-State Error Test: This test measures the steady-state error to ensure the controller maintains the desired speed with minimal deviation under constant conditions. By running the motor at a fixed speed for an extended period, it is observable the accuracy and stability of the speed control, ensuring that the steady-state error remains within acceptable limits.

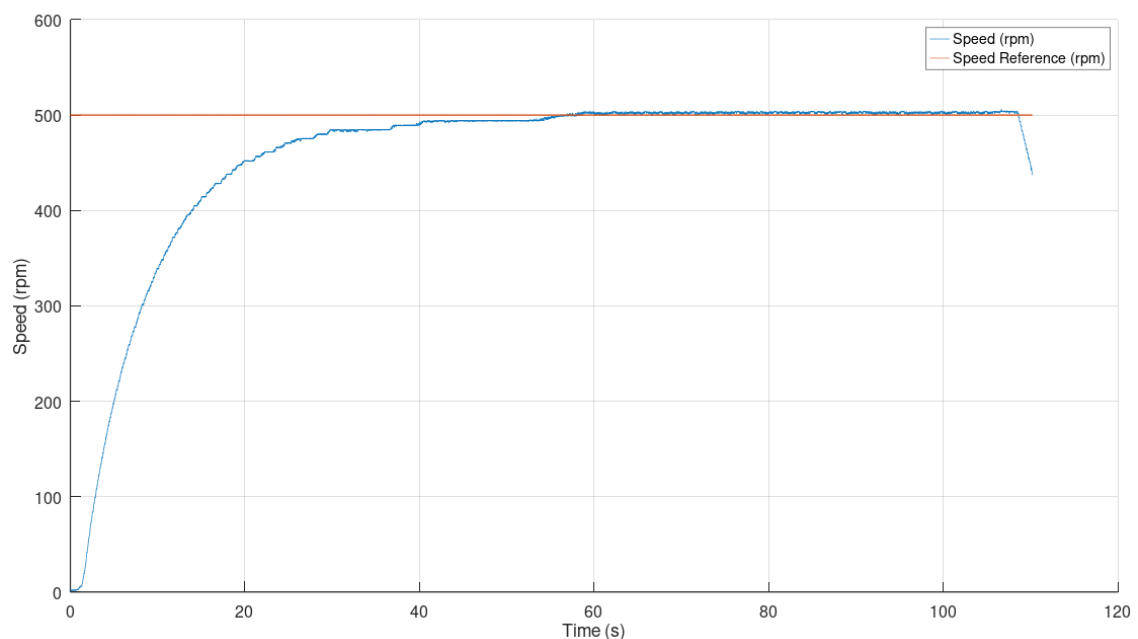


Figure 5.10: Steady State test (Speed).

From the graph in Figure 5.10, it is observable that the speed controller effectively matches the speed reference and maintains this value consistently without deviation. The motor's acceleration profile is very smooth, indicating precise control and the absence of any overshoot. This performance is a strong indicator of well-tuned PI gains, demonstrating the controller's ability to achieve and sustain the desired speed accurately.

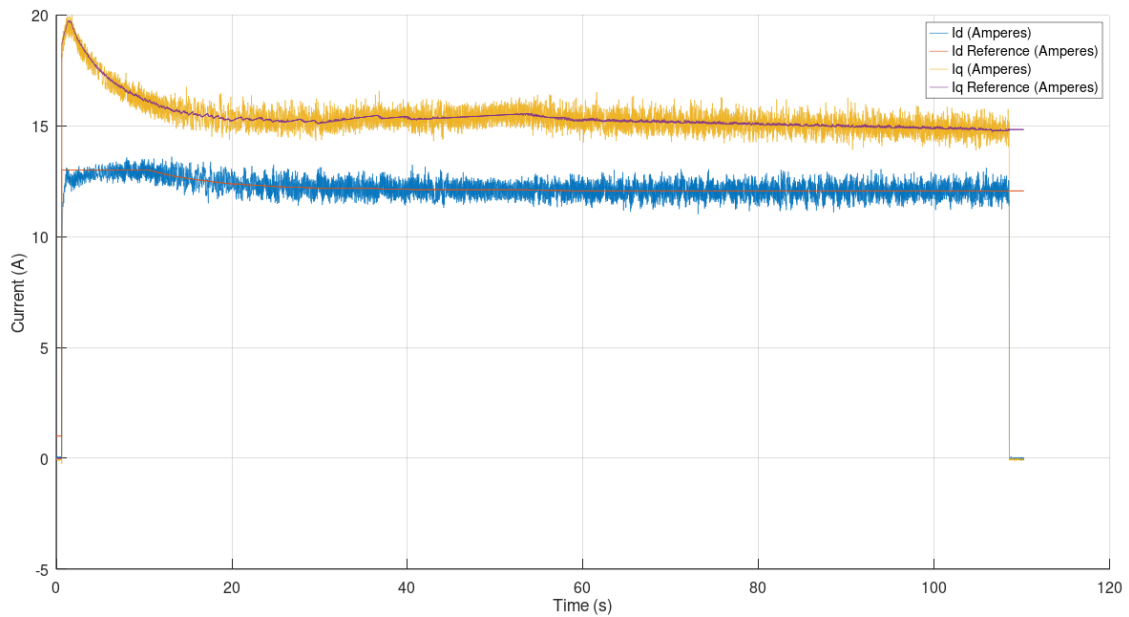


Figure 5.11: Steady State test (Currents).

The graph in Figure 5.11 illustrates the current response during the steady-state test. Initially, there is an increase in the I_q current, which is required to generate the torque needed to overcome the motor's starting inertia. As the motor begins to rotate and inertia starts to assist, the I_q current decreases and stabilizes around 15 Amperes. This stabilized value indicates the amount of torque-inducing current necessary to maintain the motor's rotation at the set speed. It's important to note that the I_q value varies with different speeds and loads. For instance, at higher speeds or lower loads, the I_q required would be lower. In scenarios with negligible load or when the rotor is free-running, the I_q in a steady-state situation at nominal speed can be close to 0 Amperes.

Additionally, the magnetizing current, I_d , also plays a crucial role. As speed increases, the I_d must be adjusted to ensure the motor remains magnetized with a slip different from zero. In this particular test, the I_d starts at approximately 13 Amperes and stabilizes around 12 Amperes. This adjustment is part of a process known as Field Weakening, which allows the motor to operate efficiently at higher speeds by reducing the magnetizing current to prevent excessive flux in the motor.

Overall, the steady-state tests for both speed and currents demonstrate the effectiveness of the speed controller in maintaining the desired operational parameters, ensuring both accuracy and stability.

5.6.2 Field Weakening

As mentioned beforehand, Field Weakening is a technique used to allow a motor to accelerate by reducing the magnetizing current, I_d . This reduction must be done carefully because an incorrect

I_d value, whether too high or too low, can cause instability in the motor currents. This instability can result in inefficient or unresponsive motor behavior and, in extreme cases, damage the equipment.

In simulations, it was employed a more aggressive Field Weakening approach, directly comparing the reference speed to the actual speed. However, applying such an aggressive reduction in a real-world scenario, like the motor in this test bench, could lead to rapid demagnetization and instability.

In the field weakening region, the magnetizing current (I_d) changes dynamically, affecting the d-axis voltage component (V_d). The adjustments dependent on the operating speed are able to ensure proper DC bus voltage utilization [28]. Although the simulated motor parameters closely matched the real motor parameters, practical tests revealed that the calculated I_d was much higher than what was actually required to control the motor effectively.

Before implementing any Field Weakening, it was observed that the motor began to demagnetize at speeds above 200 rpm. This is significantly lower than the nominal speed for a 6-pole motor, which is around 1000 rpm. When demagnetization occurred, the I_d and I_q currents started to deviate from their references, leading to distorted three-phase currents. Additionally, a distinct noise was heard from the motor, indicating unbalanced rotation.

The following image illustrates the d - q currents when the motor loses magnetization, highlighting the deviations and distortions mentioned.

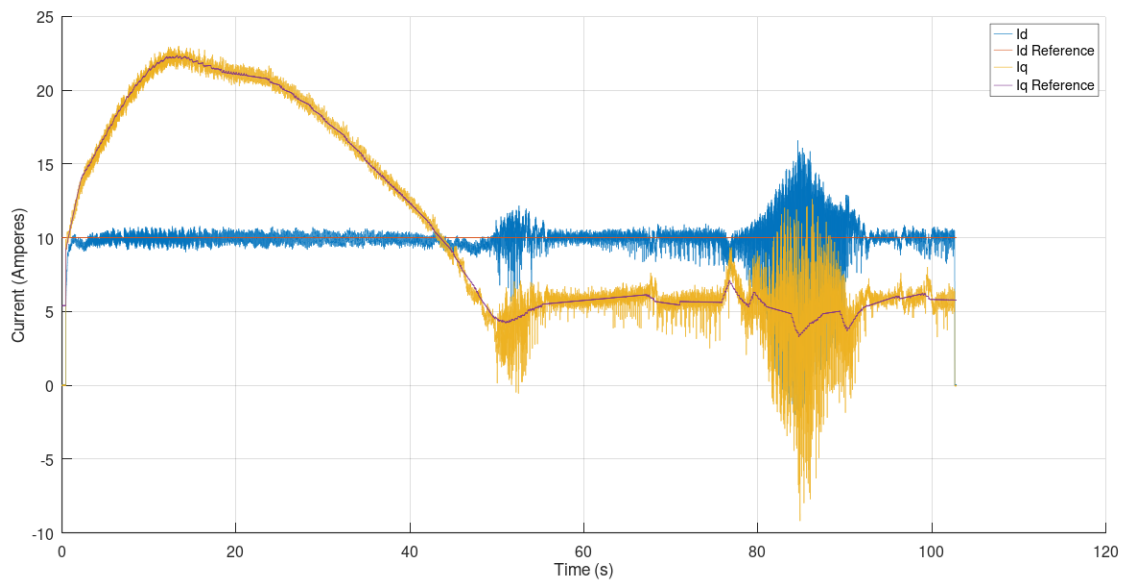


Figure 5.12: d - q Currents During Motor Demagnetization.

This figure demonstrates the importance of careful tuning and implementation of the Field Weakening technique to prevent adverse effects and maintain motor stability.

Due to time and equipment constraints, further increasing the motor speed beyond 500 rpm was not pursued. Through experimentation, an equation was developed to systematically reduce

the I_d reference value as the speed exceeded a predefined set value. For these experiments, this set value was 350 rpm. By starting with a slightly lower I_d base reference, the motor was able to exceed the 350 rpm mark without issues. After reaching this speed, the field weakening was activated.

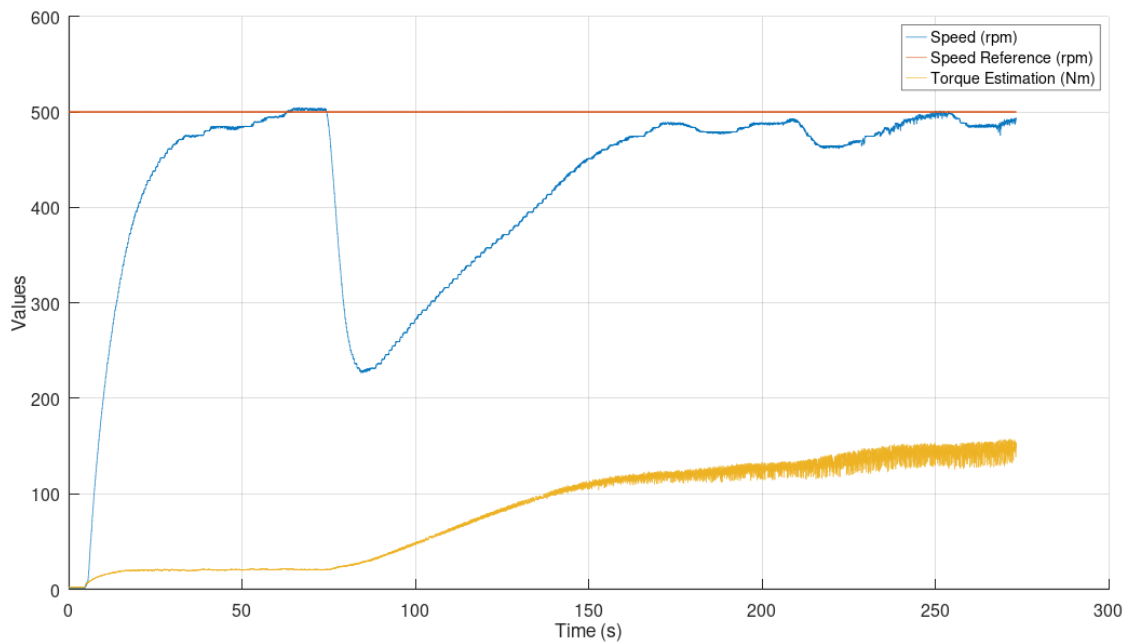
This approach allowed us to produce graphs such as those shown in the steady-state tests, particularly the one in Figure 5.11. Around the 15-second mark in this graph, it is evident that the magnetizing current begins to decline gradually. This gradual reduction enables the motor to accelerate in a stable manner, illustrating the effectiveness of this Field Weakening strategy.

5.7 Train Emulation

As mentioned earlier, in the test bench setup, the 3-phase induction motor is coupled to a DC machine. This DC motor serves as a load emulator, simulating the locomotive's physical characteristics, such as starting torque due to weight and air resistance. Additionally, it can replicate different stages of a journey, including curves, climbing inclines, descending hills, and sudden load increases. By configuring the DC motor to emulate these conditions, the controller can be adjusted and fine-tuned to closely match the performance and behavior of a controller installed in an actual locomotive. This emulation process is critical for ensuring that the control strategies are robust and effective under a variety of real-world conditions.

5.8 Load Insertion

The initial stages of train emulation lied in increasing the load of the motor utilizing the DC motor. In these situations, the driving motor would be rotating at a predefined speed and the DC motor would increase the load in a step in order to view the behavior of the controller. For these studies, the driving motor would be rotating at a speed of 500 RPM and after reaching that speed, at 75s, the DC motor would insert a given load. In these tests, it was relied on the torque estimation calculation in order to measure the behavior of the motor during these situations.

Figure 5.13: 150 Nm Load Insertion at $t=75s$.

In this study, displayed in figure 5.13, it was applied a load that resulted in an estimation of 150 Nm of torque applied in the rotor of the driving motor. It is possible to witness when the load is initially applied when the rotating speed lowers significantly to around 230 rpm, which is a little less than half of the initial speed. After a few seconds of the load being applied, the controller slowly started to increase the torque inducing current in order to overcome this substantial load that is being applied. This test was also monitored using PicoScope and with this tool it was also possible to witness the current that was being input into the motor directly from its 3-phase cables.

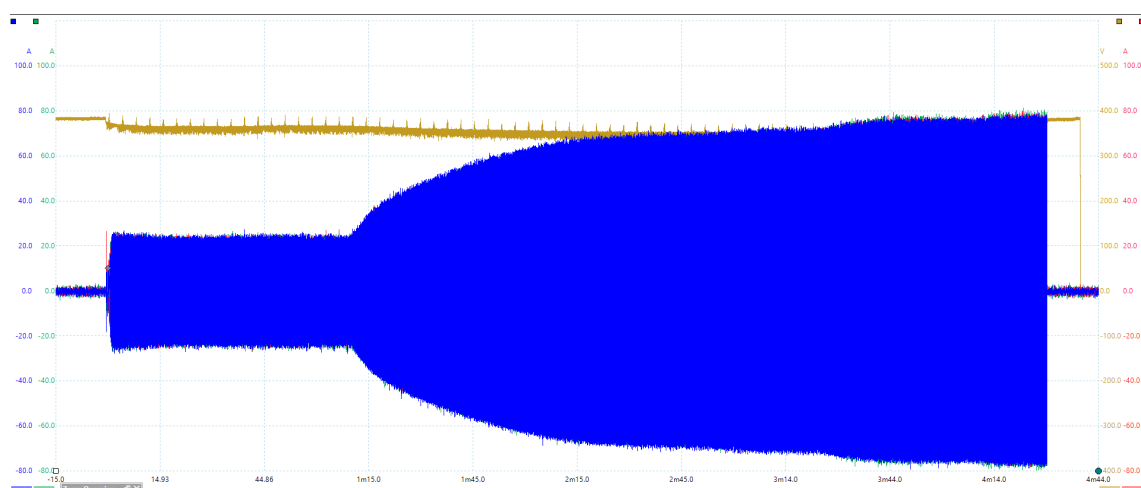


Figure 5.14: 150 Nm Load Insertion PicoScope.

In figure 5.14, it is observable the 3-phase currents (represented in blue, green, and red) overlapping each other, along with the initial voltage applied (represented in yellow). This graph clearly illustrates the moment when the load is applied and how the controller adjusts the currents entering the motor to overcome the load. The currents reach a peak amplitude of nearly 80 Amperes, which ultimately triggers the current protection mechanism, terminating the study. This test allows us to determine the maximum load that the motor can handle and provides insights into its behavior under heavy load conditions. The results can be compared to scenarios such as a train ascending a hill or encountering a sudden increase in track inclination.

5.8.1 Locomotives air resistance

Upon establishing the DC motors ability to provide a definable amount of torque proceeded in applying torque dynamically in order to start emulating real life scenarios. The first emulation regarded the phenomenon of air resistance. To apply the load, the following equation was used:

$$F = A + Bv + Cv^2 \quad (5.1)$$

Where A is the effect of rolling resistance, B is the drivetrain losses, as well as the rolling resistance effect on velocity, and C is the aerodynamic drag. With this equation, the faster the motor was rotating, more load would be applied similarly to what would happen in a moving locomotive, since the faster it is going, the more air resistance is applied. Doing this, the following Speed curve was obtained:

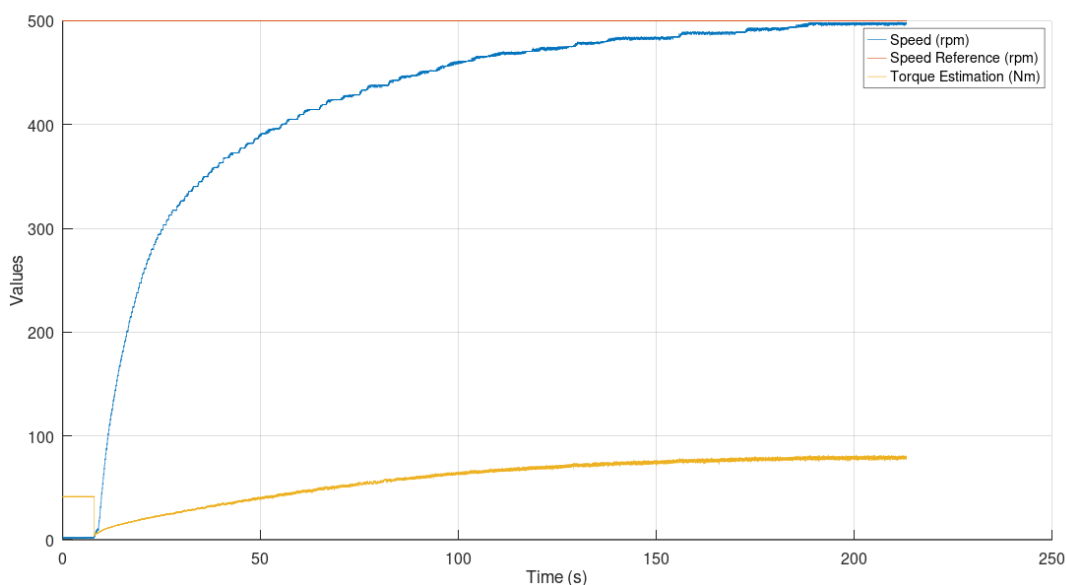


Figure 5.15: Effect of Air Resistance.

By comparing the speed curve with load applied to one without load, it is possible to observe a significant difference, in the estimated torque values. In this study, torque values close to 80 Nm

at 500 rpm were reached. In contrast, studies conducted without any air resistance showed that the motor only required a torque value of around 17 Nm, as displayed in Figure 5.16.

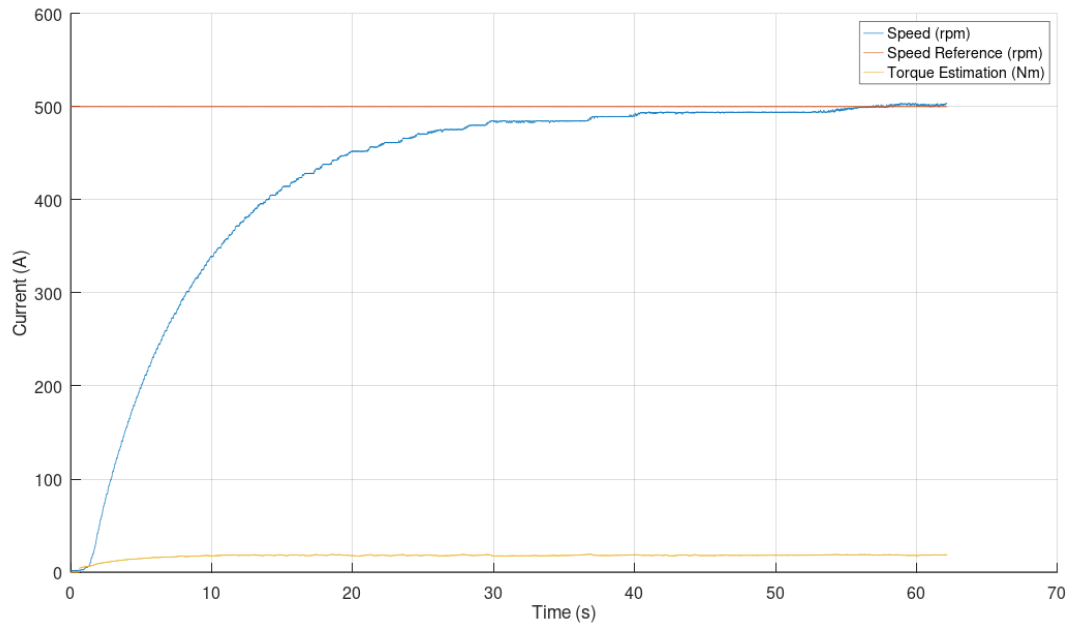


Figure 5.16: Comparison without air resistance.

It is a noticeable difference in the time that it takes to reach the speed reference, taking about a minute to reach the reference with no load and almost three and a half minutes to reach the same reference.

5.8.2 Small Journey Emulation

When all load types are combined, it is possible to simulate a study that emulates a small journey encompassing all the previously mentioned load conditions. In this simulation, the train must overcome the starting torque and air resistance, followed by navigating up hills and performing curves. By incorporating these scenarios, it is observable how the controller manages these varying conditions. The following figure showcases the behavior of the controller as it responds to these load actuations throughout the simulated journey.

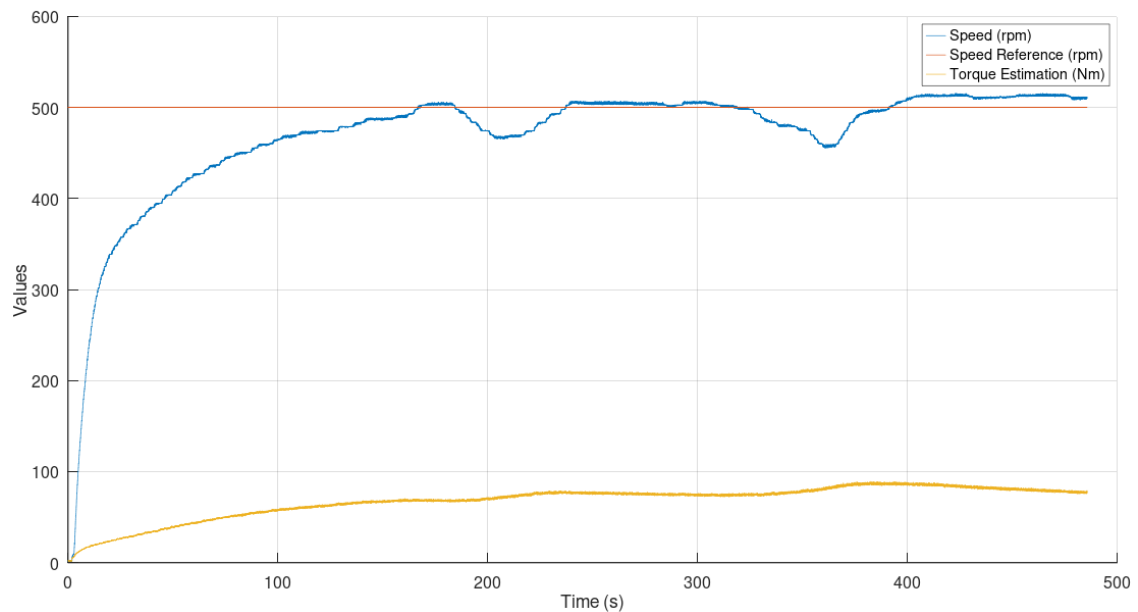


Figure 5.17: Speed and torque curve of a small journey.

Figure 5.17 illustrates the torque curve and the time it takes for the locomotive to reach the reference speed during the small journey simulation. Initially, the air resistance is in full effect as the locomotive accelerates. Upon reaching the reference speed, the locomotive encounters a small hill, followed by a quick descent. This is reflected in the torque curve, showing a temporary increase in torque needed to climb the hill and a very slight decrease some time after the descent. It is most noticeable though in the speed curve, since it is easily seen when the locomotive starts climbing the hill, when the speed starts to lower rapidly.

After a few seconds, there is a very small decrease in velocity, indicating the start of a curve. This slight reduction in speed does not pose significant challenges, as the motor promptly recovers and returns to the reference speed. Shortly after starting the curve, the locomotive climbs another hill. This hill has a sharper descent, resulting in a slight overshoot in speed. The controller adjusts accordingly, stabilizing the speed and concluding the study.

Chapter 6

Discussion

In the previous chapters, the control strategies were developed and tested extensively through simulations. These simulations served as foundation, allowing the chance to refine the methods used before moving on to physical implementation. This chapter is dedicated to presenting and discussing the results obtained from both the simulation environment and the physical test bench. By comparing these sets of results, it was aimed to validate the simulation models, identify any discrepancies, and discuss the overall performance and reliability of the control strategies. This comparative analysis is essential for understanding the effectiveness of this approach and guiding future improvements.

6.1 Simulation and Implementation: comparison of results

6.1.1 Theta Estimation

Theta estimation plays the role in accurately determining the rotor position, which is important for the efficient control of the motor. The approach to theta estimation remained consistent in both the simulation and the emulation phases of this project. In the simulation, PSIM was utilized to calculate the Theta frequency directly, resulting in a value of approximately 25 Hz when the motor operates at 500 rpm.

For the emulation, the same estimation method was applied. Despite the presence of under-sampling in STMCubeMonitor, which made the Theta graph less readable compared to the simulation output, the estimated Theta frequency was also found to be around 25 Hz. This close match in Theta frequencies between the simulation and emulation validates the accuracy and reliability of this estimation technique.

6.1.2 Current Control

The implementation and performance of the Current Controller revealed noticeable differences between the simulation and the real-world test bench scenarios. These differences are primarily due to the unavailability of exact conditions and precise motor parameters in the simulation

environment, making it challenging to create an exact replica of the test bench setup.

In the simulation, achieving the target speed of 500 rpm required a significant amount of current to ensure a controlled and smooth acceleration. The d - q currents, as depicted in Figure 6.1, illustrate this requirement. The simulation predicted that a substantial current was necessary to maintain stability and follow the speed reference effectively.

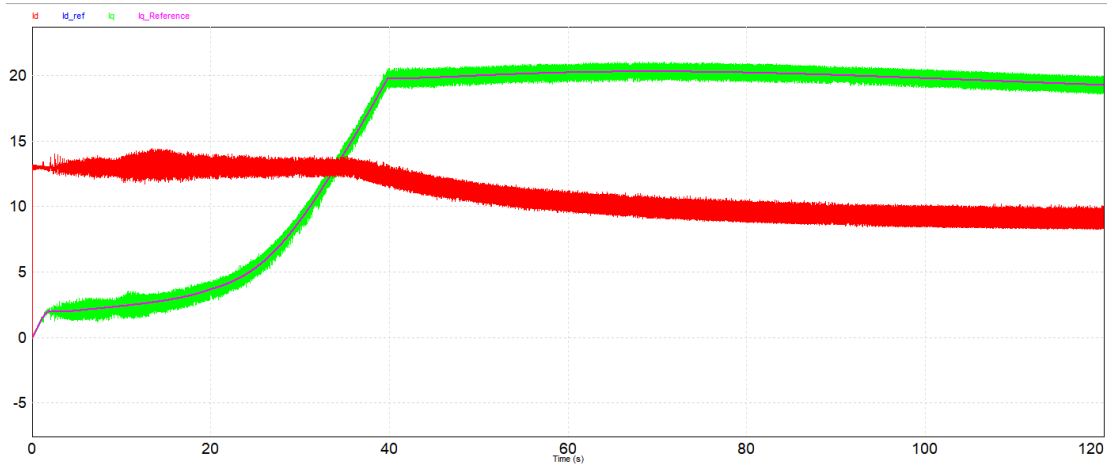


Figure 6.1: d - q currents in simulation.

Conversely, the results from the test bench were quite different. In the real-world scenario, the motor reached the desired speed of 500 rpm with approximately half the current that the simulation predicted. This discrepancy highlights the challenges of simulating complex physical systems accurately without precise data. The d - q currents obtained from the test bench, shown in Figure 6.2, demonstrate a more efficient current utilization compared to the simulation.

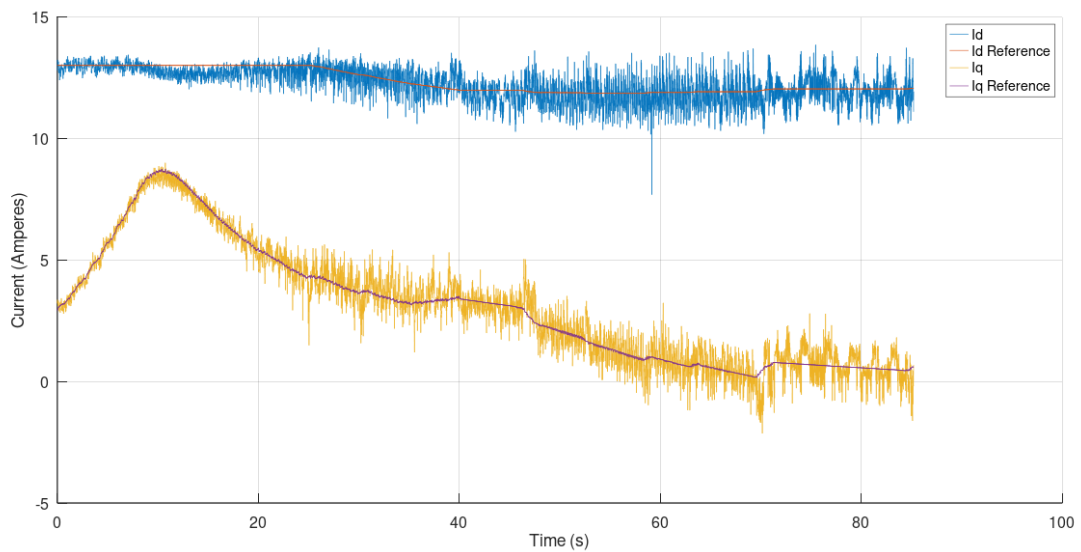


Figure 6.2: d - q currents in emulation.

Adding to the difference in magnitude, it can also be observed a distinct difference in the current curves. While the general pattern over time is similar in both the simulation and the emulation — with I_q increasing to overcome the starting torque and then decreasing after passing the reference — the behavior in the simulation is much more controlled and thus significantly slower in its progression.

In the simulation, the increase in I_q and its subsequent stabilization are noticeably delayed. This delay causes I_q to start decreasing only when the speed reference reaches the 500 rpm mark. At this point, the reference stops increasing, leading to the stabilization of I_q . After a few seconds, once the system has fully stabilized, it is possible to finally see a decrease in I_q .

6.1.3 Speed Control

The differences observed in the Speed Controller are relatively minor. Despite slight variations, the speed curves exhibit similarity between simulation and emulation scenarios. To simulate a realistic scenario, it was implemented a linear speed increase program, aiming to reach the 500 rpm mark within 40 seconds.

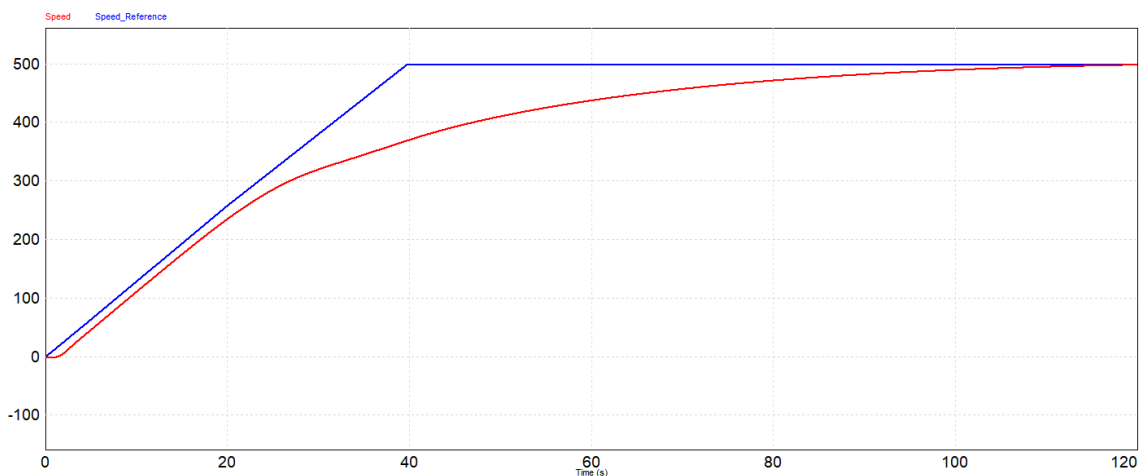


Figure 6.3: Speed curve in simulation.

In the simulation phase, significant efforts were directed towards fine-tuning the speed controller to achieve a smooth curve typical of locomotive operations. As detailed in previous chapters, this involved deliberately slowing down the controller to attain the curve depicted in Figure 6.3.

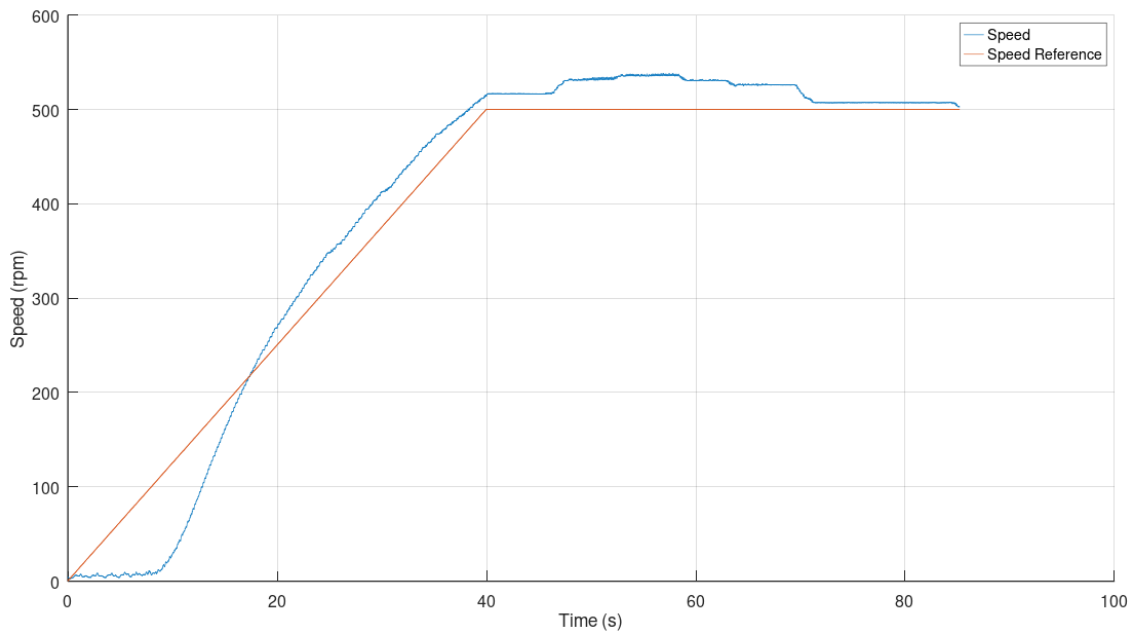


Figure 6.4: Speed curve in emulation.

During emulation, as displayed in figure 6.4, the controller exhibits a similarly smooth response but acquires speed more swiftly due to inherent inertia. This characteristic results in an initial overshoot above the reference speed, which is subsequently corrected with a slight margin. Upon reaching the 500 rpm set value, the speed plateaus momentarily before experiencing a slight increase a few seconds later, influenced by the controller's higher time constant. As the Integrator block adjusts, the speed converges to and stabilizes at the reference value.

6.1.4 Flux Command

In terms of the Flux Command, it was utilized Field Weakening as previously discussed. Both the simulation and the emulation were configured to start field weakening upon surpassing the reference speed of 350 rpm.

In the simulation, Field Weakening served primarily to smooth the speed curve. As illustrated in Figure 6.1, once the speed exceeds the 350 rpm mark, the reference begins to lower, causing a gentler slope in the speed curve. This effect is evident in Figure 6.3, where a noticeable decrease in the slope occurs at around 350 rpm. This approach ensured a controlled and gradual increase in speed, preventing an overshoot, allowing for a smoother acceleration.

In the emulation, Field Weakening was applied for its intended purpose. Despite operating at slightly above the rated voltage of the motor (400 V), initial tests indicated the necessity to reduce the magnetizing current (I_d) when surpassing the 350 rpm mark. Initially, it was attempted to lower the reference from the start, but this led to insufficient magnetizing current for effective startup, causing operational issues. Consequently, field weakening was introduced at 350 rpm to manage this problem, allowing the motor to reach 500 rpm smoothly.

Comparing the simulation and emulation, the behavior of the I_d reference remained consistent, which is evident when comparing Figure 6.1 with Figure 6.2. Both figures demonstrate that, upon reaching 350 rpm, the I_d reference begins to decrease, highlighting the effective implementation of field weakening in both scenarios. This consistency underscores the robustness of the approach and its applicability in both simulated and real-world environments.

The use of Field Weakening not only ensured smoother speed transitions, but also provided the necessary control over the magnetizing current, which is necessary for maintaining motor performance and efficiency. By starting field weakening at 350 rpm, it had been successfully managed the trade-off between speed and torque, achieving a balance that allowed for stable and efficient operation across the motor's speed range. This approach highlights the importance of fine-tuning control strategies to match the specific requirements of both simulated and real-world applications.

6.2 Discussion

The comparative analysis between simulation and emulation results has provided valuable insights into the performance and robustness of these control strategies. While there are discrepancies between the two sets of results, these differences do not undermine the validity or applicability of these methods. Instead, they highlight the complexities and challenges inherent in accurately modeling and emulating physical systems.

Overall, the control system demonstrated reliable behavior across various tests. The theta estimation, despite being less visually clear in the emulation due to undersampling of STMCube-Monitor, it still proved that the Theta value in both the simulation and emulation was consistent. This consistency reinforces the robustness of this theta estimation method, ensuring precise rotor position information for correct motor control.

In the realm of current control, the emulation revealed a more efficient current utilization compared to the simulation. Although the simulation required more current to reach the target speed, this discrepancy is attributable to the inherent differences between simulated and real-world conditions. The overall pattern of current variation, with initial increases to overcome starting torque followed by stabilization, was consistent across both scenarios, displaying the effectiveness of the control strategy.

Speed control tests yielded similar speed curves in both the simulation and emulation, with minor variations. The implementation of a linear speed increase program allowed for a realistic assessment of the controller's performance. Despite the emulation showing a slightly faster speed gain due to inertia, both environments achieved smooth transitions and minimal overshoot, validating the controller's suitability for locomotive applications.

The application of field weakening further demonstrated the system's capability to manage speed transitions effectively. Both the simulation and emulation employed field weakening upon surpassing 350 rpm, ensuring controlled and gradual speed increases. This strategy successfully balanced speed and torque, maintaining motor performance and efficiency across the speed range.

The consistent behavior of the I_d reference in both scenarios highlights the robustness of this approach.

In conclusion, while the simulation and emulation results are not identical, they collectively affirm the viability of these control strategies. The discrepancies observed are valuable for refining the models and improving accuracy in future work. The ability to achieve smooth speed curves, and effective current and flux management in both environments speaks to the overall success of this project. These findings provide a solid foundation for further development and optimization of railway traction converter controllers, paving the way for advancements in locomotive motor control technology.

Chapter 7

Conclusion

7.1 Summary

For this dissertation, the primary objective was to simulate and implement a Railway Traction Converter Controller, a critical component responsible for driving the motor based on the train driver's commands.

To achieve this goal, an in-depth study of the relevant theory was previously undertaken, ensuring a comprehensive understanding of the subject matter. It was explored the types of semi-conductors involved, evaluated various control strategies, selecting Field Oriented Control, and examined modulation techniques, opting for Space Vector Modulation.

The next phase involved extensive simulation efforts to recreate the Converter Controller in a virtual environment. This stage was necessary for calibrating and fine-tuning the controller, specially in terms of the PI controller gains, to meet the requirements of railway applications.

Upon achieving satisfactory simulation results, it was transitioned to implementing the controller on a physical test bench. This step aimed to replicate real-world conditions as closely as possible. After additional rounds of calibration and adjustments of the PI gains, the system was subjected to testing by varying the motor load, simulating realistic locomotive journey conditions to evaluate its capabilities.

In the final analysis, the results were compared from the simulation and emulation, drawing conclusions about the reliability and performance of this work.

Overall, this project was a success. Accomplishing the goal of simulating and implementing a Railway Traction Converter Controller. Although it was not developed a controller ready for immediate use on the LE5600 series locomotive, it was laid a solid foundation and gained the necessary knowledge to build one from scratch in the future or adapt it for other locomotive models.

7.2 Contributions

The contributions of this dissertation address the objectives outlined in the introduction chapter:

Contribution 1: The results presented in the controller computational analysis chapter demonstrate that the fundamentals and simulation methodology are effective for the development of a traction controller. This addresses the first and second objectives of the dissertation.

Contribution 2: The results detailed in the controller implementation chapter show that the fundamentals and implementation methodology are practical and useful for developing a traction controller, therefore answering the first and third objectives of the dissertation.

Contribution 3: This dissertation demonstrates a methodology for developing traction controllers through iterative processes between simulation and implementation. This contribution directly aligns with the dissertation's title.

7.3 Future Work

Despite this comprehensive exploration of this project's scope, certain aspects remained unaddressed due to various constraints. Future work could focus on several key areas to enhance and expand upon these findings:

1. **Achieving and Exceeding Rated Motor Speed:** Investigating the Field Weakening strategy further to reach and potentially surpass the motor's rated speed of 1000 rpm, coupled with an increase in input voltage to 500V.
2. **Exploring Torque Control:** Since some locomotives utilize torque control rather than speed control, examining this control strategy could provide valuable insights and broaden the applicability of this converter controller.
3. **Obtaining Accurate Motor Parameters:** Conducting locked rotor and no-load tests on both the test bench motor and the motors from the LE5600 series locomotive to acquire precise motor parameters. This data would enable the creation of more accurate and fitting converter controllers.

By addressing these areas, future research can build upon the groundwork laid in this dissertation, further refining and optimizing railway traction converter controllers for real-world applications.

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