CONTROL OF A PMSG BASED WIND ENERGY CONVERSION SYSTEM USING ABC-FRAME UNDER ASYMMETRICAL AND DISTORTED GRID-VOLTAGE CONDITIONS

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Aos meus pais, Dimas e Kátia. À minha querida irmã Talitha. À Sarah, amada noiva. Aos meus avós, Rui e Maria Odila, Dona Onilda e Sr. Geraldo (in memoriam). Aos meus amigos.
"This is a faithful saying, and worthy of all acceptation, that Christ Jesus came into the world to save sinners; of whom I am chief."

I Timothy 1:15
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RESUMO

A demanda por energia elétrica tem sido crescente, gerando preocupação e ampla discussão no cenário mundial no que concerne ao esgotamento dos recursos naturais e emissão de carbono na atmosfera. Neste sentido, é consenso por parte de órgãos governamentais e ambientais, que o suprimento desta demanda necessita ser feito de forma sustentável. Com isto, investimentos maciços em fontes sustentáveis de energia elétrica tem sido uma realidade em muitos países. Tais investimentos evocam o desenvolvimento de pesquisas em tecnologias de conversores eletrônicos, que são os responsáveis pelo condicionamento da energia de muitas destas fontes para conexão destas na rede elétrica. Dentre elas, a energia eólica cumpre um papel de suma importância por ter uma grande participação na geração sustentável de energia elétrica. Além do investimento expressivo em fontes sustentáveis por parte dos governos, outros fatores como, o avanço tecnológico da eletrônica de potência e dos geradores de fontes renováveis, estes apresentando melhorias em eficiência de conversão de energia, tem contribuído para um aumento da participação de fontes renováveis através da geração distribuída ao longo das redes de média e baixa tensão. No contexto de geração distribuída, os conversores eletrônicos podem ser mais bem aproveitados se a eles são adicionadas capacidades que vão além do simples fornecimento de potência ativa. Eles podem ser utilizados para também dar suporte à rede elétrica em compensação de potência reativa e de distúrbios de cargas. A predominância de cargas desbalanceadas e não lineares ao longo das redes de distribuição, são motivo de assimetria e distorção na tensão do ponto de acoplamento comum, desafiando os inversores e controladores de corrente que devem ser impermeáveis a tais perturbações. Assim, este estudo propõe avaliar o desempenho operacional de um sistema de energia eólica baseado em uma máquina síncrona de ímã permanente, que foi concebido um esquema de controle desenvolvido em uma estrutura em coordenadas naturais sob condições de tensão da rede não ideais. Considerou-se que tal sistema também pode realizar compensação de corrente de carga, agregando mais funcionalidades e aumentando sua relação custo-benefício. Além do mais, uma solução para o desacoplamento entre torque e fluxo para o controle da máquina Permanent Magnet Synchronous Generator em abc foi desenvolvida. Através de simulação computacional, o desempenho do controle do lado da rede foi avaliado compensando-se a componente de corrente não ativa para carga não linear, frente a distúrbios de assimetria e distorção harmônica das tensões do ponto de acoplamento comum. A operação do gerador com a solução de controle apresentada, frente a variações de torque devidas às variações na velocidade do vento, também foi avaliada. Por fim, implementação experimental validou o controle de corrente para o lado da rede, bem como a solução apresentada para o lado da máquina.

Palavras-chave: Energia Eólica; Coordenadas ABC; Gerador Síncrono de Ímãs Permanentes; Compensação de Carga.
The steady growth in the power demand has leading to wide discussion about natural resources exhaustion and carbon emission, worldwide. In this sense, it is a consensus on the part of governmental and environmental agencies that such demand should be supplied in a sustainable way. Thus, massive investments in sustainable energy resources have been a reality in many countries. Such investments evoke some development in electronic converters technology through researches, once power converters are responsible for energy conditioning for grid connection of many renewable power generators. Among renewables, wind power plays a key role because of its great share in sustainable power generation market. Besides significant investment in sustainable resources by governments, other factors such as technological advancement of power electronics and renewable energy devices, which have achieved improvements in energy conversion efficiency, have contributed to the share of renewable sources in distributed generation throughout medium and low voltage networks. In distributed generation context, electronic converters can be better utilized if to them are added capabilities that go beyond simple active power supply. Power converters can also be utilized for support the grid, compensating reactive power and load disturbances. The predominance of unbalanced and non-linear loads along distribution networks, causes asymmetry and distortion in voltages at common coupling point. This challenges power inverters and their current controllers, which must be impermeable to such disturbances. So, this study proposes to evaluate operational performance of a permanent magnet synchronous generator based wind power system under non-ideal grid voltage conditions, for which a control scheme in abc-reference-frame was conceived. It was considered that such a system can also perform load current compensation, which adds more functionality and increases its cost-effectiveness. In addition, a solution for flux and torque decoupling was developed for permanent magnet synchronous machine abc-reference-frame control. Through computational simulation, the performance of the grid-side control was evaluated for non-linear-load current compensation, in face of asymmetric and distorted grid-voltages. Furthermore, operation of the generator against torque disturbances due to variations in wind speed was also simulated, then evaluating the performance of the developed solution for machine-side-control in abc-reference-frame. Finally, experimental results validated the current control for machine-side, as well as the presented solution for machine-side.

**Keywords:** Wind Power; ABC-Reference-Frame; Permanent Magnet Synchronous Generator; Load Compensation.
# LIST OF ABBREVIATIONS

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Full Form</th>
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<tbody>
<tr>
<td>abcRF</td>
<td>abc Reference-Frame</td>
</tr>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>CPT</td>
<td>Conservative Power Theory</td>
</tr>
<tr>
<td>CPWM</td>
<td>Continuous Pulse-Width-Modulation</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DFIG</td>
<td>Doubly-Fed Induction Generator</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processor</td>
</tr>
<tr>
<td>EMF</td>
<td>Electromotive Force</td>
</tr>
<tr>
<td>FLFW</td>
<td>Four-Leg Four-Wire</td>
</tr>
<tr>
<td>GS</td>
<td>Grid-Side</td>
</tr>
<tr>
<td>IGBT</td>
<td>Insulated Gate Bipolar Transistor</td>
</tr>
<tr>
<td>IPT</td>
<td>Instantaneous Power Theory</td>
</tr>
<tr>
<td>LCCE</td>
<td>Energy Conversion and Control Laboratory</td>
</tr>
<tr>
<td>MPPT</td>
<td>Maximum Power Point Tracking</td>
</tr>
<tr>
<td>MS</td>
<td>Machine-Side</td>
</tr>
<tr>
<td>PCC</td>
<td>Point of Common Coupling</td>
</tr>
<tr>
<td>PFC</td>
<td>Power Factor Correction</td>
</tr>
<tr>
<td>PI</td>
<td>Proportional-Integral</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<td>--------------</td>
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</tr>
<tr>
<td>PLL</td>
<td>Phase-Locked-Loop</td>
</tr>
<tr>
<td>PMSG</td>
<td>Permanent Magnet. Synchronous Generator</td>
</tr>
<tr>
<td>PMSM</td>
<td>Permanent Magnet Synchronous Machine</td>
</tr>
<tr>
<td>PR</td>
<td>Proportional Resonant</td>
</tr>
<tr>
<td>pu</td>
<td>Per-Unit</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse-Width-Modulation</td>
</tr>
<tr>
<td>SMO</td>
<td>Sliding-Mode Observer</td>
</tr>
<tr>
<td>SRF</td>
<td>Synchronous Reference Frame</td>
</tr>
<tr>
<td>SVPWM</td>
<td>Space-Vector Pulse-Width-Modulation</td>
</tr>
<tr>
<td>THD</td>
<td>Total Harmonic Distortion</td>
</tr>
<tr>
<td>TLFW</td>
<td>Three-Leg Four-Wire</td>
</tr>
<tr>
<td>TLTW</td>
<td>Three-Leg Three-Wire</td>
</tr>
<tr>
<td>UFMG</td>
<td>Federal University of Minas Gerais</td>
</tr>
<tr>
<td>VSI</td>
<td>Voltage Source Inverter</td>
</tr>
<tr>
<td>WECS</td>
<td>Wind Energy Conversion System</td>
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<tr>
<td>WRSM</td>
<td>Wound Rotor Synchronous Machine</td>
</tr>
<tr>
<td>WT</td>
<td>Wind Turbine</td>
</tr>
<tr>
<td>VSR</td>
<td>Voltage Source Rectifier</td>
</tr>
<tr>
<td>DG</td>
<td>Distributed Generator</td>
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CHAPTER 1

Introduction

1.1 Contextualization

Some factors as the growth of world population, the constant need for more food and industrial production and development of new technologies have been rising up the global electric power demand and consumption. The data presented in Figure 1.1 confirms this statement as it shows an upward trend in electricity consumption in the world from 1990 to 2017.

In this context, sustainability has been a general global concern. In order to avoid carbon emission increasement and natural resources exhaustion, the most of the political efforts in increase power generation is to make this throw clean energy sources like solar, wind, wave sea, geothermal and many others. In this scenario, wind energy conversion systems (WECSs) play a key role on large, medium and small scale electric generation, due to the technological advances in materials and size of installations, which allows better use of wind mechanical energy.

Figure 1.1 – Trend of global electric power consumption (ENERDATA, 2018)
It is estimated that between 2000 and 2016 the average growth rate was 23.8% a year. Moreover, in 2016 there was an increase of 50.2 GW in the global wind power generation. It accounts for 23% of the world generating capacity. In this context, China had the largest share of global wind power generation. It was a participation in about 25.1%, overcoming the United States. Denmark is pioneering in wind power system, having notably 100% of the WECS of all world in 1980. Nowadays, Denmark is among the 20 largest wind energy producers in the world and is still in the first position, having a participation of 42.5%. In the second position it is found Portugal, followed by Spain. Brazil in this scenario has a significant role, as it was the 7th country in WECS, 5th in power expansion, and 1st in capacity factor in 2016 (Energia Eólica no Brasil e no Mundo: Ano de Referência - 2016, 2017).

WECS can be composed of utility-scale wind farms, which are connected to a national electric grid and may be located onshore or offshore (YARAMASU, WU, et al., 2015). Such systems are typically greater than 20MW and are composed of a group of large wind turbines over 100kW capacity each one (AMERICAN WIND ENERGY ASSOCIATION, 2018), (WIND ENERGY TECHNOLOGIES OFFICE, 2018).

Wind power systems can also be connected, near or at the point of end-use, in a distributed generation approach. Agricultural, commercial, industrial, and community sites are examples of places where distributed wind power plants may be found at. Wind turbines in a common range from 1kW to 10kW have been installed at homes, being connected at the point of common coupling (PCC), on the customer side of the meter. Also, it is found distributed wind turbines in communities, in order to supply local electricity. Figure 1.2(a) and (b) illustrates those statements, where it is shown a small wind turbine with a rated capacity of 1.8kW at a house and another of 0.6MW at a community, respectively (WIND ENERGY TECHNOLOGIES OFFICE, 2018). Furthermore, it is noted that there are some photovoltaic modules on the roof of the house, depicting the renewables plurality in distributed generation.

The aforementioned factors show that the effort in renewables and the advance of power electronics technology have boomed the penetration of distributed generators (DGs) over medium and low-voltage grids. Such distributed power generation paradigm increases the use of power electronics and the amount of electric power generated with efficiency, reliability and diversity. However, this task brings new challenges into the electric power system. As the number of distributed sources increases, it changes the voltage level profile along the power lines.
The heavy active power generation gradually increments the voltage level on the direction starting from distribution transformers towards loads, in radial feeders (WIDEN, 2010), (ANGELOPOULOS, 2004). In this context, some support on voltage level improvement is necessary. Thus, reactive power compensation capability may be useful as an additional service of DGs. Another relevant functionality related to reactive power injection is voltage support at the PCC during voltage sags, as developed in (MIRET, CAMACHO, et al., 2013).

Nevertheless, another power quality issue remains and may be targeted of improvement. This is line current and voltage distortion due to industrial and domestic non-linear loads over the grid. So, the current scenario leads to a conclusion that current harmonic compensation is a suitable and important functionality in renewable sources power converters, which is widely addressed in literature (CHAUDHARY, LASCU, et al., 2016), (ZHAO, MENG, et al., 2018), (BURGOS-MELLADO, HERNÁNDEZ-CARIMÁN, et al., 2017).

In fact, multiple additional functions, besides the active power delivering, have been considered for implementation in DG power conditioners. This is depicted in some works as, (MARAFÃO, BRANDÃO, et al., 2015), (BUBSHAIT, MORTEZAEI, et al., 2017) and (BONALDO, PAREDES e POMILIO, 2016), in which multiple functions for support on power quality enhancement are explored, such as load harmonic, unbalancing and reactive power compensation, voltage support and others. In the context of WECS, for multi-functionality of a DG to improve the power quality at the PCC, it can be highlighted in (BUBSHAIT, MORTEZAEI, et al., 2017).
1.2 Motivation

Power theories have been developed (KUKAčKA, KRAUS, et al., 2016) and are good tools for multi-functionality of DGs. However, they have been source of divergence in some aspects concerning physical meaning of current and power terms. These divergences are amplified under distorted and unbalanced voltages condition, as shown in (MONTERO, CADAVAL e GONZALEZ, 2007). In this context, it compares the performance of three-phase four-wire active power filters using instantaneous \( p-q \) strategy (\( p-q \) theory), \( i_r-i_q \) method, also known as synchronous reference frame, unity power factor (UPF) strategy, and perfect harmonic cancelation (PHC) strategy. Such paper concludes that \( p-q \) strategy and \( i_r-i_q \) method are the most sensitive to distortion and asymmetric voltages. On the other hand, it shows that stationary frame approaches demonstrate better performance. In fact, they have the advantage of dealing directly with unbalances in current references and grid voltages (HAYASHI e MATAKAS, 2017).

In the recent scenario of micro-grids, weak grids are not able to maintain steady frequency and voltage and new power quality issues show up, with the need of higher selectivity in signal decomposition by modern power theories. So, (HARIRCHI e SIMOES, 2018) develops a method to enhance the instantaneous power theory and then compared it to the Conservative Power Theory (CPT). Both are the main power theories nowadays.

Within this context, the CPT is used herein, because it is employed in natural-reference-frame (\( abc \)). It is known that \( abc \)-reference-frame has a stationary characteristic and it does not employ frame transformation. Moreover, this theory presents a high selectivity of orthogonal current and power terms. These factors make this theory suitable for modern power networks in even critical non-sinusoidal voltage and non-linear and unbalanced currents. Moreover, the CPT is focused on load disturbances decomposition, which makes this theory a good choice for power quality improvement in modern distribution grids, which are depleted of non-linear loads.

Then, based on the conclusions of (MONTERO, CADAVAL e GONZALEZ, 2007) regarding better operation performance under non-ideal voltage condition, and considering WECS based on permanent magnet synchronous generator (PMSG) for modern power network application, this work is motivated to devise the PMSG control scheme in \( abc \)-frame. It is expected to operate that this system operates without loss of performance and gains in flexibility in terms of compensation, particularly under condition of non-ideal voltage.
1.3 Problem Statement and Objectives

Distorted and asymmetrical grid-voltage conditions impose challenges for power converters operation in terms of undesired current circulation. Especially when control is taken in $d$-$q$ reference frame, direct and quadrature quantities present oscillating components due to voltage and current disturbances and separate them is not an easy task. Typically, low-pass filters are applied to mitigate those oscillating components which characterize loss of information. If someone wants to deal with those oscillating components then more complex controllers, like tunes controllers, must be applied to, instead of typically used proportional-integral controllers. By doing so, this control scheme approached to the $abc$-frame implementation, in which this latter avoids using the frame transformation.

In the light of that issue, this work proposes to study and analyze a permanent magnet synchronous generator (PMSG) based wind power system devised in $abc$-reference-frame ($abc$RF), acting as active power generator and, simultaneously, as active power filter under distorted and asymmetrical PCC voltage and load conditions, as well as three-phase voltage sags.

1.4 The Studied System

Figure 1.3 represents the considered system for study with all possible measurements. The grid-side control has, as main objective, to guarantee active power supplying with high power factor and low current or voltage THD for both grid-connected or island operation, respectively. It must meet the minimum power quality requirements for power converters in these operation modes.

Besides the active power delivering, additional functionalities may be included, as compensate selectively (XAVIER, CUPERTINO, et al., 2016) or fully load harmonic content and even provide voltage support (BRANDÃO, GUILLARDI, et al., 2016).

At the generator-side, the power converter controls the generator speed. It is performed so that the generator makes power conversion with low losses and takes advantage of the maximum power available in the wind. The DC-bus or the machine’s stator values can be used for power feedback control. Also the mechanical speed can be used for feedback control, being measured for an encoder or a sensorless estimation algorithm (XU, WANG, et al., 2018). Such power and control details will be given in the following chapters.
1.5 **Organization**

Firstly, Chapter 2 presents the mathematical model of a wind turbine related to mechanical energy conversion and MPPT issues. Further, a brief review of some electrical generators and power converter topologies for WECS is presented and the choice of that used in this work is then justified. Furthermore, it emphasizes the MPPT approach and the machine used in this study, despite of some variations found in literature (HUI, BAKHSHAI e JAIN, 2016), (WEI, ZHANG, et al., 2016).

The third Chapter treats the control system approach. It starts presenting the grid-side (GS) modeling, for both output inverter currents and DC-bus voltage. It highlights the current regulation strategy and the discussion about the current reference generation regarding three-leg three-wire (TLTW) voltage source inverter (VSI) constraints. Further, the PMSG control scheme developed in this work is presented and discussed.

In the Chapter 4, a computational model was implemented in Matlab/Simulink representing the complete dynamics of a wind energy conversion system based on PMSG with power quality improvement capability. It should be noted that all parameters considering in the control equations and system simulations are the same of the used test bench.

Chapter 5 validates the stator currents regulation in abcRF simulated before and presents the test bench used in this work, followed by conclusions in the further chapter.
CHAPTER 2

Wind Power System

2.1 Introduction

This chapter is aimed at presenting the main topics related to WECS that are relevant to this work. Some mathematical expressions that model the approximate behavior of a wind turbine, concerning power transferring, are presented here. From these equations, a maximum power point (MPP) curve can become estimated. It may be applied in the control of the generator to perform the electro-mechanical energy conversion with maximum efficiency. So, a basic maximum power point tracking (MPPT) method is presented in this Chapter.

Another relevant subject is related to the state-of-the-art topologies used in WECS. Further, some power converters used in PMSG based wind power system are overviewed. Finally, the considered system is presented in detail.

2.2 Wind-Turbine Model

2.2.1 Behaviour of the Mechanical Power

The mechanical power available in the wind [W] can be represented by (2.1) (HEIER, 2014). Where, \( \rho \) [kg.m\(^{-3}\)] is the air density, \( R_t \) [m] is the wind turbine’s radius and \( v_{wind} \) [m.s\(^{-1}\)] is the wind speed.

\[
P_{wind} = \frac{1}{2} \rho \pi R_t^2 v_{wind}^3
\]  

(2.1)

The power transferred to the shaft of the wind-turbine is weighted by the power coefficient \( C_p \), that is dependent on the turbine’s construction and dictates the efficiency of the power transfer, presented in (2.2).

\[
P_{shaft} = P_{wind} C_p(\lambda, \beta)
\]  

(2.2)
See $C_p$ equation in (2.3), where the factors $c_1$ to $c_6$ are constants whose values are, respectively, 0.5176, 116, 0.4, 5, 21 and 0.0068. $\beta[^\circ]$ is the blade pitch angle (SANTOS, 2015).

$$C_p(\lambda, \beta) = c_1 \left( \frac{c_2}{\lambda_i} - c_3 \beta - c_4 \right) e^{-c_5 \lambda_i} + c_6 \lambda$$  \hspace{1cm} (2.3)

The term $\lambda$ is the blade tip speed ratio due to the wind speed in a time instant, explicit in (2.4), where $\omega$ [rad.s$^{-1}$] is the mechanical angular speed of the wind turbine’s axis.

$$\lambda = \frac{\omega R_t}{v_{wind}}$$  \hspace{1cm} (2.4)

Considering a gear relation, $k_{gr}$ between the turbine’s and the generator’s shafts, the wind speed can be represented as,

$$v_{wind} = \frac{\omega_m R_t}{k_{gr} \lambda}$$  \hspace{1cm} (2.5)

where $\omega_m$ is the mechanical speed at the electrical machine axis. The $\lambda_i$ is dependent on the $\lambda$ and $\beta$ factors.

$$\frac{1}{\lambda_i} = \frac{1}{\lambda + 0.08 \beta - 0.035 \beta^3 + 1}$$  \hspace{1cm} (2.6)

### 2.2.2 Mechanical Dynamics

The generator shaft dynamics is represented by (2.7) (SINGH e SANTOSO),

$$J_{tg} \frac{d\omega_m}{dt} = T_m - B \omega_m - T_e$$  \hspace{1cm} (2.7)

where $J_{tg}$ is the total moment of inertia of the rotating mass set turbine-generator, $B$ is the viscous damping coefficient, $T_m$ is the mechanical torque at the generator shaft and $T_e$ represents the electromagnetic torque.
2.3 The Maximum Power Point Tracking

The maximum mechanical power transfer from the wind to the turbine’s shaft, obviously occurs when $C_p$ has its maximum value. It occurs when, for a given blade pitch angle, the wind and tip speed give a relation such that the mentioned coefficient is maximum (HEIER, 2014).

The graphic of (2.3) is represented in Figure 2.1(a). The aforementioned power transferring has its maximum efficiency when $\beta=0^\circ$ and $\lambda = 8.129$ (i.e. $\lambda_{max,o}$), with $C_p=0.48$ (i.e. $C_{p,max,o}$). Figure 2.1(b) shows the same curve for $\beta=0^\circ$. Note that there is a local maximum point ($\lambda_{max}, C_{p,max}$) for each pitch angle value. One can replace $\lambda_{max}$ in (2.3) and (2.5), then replace the resulting equation for $v_{wind}$ in (2.1). Thus, for a fixed $\beta$ value, an expression for the maximum power point tracking (MPPT) of the generator shows up in (2.8):

$$P_{m,max} = \frac{1}{2}\rho \pi R_t^2 \left(\frac{1}{\lambda_{max} \cdot k_{gr}}\right)^3 \cdot C_{p,max} \omega_m^3$$ (2.8)

So, $P_{m,max}$ can be represented in terms of constant $K_{max}$, as in (2.9).

$$P_{m,max} = K_{max} \omega_m^3$$ (2.9)

Note that the generator power is function of its rotor speed and can be plotted as a family of curves, each one due to a specific wind speed, as shown in Figure 2.2, where $v_1 < v_2 < v_3 < v_4 < v_5$. 

![Figure 2.1 - Power coefficient ($C_p$) x Tip Speed Ratio ($\lambda$) x Pitch Angle ($\beta$) (a); Power coefficient ($C_p$) x Tip Speed Ratio ($\lambda$), with $\beta = 0^\circ$ (b).](image-url)
The plot of (2.9) over these curves shows that it is on the maximum power point (MPP) of each one of them.

The tracking of the MPP is also shown on Figure 2.2 for a specific wind speed (refer to \( v_4 \)). For mechanical power feedback, if in a first time the sensing power is \( P_1 \), thus, the reference speed for the generator drive will be \( \omega_2 \), so the mechanical speed increases and the generated power is increased to \( P_2 \) (SANTOS, CUPERTINO, et al., 2015). And so on it occurs since the MPP around \((\omega_4, P_4)\) is reached. Note that this tracking can also occur in the same way for a mechanical speed feedback (SINGH, KHADKIKAR e CHANDRA, 2011).

![Figure 2.2 - Mechanical speed versus generated power for several wind speeds: \( v_1 < v_2 < v_3 < v_4 < v_5 \); behavior of the MPPT scheme.](image)

2.4 **Wind Power Conversion System Topologies**

A modern wind electromechanical system is composed of some fundamental parts, as shown in Figure 2.3. From the mechanical perspective, the turbine itself has blades that absorb the mechanical energy from an air mass displacement. This makes the turbine’s shaft to spin and then activates a gear train that makes a mechanical transmission from a lower to a higher speed. In this sense, the high speed shaft makes an electrical generator to work and convert mechanical into electrical energy. Note that it may be either synchronous or asynchronous, depending on the system. Besides, it is important mentioning that the gear train is an optional item (YARAMASU, DEKKA, et al., 2017), (SINGH e SANTOSO). Generally, a PMSG used in wind systems has a number of poles which are sufficiently high to provide an adequate
electrical angle variation. So, even if the mechanical temporal angle variation is slow, a direct speed transmission may be used in this case (YARAMASU, WU, et al., 2015).

Schematically, the Figure 2.3 also represents the two most common wind power conversion topologies, the DFIG and the PMSG based, detailed in the next subsections. The first consists in a wound-rotor doubly-fed electric asynchronous three-phase generator, in which the stator is directly connected to the grid (orange arrow, at the top), while the rotor is connected through a power converter in series (green arrows way, at the bottom). Those connections are usually made through a transformer in order to either provide galvanic isolation between the grid and the conversion system or also to match properly the distribution voltage level of stator and rotor (SINGH e SANTOSO). For the PMSG power system, a permanent magnet synchronous three-phase generator is directly connected to the grid through a back-to-back power converter.

![Figure 2.3 - Wind turbine diagram (adapted from (SINGH e SANTOSO)).](image)

### 2.4.1 Wind Turbine with Partial-Scale Frequency Converter

In a partial scale WECS, the machine is a doubly fed induction generator (DFIG), whose connection to the grid is made in a double way. Its stator is directly connected to the mains while its rotor has such connection made through a power converter. This indirect link enables control the rotor speed.

The power rating of this partial scale power converter defines the range of speed control, which is typically ±30% around synchronous speed. As shown in Figure 2.4, that converter is composed of a rectifier stage followed by a DC-bus and an inverter stage. So, compared to the fixed speed conversion system, and partial-fixed speed based topologies, it can be pointed out
some advantages as: an extended speed range, the improvement of the power conversion efficiency by enabling the use of a MPPT algorithm, smooth grid interconnection and better dynamic response toward grid disturbances. The advances of this configuration have made it one of the most used technologies in WECS industry nowadays (BLAABJERG, LISERRE e MA, 2011), (YARAMASU, WU, et al., 2015).

Despite of those aforementioned advances, the DFIG configuration has some drawbacks that prevent its use in offshore system, such as: a limited fault ride through capability because of the partial scale controllability of the power converter and the need of regular maintenance because of the gearbox presence and the slip rings (average-life of 6 to 12 months) that connect the converter to the rotor (YARAMASU, WU, et al., 2015).

2.4.2 Wind Turbine with Full-Scale Power Converter

An example of full-scale configuration is shown in Figure 2.5. It is consisted of a generator that can be a PMSG, a WRSG, or even a SCIG, being this first the most suitable because of the absence of slip rings. The electrical generator is connected to the grid through a power converter so that it is decoupled from the power system. Note that it consists of a rectifier stage capable of controlling the generator’s speed within full range, a DC-bus and an inverter stage. With a 100% speed range operation, this topology achieves better performance on FRT compliance with robustness against grid faults, compared to the other configurations.

Aftermore, the gear box is an optional part because a high-pole number synchronous generator can be used instead of an induction one. As partial-scale configuration, full-scale one can perform MPPT, but with higher wind energy conversion efficiency. On the other hand, the power converter must be rated as the generator’s power rating, which increases its size, complexity, losses and cost (YARAMASU, WU, et al., 2015).
2.5 Power Converters for PMSG Based Wind Power Systems

For low power and voltage wind turbines, a cost-effective configuration could be that the rectifier is consisted of a three-phase diode bridge and a DC-bus followed by a three-phase two-level voltage source inverter (BLAABJERG, LISERRE e MA, 2011), as shown in Figure 2.6. But the power factor at the output of the PMSG is low, what brings low efficiency at the machine’s side. As an alternative to mitigate this problem, a z-source inverter instead of a voltage one was used in (DU e BHAT, 2016) and the system efficiency is improved.

The DC-bus voltage is variable in such configuration and must remain always great enough to guarantee active power flowing towards the grid, which limits substantially the speed range operation. A solution to this issue, illustrated in Figure 2.7, is to use a two-stage rectifier by inserting a DC-DC step-up chopper between the diode bridge output and the DC-bus capacitor. Thus, its voltage is regulated to be always high enough to provide active power flow from the generator to the grid, so allowing variable speed operation with stable DC-bus voltage. Furthermore, the total harmonic distortion (THD) of stator phase currents are reduced and a better power factor at the generator’s output (DU e BHAT, 2016), (BLAABJERG, LISERRE e MA, 2011). Another advantage of this topology is to provide decoupling from generator side to the grid side, despite of transients in the PMSG (YARAMASU, WU, et al., 2015). The decoupling performance is related to the size of DC-capacitor.

In this way, another configuration with full-scale power conversion is presented in Figure 2.8. Its difference from the previous examples is that it consists of a two-voltage-level fully controlled VSR at the generator’s output, in a back-to-back structure.
This topology is widely used in PMSG based WECS and has been proven itself to be robust and reliable, due to its simplicity and few components. Moreover, it also presents a proper decoupling with respect to voltage transients at the point of common coupling (PCC).

On the other hand, the aforementioned solutions might become unsuitable for wind turbines depending on its rated voltage and power. For a range from medium to high voltages, some drawbacks are noted. The two level modulated voltages at the rectifier’s input will present high gradient \( \frac{dv}{dt} \), what causes stresses to the electrical machines (generator and isolating transformer) and high current THD. This leads to the use of relatively bulky passive
filters at both machine’s and generator’s sides. Another issue is that available power transistors should be mounted in parallel and/or series to achieve the required nominal quantities.

This latter problem introduces the field of the multilevel power converters for back-to-back applications in variable-speed WECS. They are often used in rated powers from 3MW to 7MW (due to cost-effectiveness) and can be applied through some different structures. In fact, those converters are capable of introducing more than two voltage levels and thus allow the reduction of filter sizes. On the other hand, multilevel converters may bring some reliability and power density problems for the use of higher number of switching devices. However, these issues open the field of the multiple-cells approach for both multilevel and standard power converters for big size WECS (BLAABJERG, LISERRE e MA, 2011).

2.6 The Considered PMSG Based Distributed Generator

The arguments presented in sections 2.3 and 2.4 reinforce the cost effectiveness of the back-to-back VSC based topology to be used in full-scale low and medium voltage wind systems. So, the PMSG based WECS is chosen in this work. Figure 2.9. shows the considered DG under study with non-linear load. This system mainly consists of the following components: wind-turbine, PMSG, a back-to-back two-level converter connected to the mains through an inductive filter and non-linear loads connected to the PCC. Later the unwanted load current terms will be compensated in order to give support to the grid, assisting to improve the power quality at the PCC.

The DC-bus is composed of a capacitor that makes the DC-voltage filtering, which equivalent capacitance is represented by $C_{dc}$. At the grid-side, the three inductors, of inductance represented by $L_f$, perform the filtering of the output current of the inverter to attenuate the switching ripple. The nominal line voltage and frequency at the PCC are 220V and 60Hz, respectively.

The generator-side and grid-side control blocks shown in Figure 2.9 are detailed and explained in Figures 2.10 and 2.11, respectively. The generator-side control block is composed of a MPPT, a shaft speed estimation algorithm and stator current and mechanical speed control loops. Referring to Figure 2.10, it is valid the following description: the measured and conditioned DC-bus voltage $\bar{v}_{dc}$ is multiplied by the measured and filtered DC-current $\bar{i}_{dc}$ to obtain the instantaneous DC power $\bar{P}_{dc}$, which in turn is applied to the equation presented in (2.9). Then, the instantaneous value of the mechanical angular speed reference ($\omega_m^*$) is obtained
by the MPPT algorithm. In this control loop, the mechanical speed of the generator’s shaft may be measured by an encoder. However, in this work, a sensorless algorithm is used instead. So, the estimated value of $\hat{\omega}_m$ is generated from this algorithm, using the stator current ($i_{abc}^s$) and the terminal phase-voltage ($v_{abc}^t$) measured values. The output of the speed control loop gives a reference value, $i_{abc}^{*s}$, for the stator current loop, which in turn generates an output signal that is the input of the PWM modulator. This latter builds the PWM control voltages for the IGBTs of the generator-side converter, represented as $V_{PWM}$.

A DC-bus voltage and an output filter current control loop, as well as a load current decomposition scheme, comprise the grid-side control block. Referring to Figure 2.11, the output of the DC-bus voltage control loop gives a current reference which is related to the active power that must be delivered at the output of the inverter. This reference is summed with another reference that is given by a load current decomposition block, which generates a compensation current reference. This block, in turn, uses measured values of the PCC phase voltages and load currents. The output of the current control loop gives the input signal to generate the PWM control voltages, $V_{PWM}$. The PMSG control system with the elements of the control blocks are deeply explained in Chapter 3.

2.7 Final Considerations

It was presented in this Chapter some important aspects concerning WECS, which have given the necessary understanding about wind power in order to develop this work. It has shown the variable characteristic of the power coefficient ($C_p$) and because of that, the wind turbine should be controlled so that this coefficient is held up in its maximum value. The way to perform this is to control the generator accordingly the equation presented in (2.9). Thus, it can be concluded that the power conditioner must be able to control the power or mechanical speed of the machine in a range which is wide enough so that energy conversion is performed with maximum efficiency.

The most important wind power system configurations (partial-scale and full-scale) were presented. As well as the main power converter topologies for a PMSG based WECS in the context of this work. From this knowledge found in literature, the feasibility of the chosen DG configuration in controlling energy conversion holding maximum possible efficiency is then justified for the power level considered in this work.
Figure 2.9 – Considered PMSG based DG.

Figure 2.10 – Control blocks of the considered system: generator-side.

Figure 2.11 – Control blocks of the considered system: grid-side.
CHAPTER 3

The Control System

3.1 Introduction

In this chapter, elements of the control system are presented, showing the abc-frame control scheme and dynamic modeling for grid and generator-side. Furthermore, the issues related to coupling among phase quantities are analyzed, such as: inverter’s modulated voltages; and possible solutions for decoupling them are presented. Then, the considered PWM structure is described. Another relevant discussion regarding the proposed control system is related to the decoupling between torque and flux for control the stator currents of the PMSG, which is also analyzed herein. Finally, the design of the control loops is described.

3.2 abc-Frame Control Structure

3.2.1 Grid-Side

3.2.1.1 Current Reference Generation

As mentioned before, the power converters are used in distributed generation to inject active power into the grid, however, other capabilities can be added concomitantly, as for example active filtering. In short, it is achieved by adding proper reference current to the grid-side control loop (MONTERO, CADAVAL e GONZALEZ, 2007). More specifically, the measured load current can be decomposed in some current terms related to its active and reactive power consumption, asymmetry and harmonic content (BRANDÃO, GUILLARDI, et al., 2016), (YAN, WANG, et al., 2018), (MALENGRET e GAUNT, 2008), (LU, XU e LI, 2011).

The CPT is used in this work to perform the load decomposition, as it deals directly with active and non-active current terms in the abcRF. In this approach, load current of each phase are the instantaneous sum of its active, reactive and void current terms. Such quantities are
represented as vectors in the abcRF at the right side of (3.1) (PAREDES, 2011). The selectivity of such power theory is higher than the presented in this work as reactive and void terms can be further decomposed in more components, but it is out of scope of this work.

\[ \vec{i} = \vec{i}_a + \vec{i}_r + \vec{i}_v \]  

(3.1)

Accordingly to such power theory, the active, and reactive currents can be split into balanced and unbalanced active current terms, as follows:

\[ \vec{i}_a = \vec{i}_a^b + \vec{i}_a^u \]
\[ \vec{i}_r = \vec{i}_r^b + \vec{i}_r^u \]  

(3.2)

And for this latter, the non-active current can be isolated, as shown in (3.3), in order to be compensated, i.e., supplied by the power conditioner.

\[ \vec{i}_{na} = \vec{i} - \vec{i}_a = \vec{i}_a^u + \vec{i}_r^b + \vec{i}_r^u + \vec{i}_v \]  

(3.3)

The balanced active current vector is defined in (TENTI, PAREDES e MATAVELLI, 2011) as,

\[ \vec{i}_a^b = \frac{\langle \vec{v}_{pcc}, \vec{i} \rangle}{\| \vec{v}_{pcc} \|^2} \cdot \vec{v}_{pcc} \]  

(3.4)

where \( \vec{v}_{pcc} \) is the vector of PCC phase voltages. The numerator indicates an inner product of it with load currents. The denominator indicates the norm of PCC voltages vector squared. This latter can be rewritten as,

\[ \vec{i}_a^b = \frac{P}{V_{pcc}^2} \cdot \vec{v}_{pcc} \]  

(3.5)

such that \( P \), the active power absorbed at the port where the measurement is made (PCC voltages and load currents in this case, as it is in Figure 3.1) and \( V_{pcc} \), the norm of considered PCC voltages vector. So, an equivalent balanced conductance, that is \( G_b = \frac{P}{V_{pcc}^2} \), is defined. Thus, balanced active currents are equivalent to the currents drawn by a symmetrical resistive load with same active power that is consumed by the actual load.

For additional functionality, PCC voltage and load current measurements are used to calculate the non-active component to be supplied to the load, set as additional reference for the
current controller. It is worth mentioning that not only $\vec{i}_{na}$, that is the sum of all compensation components, may be set as reference for compensation, but also other possible combinations of the current decomposition such as: $\vec{i}_v, \vec{i}_r, \vec{i}_u = \vec{i}_a + \vec{i}_b + \vec{i}_c$.

Extra reactive current, generated by some reactive power reference or voltage support, can be also added, represented by $\vec{i}_R$ as shown in Figure 3.2. The reference $\vec{i}_{fa}$ is the output of the DC-bus voltage controller, responsible for DC-bus voltage regulation through active power balance between AC and DC sides.

Figure 3.1 – Reference current generation for load compensation.

3.2.1.2 Line-Current Dynamics

A grid-connected VSI can be dynamically modeled considering a schematic as shown in (HOLMES, LIPO, et al., 2009), (RYAN, LORENZ e DONCKER, 1999). The inverter is connected to the grid through an inductive filter, whose the flowing currents are regulated by the output voltages that are driven by the PWM switching-signals.

By Kirchhoff’s voltage law (3.6) the circuit of Figure 3.1 can be dynamically modeled considering only the fundamental value of each phase-voltage. It is worth mentioning that “x” and “X” are denoted herein as the subscripts that indicates the phases “a,b,c” and “A,B,C” as represented in Figure 3.1, respectively. The inverter line-currents are denoted by $i_{f,x}$. 
\[ v_{X,N} = R_f i_{f,X} + L_f \frac{di_{f,X}}{dt} + v_{X,N} \mid v_{*,N} = v_z \]  \hspace{1cm} (3.6)

Then, considering the single-phase representation, it is possible to build three independent voltages, \( v_{X,N} \), using the Laplace transform and considering the grid voltage, \( v_{X,N}(s) \) as a short-circuit. The per-phase control transfer function is represented as:

\[ G_{i_f}(s) = \frac{i_{f,X}(s)}{v_{X,N}(s)} = \frac{1}{R_f + sL_f} \]  \hspace{1cm} (3.7)

### 3.2.1.3 DC-bus Voltage Dynamics

Considering the active power balance between the grid and DC-bus sides, the expression in (3.8) models the variations of DC voltage related to peak line current, taking into account symmetric grid quantities. Appendix A gives detail about this modeling.

\[ G_{v_{ac}}(s) = \frac{\tilde{v}_{dc}(s)}{i_{f,a}^k(s)} = -\frac{3V_{ac}}{\sqrt{2V_{dc}}} \frac{1}{SC} \]  \hspace{1cm} (3.8)

As illustrated in Figure 3.3, the necessary variation of the DC current at DC-side is dictated by an active-current peak value variation through a gain that contains the nominal values of the DC-bus voltage, which is \( V_{dc} \), and that of the grid-phase-voltage \( rms \) value, that is \( V_{ac} \). It also illustrates perturbations that compose the DC-side current and consequently may be source of perturbations in DC-bus voltage. A considered perturbation may be variations of the source current as consequence of variations in the instantaneous generated power.
There are other sources of oscillation in the DC-bus voltage, represented by the “general” oscillation term, and are reflections of perturbations at the grid-side. Harmonic currents injected by the VSI used by compensation, bring a component of oscillation in DC-current. Moreover, asymmetry in the grid-voltages generates an oscillating component in the instantaneous power balance, which is reflected as an additional term of perturbation in the DC-current, and so on in the DC-bus voltage.

3.2.2 Generator-Side

To model the PMSG electric behavior, the circuit in wye connection presented in Figure 3.4 is considered, from which the stator’s electric equations are developed. The stator currents, $i_{s,X}$, and flux linkages, $\varphi_{r,X}$, are considered to be sinusoidal and symmetric.

As the machine used in this work has salient poles, the self and mutual inductances vary sinusoidally with the rotor position. In literature, that variation is modeled as the double of the instantaneous electric angle (KRAUSE, WASYNCZUK e SUDHOFF, 2002), (BOLDEA, 2006), (BOSE, 2002), (AHMED-ZAID, CHAUDHRY e DEMERDASH, 1995).

The behavior of the stator phase self-inductances is represented as the following equations, from (3.9) to (3.11) (KRAUSE, WASYNCZUK e SUDHOFF, 2002), (KOCHE, 2015).
\[ L_A = L_s - L_m \cos(2\theta_e) \quad (3.9) \]
\[ L_B = L_s - L_m \cos\left(2\theta_e - \frac{4\pi}{3}\right) \quad (3.10) \]
\[ L_C = L_s - L_m \cos\left(2\theta_e + \frac{4\pi}{3}\right) \quad (3.11) \]

And the mutual inductance expressions are shown in (3.12) to (3.14).
\[ M_{AB} = L_s - L_m \cos\left(2\theta_e - \frac{2\pi}{3}\right) \quad (3.12) \]
\[ M_{BC} = L_s - L_m \cos(2\theta_e + 2\pi) \quad (3.13) \]
\[ M_{CA} = L_s - L_m \cos\left(2\theta_e + \frac{2\pi}{3}\right) \quad (3.14) \]

The terminal phase voltage equations are obtained by the Kirchhoff’s voltage law in (3.15),
\[
\begin{bmatrix}
    v_{t_{AN}} \\
    v_{t_{BN}} \\
    v_{t_{CN}}
\end{bmatrix} = -
\begin{bmatrix}
    R_s & 0 & 0 \\
    0 & R_s & 0 \\
    0 & 0 & R_s
\end{bmatrix}
\begin{bmatrix}
    i_{s,A} \\
    i_{s,B} \\
    i_{s,C}
\end{bmatrix} + \frac{d}{dt}
\begin{bmatrix}
    \varphi_{s,A} \\
    \varphi_{s,B} \\
    \varphi_{s,C}
\end{bmatrix} \quad (3.15)
\]

Its matrix equation is presented in (3.16), where \( v_{tn} \) is the 3x1 matrix of the terminal phase voltages. The term \( R_s \) is the matrix of the stator phase windings series resistance and \( i_s \) is that which represents the three phase stator currents.
\[
v_{tn} = -R_s i_s + \frac{d\varphi_s}{dt} \quad (3.16)
\]

The stator’s flux linkage equations are expressed by the \( \varphi_s \) 3x1 matrix and presented in (3.17).
\[
\begin{bmatrix}
    \varphi_{s,A} \\
    \varphi_{s,B} \\
    \varphi_{s,C}
\end{bmatrix} = -
\begin{bmatrix}
    L_A & M_{AB} & M_{AC} \\
    M_{AB} & L_B & M_{BC} \\
    M_{AC} & M_{BC} & L_C
\end{bmatrix}
\begin{bmatrix}
    i_{s,A} \\
    i_{s,B} \\
    i_{s,C}
\end{bmatrix} -
\begin{bmatrix}
    \varphi_{r,A} \\
    \varphi_{r,B} \\
    \varphi_{r,C}
\end{bmatrix} \quad (3.17)
\]

The aforementioned equation is depicted and (3.18) in a concise form, where \( L_s \) is the stator inductances matrix and \( \varphi_r \) is the three-phase flux linkage established between the rotor permanent magnet and stator phase windings.
\[
\varphi_s = -L_s i_s - \varphi_r \quad (3.18)
\]
3.3 abc-Frame Control Issues

3.3.1 Coupling Between Control Variables

The inverter “neutral” point (the DC-bus center point (HOLMES, LIPO, et al., 2009)) is named here, “o”. As done in (MATAKAS JÚNIOR, 2012), herein the voltage components whose instantaneous sum is zero are also called “balanced” terms, otherwise, they are called “unbalanced” terms. So, positive and negative sequence components are defined herein as balanced terms and zero sequence as unbalanced terms. In this way, if the phase-voltages have only positive and negative components, it is balanced. If it has some zero sequence component, then it is unbalanced. Furthermore, it is worth mentioning that the presented voltages in the following discussion are considered to have only fundamental frequency order, not taking into account switching harmonics.

Representing the zero-sequence component of grid voltages have separately from their balanced components. So that,

\[ v_z = v_{*,N} = \frac{1}{3} (v_{a,N} + v_{b,N} + v_{c,N}) \]  

(3.19)

and, the balanced components,

\[ v_{a,*} + v_{b,*} + v_{c,*} = 0 \]  

(3.20)

So, \( v_{X,o} \) are the VSI modulated voltages. Even if they are intended to be balanced, their switching characteristic brings an intrinsic additional component of common mode (zero sequence) (JULIAN, ORITI e LIPO, 1999), (KRUG, KUME e SWAMY, 2004). It can be confirmed by simulation in Figure 3.5, where sinusoidal balanced three-phase signals with a triangular carrier (CPWM) are imposed to control the switches of a TLTW VSI. At the top, the terminal voltages are presented normalized by 100. At the bottom, the common mode voltage, \( (v_{A,o} + v_{B,o} + v_{C,o})/3 \), is divided by the same factor and its temporal mean value is presented in original dimension.

The equivalent circuit of Figure 3.2 is shown in Figure 3.6 (a) and (b), as its three-phase and per-phase representations, respectively. Referring to it, the desired modulated voltages are represented by \( v_{X,M} \) and its additional “undesired” common mode component promoted by the switching regime is \( v_{M,o} \). Thus, if balanced modulating signals are imposed,

\[ \frac{v_{A,M} + v_{B,M} + v_{C,M}}{3} = 0 \]  

(3.21)
Thus,

$$\frac{v_{A,o} + v_{B,o} + v_{C,o}}{3} = v_{M,o} \quad (3.22)$$

By the Kirchhoff’s voltage law, the considering per-phase dynamic linear equations are shown in (3.23).

$$v_{A,N} = v_{A,M} + v_{M,o} - v_{N,o} = R_f i_{f,A} + L_f \frac{di_{f,c}}{dt} + v_{a,*} + v_{*,N}$$

$$v_{B,N} = v_{B,M} + v_{M,o} - v_{N,o} = R_f i_{f,B} + L_f \frac{di_{f,c}}{dt} + v_{b,*} + v_{*,N} \quad (3.23)$$

$$v_{C,N} = v_{C,M} + v_{M,o} - v_{N,o} = R_f i_{f,C} + L_f \frac{di_{f,c}}{dt} + v_{c,*} + v_{*,N}$$

Note that $v_{X,M} + v_{M,o} = v_{X,o}$ and $v_{X,*} + v_{*,N} = v_{X,N}$. As the sum of the phase-currents is null ($i_{f,A} + i_{f,B} + i_{f,C} = 0$), by summing the three phase-equations of (3.23), important conclusions are done (HOLMES, LIPO, et al., 2009):

![Figure 3.5 – Modulated voltages and their common mode component due to switching behavior.](image)

![Figure 3.6 – Equivalent circuit of the grid-connected VSI in question (a); Equivalent per-phase (b).](image)
1. **If the commanded and grid voltages are unbalanced:**
   
   In this case, the relation of (3.24) is different from zero, and not only the common mode component $v_{M,o}$ appears in the VSI output voltages. Thus, the following relation remains,
   \[
   \frac{1}{3} (v_{A,M} + v_{B,M} + v_{C,M}) + v_{M,o} - v_{N,o} = v_{*,N} \tag{3.24}
   \]
   
   - Note that, substituting (3.24) in (3.23), a strong coupling is noted between the commanded voltages.

2. **The commanded voltages are unbalanced and the grid-voltages are balanced:**
   
   Observe that $v_{*,N} = 0$ and,
   \[
   -\frac{1}{3} (v_{A,o} + v_{B,o} + v_{C,o}) = v_{M,o} - v_{N,o} \tag{3.25}
   \]
   
   - Substituting (3.25) in (3.23), a strong coupling is noted between the commanded voltages again.

3. **The commanded voltages are balanced and the grid-voltages are unbalanced:**
   
   In this case, the relation of (3.21) is valid and,
   \[
   v_{M,o} - v_{N,o} = v_{*,N} \tag{3.26}
   \]
   
   - And substituting (3.26) in (3.23), the following phase-equation is given:
   \[
   v_{X,N} = v_{X,M} = R_f i_{f,X} + L_f \frac{df_x}{dt} + v_{x,*} \tag{3.27}
   \]
   
   In this case, it is noted that the unbalanced term contained in both grid and VSI output switched voltages (related to the DC-bus neutral point “o”) do not affect the current control dynamics. Being $v_{X,M}$, the input signal to current control. Furthermore, no coupling is noted between phase commanded voltages.

4. **The commanded and grid-voltages are balanced:**
   
   Also $v_{*,N} = 0$ and,
   \[
   v_{M,o} - v_{N,o} = 0 \tag{3.28}
   \]
   
   - Herein, substituting (3.28) in (3.23), the phase-equation in (3.27) still remains.
   
   Once described the aforementioned situations, it is concluded that, if in some way it is possible to guarantee that the modulated voltages are balanced, it is feasible to use three
independent current regulators with stable control, even if some zero-sequence component is present in grid voltages. Otherwise, as the cases 1 and 2, such coupling causes instability.

In this way, by Laplace transform, the per-phase control equation can be represented as,

\[
G_i(s) = \frac{i_{fX}(s)}{v_{xM}(s)} = \frac{1}{R_f + sL_f}
\]

while by superposition, \(v_{x^*}(s)\) is made a perturbation term.

### 3.3.2 Strategies for Control Variables Decoupling

#### 3.3.2.1 Line Current Control with Two Feedback Loops

A solution for decoupling the control variables that is widely described in literature is shown in (MATAKAS JÚNIOR, 2012) and (HAYASHI e MATAKAS, 2017). It consists in implementing only two control loops. Referring to Figure 3.7, the modulating signals for phases “a” and “b” are \(m_a\) and \(m_b\), respectively. The third control signal, phase “c”, is made the negative sum of the others (- \(m_a - m_b\)). Therefore, some zero-sequence component in the modulation signal is avoided, as its final sum is always zero. So, this approach puts the use of the stationary abc-frame in some advantage compared to frame-transformation based control for three-leg three-wire system, as it saves computational power by avoiding frame transformation and using two controllers.

![Figure 3.7 – Control in abcRF with two controllers.](image)

#### 3.3.2.2 Control Variable Measurement for Current Control with Three Independent Feedback Loops

In (MATAKAS JÚNIOR, 2012), it is also shown that there is a solution in order to use three independent control loops. It is known that for stable current control, the sum of the measured line-currents must be null complying with Kirchhoff’s current law \((i_A + i_B + i_C = 0)\). This issue is quite irrelevant in a first moment. It is a three-wire system and obviously such
measurement does not bring any zero sequence components. But it is worth pointing out that there is usually quite little difference in the current sensor offsets and gains that may generate some virtual zero sequence measurement. Consequently, it would generate a zero-sequence component in the error signal. As this system is not able to generate unbalanced currents, in an attempt to reduce the error, the converter would lose stability. An efficient solution is to subtract the possible instantaneous unbalanced term (3.30) and (3.31) from each phase measurement. Where “k” is the k-th interaction, within digital control context.

\[ i_{f,z}(k) = \frac{1}{3} (i_{f,A}(k) + i_{f,B}(k) + i_{f,C}(k)) \]  \hspace{1cm} (3.30)

\[ i_{f,X}(k) = i_{f,X}(k) - i_{f,z}(k) \]  \hspace{1cm} (3.31)

Another possible solution is to measure two phases, and the third making the negative sum of the others.

### 3.3.3 Continuous Pulse Width Modulation with Optimal Zero Sequence Injection

Another relevant fact to be noted is that zero sequence injection to be summed to the modulating signals does not affect the low frequency behavior (HAYASHI e MATAKAS, 2017) and it can be used to take advantage in the CPWM by adding an “optimal zero sequence signal”, shown in (3.32).

\[ v_{z,\text{optm}}(k) = \frac{-[\max\{v_a^*(k), v_b^*(k), v_c^*(k)\} + \min\{v_a^*(k), v_b^*(k), v_c^*(k)\}]}{2} \]  \hspace{1cm} (3.32)

It is was demonstrated in (HOLMES e LIPO, 2003) that such sum gives to the continuous PWM the same behavior of the so-called space vector modulation (SVPWM), with the same linear modulation range, grater then that of standard CPWM. Those aforementioned improvements that may be done in abcRF control are depicted in Figure 3.8.
3.4 **Torque Control**

In short, the generator speed control regulates the maximum power point of the wind system. So, the controlled rectifier (generator-side converter) regulates the active power delivery from the wind turbine. Direct speed control of synchronous machines is performed by directly controlling the electromagnetic torque. This one opposes the mechanical torque in order to vary acceleration and regulate the machine’s shaft speed. To achieve torque control, the following requirements must be maintained every time:

1. constant rotor field flux (as it is for PMSG);
2. armature current being independently controlled;
3. an orthogonal spacial angle orientation between the armature magnetomotive force and the pole field flux (so that there is no interaction between both variables). The manner it is matched here is explained in the following subsection.

### 3.4.1 abc-Frame Considerations for PMSG Control

For a PMSG, according to generating’s convention (BOLDEA, 2006), Figure 3.9 represents a synchronous frame vector diagram with the armature voltage (blue), the permanent magnet field flux (green) and the stator current (red) decoupled into its direct and quadrature components.

\[
\begin{align*}
T_e &= \frac{3}{2} N_{pp} \left[ \frac{E_a f I_s}{\omega_e} + (L_d - L_q) I_s^2 \cos(\alpha) \sin(\alpha) \right] \\
\end{align*}
\]

(3.33)

The left hand term is related to the reaction torque and the right hand term is related to the so called reluctance torque. Note that if \( \alpha \) is regulated in 0° (motor convention) or 180°
(generator convention), the reluctance torque component is left to zero, and the torque magnitude is directly controlled through the stator current magnitude. So, a well-known technique to perform the speed control is split the stator current into direct \((i_d)\) and quadrature \((i_q)\) components. Thus, \(d\)-component current is regulated in zero, while \(q\)-component has a reference value that is necessary to give the appropriate electromagnetic torque. So, in steady state, the stator current has only the \(q\)-component in-phase with the armature voltage (LIPO, 1996). For speed regulation in a natural frame (abc) approach, the same field orientation can be achieved by synchronization of stator current with the armature induced voltage. It can be carried out through measuring the machine angular frequency through an encoder or by tracking the electric angular frequency using a sensorless estimation method, as it is developed in (SANTOS, CUPERTINO, et al., 2015) and (MERABET, TANVIR e BEDDEK, 2016).

### 3.4.2 Torque and Power Equations in abc-Frame

Assuming the magnetic system is linear, its co-energy is as follows (BOLDEA, 2006),

\[
W_m(i_s, \theta_e) = N_{pp} i_s^T \left( \frac{1}{2} L_s i_s + \varphi_r \right)
\]  
(3.34)

being \(N_{pp}\) the number of pole pairs in the machine’s rotor.

The instantaneous electromagnetic torque comes from (3.35), being its derivative with respect to the electric angle, which expresses the rotor position variation as:

\[
T_e = -\frac{\partial W_m(i_s, \theta_e)}{\partial \theta_e}
\]  
(3.35)

which gives (3.36),

\[
T_e = N_{pp} i_s^T \left( \frac{1}{2} \frac{dL_s}{d\theta_e} i_s + \frac{d\varphi_r}{d\theta_e} \right)
\]  
(3.36)

The previous equation can be expanded and represented as three separated components, \(T_1, T_2\) and \(T_3\), as follows.

\[
T_e = T_1 + T_2 + T_3
\]  
(3.37)

Each one is shown in (3.38), (3.39) and (3.40) (WEIZHE, 2004).

\[
T_1 = \frac{N_{pp}}{2} \left( i_{s_A}^2 \frac{dL_A}{d\theta_e} + i_{s_B}^2 \frac{dL_B}{d\theta_e} + i_{s_C}^2 \frac{dL_C}{d\theta_e} \right)
\]  
(3.38)
The terms $T_1$ and $T_2$ are resultant of the self and mutual inductances, respectively. They, summed, compound the reluctance torque, as shown in (3.41).

\[ T_1 + T_2 = T_{relt} \]  

(3.41)

The last torque component, named reaction torque, is due to the back-EMF and the stator phase currents and gives the major contribution.

\[ T_3 = T_{retn} \]  

(3.42)

Finally, the electromagnetic torque might be expressed by the sum of the reluctance and reaction components.

\[ T_e = T_{relt} + T_{retn} \]  

(3.43)

By trigonometric relations, it is possible to prove that, if stator currents are symmetric and in-phase with the back EMF (also symmetric), the terms of $T_1$ and $T_2$ become zero and the reaction torque dictates the electromagnetic one by,

\[ T_e = \frac{3}{2} N_{pp} \phi_{pm} I_s \]  

(3.44)

being $I_s$ the peak value of stator currents and $\phi_{pm}$ the peak value of the permanent magnet flux linkages.

### 3.5 Control Loops Project

All controllers along this section were designed based on a frequency method, in which at the desired cross-over angular frequency, $\omega_c$, the magnitude of the open loop transfer function is equal to one, and its phase is equal to the desired phase margin (PM).

\[ |C(\omega_c) \cdot G(\omega_c)| = 1 \]

\[-180^\circ + PM = \text{phase}\{C(\omega_c) \cdot G(\omega_c)\}\]  

(3.45)
Where $C(\omega_c)$ and $G(\omega_c)$ are the transfer functions of the controller and open-loop plant, respectively, at the desired cross-over frequency; $PM$ is the desired phase margin. This calculation is carried out in MATLAB, for rapid parameter update. Then, the calculated controllers were discretized with the Tustin method for implementation in both simulation and experimental results.

### 3.5.1 Grid-Side Control Loop

Figure 3.10 shows the grid-side control loop, which parameters are shown in Table 3.1. An inner closed-loop has a fast response and makes the regulation of the input filter current (i.e. output inverter current). The DC-bus voltage loop is closed externally to it, is lower and gives the necessary active current reference to the inner loop for active power balance. Details of both closed-loops are depicted further.

#### Table 3.1 - Grid-Side Control Circuit Parameters

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter inductance ($L_f$)</td>
<td>4 mH</td>
</tr>
<tr>
<td>Filter inductor series resistance ($R_f$)</td>
<td>0.16Ω</td>
</tr>
<tr>
<td>DC-Bus capacitors bank ($C$)</td>
<td>3.06 mF (equivalent)</td>
</tr>
<tr>
<td>DC-bus nominal voltage ($V_{dc}$)</td>
<td>500V</td>
</tr>
<tr>
<td>Nominal grid phase-voltage ($V_{ac}$)</td>
<td>127 V (rms)</td>
</tr>
<tr>
<td>Nominal PCC phase-voltage peak value ($V_{ac}^{pk}$)</td>
<td>180V</td>
</tr>
<tr>
<td>Grid nominal frequency</td>
<td>60 Hz</td>
</tr>
<tr>
<td>*Line resistance</td>
<td>0.05 Ω</td>
</tr>
<tr>
<td>*Line inductance</td>
<td>0.5 mH</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>12 kHz(simulation), 6 kHz (experimental)</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>12 kHz</td>
</tr>
</tbody>
</table>

* Values considered only in simulation

![Figure 3.10 - Closed-loop control diagram - current regulation](image-url)
3.5.1.1 Inner Control Loop (Output Filter Current)

The current control loop is depicted in Figure 3.11. The phase-locked-loop (PLL) algorithm gives a unitary magnitude three phase signal in phase with PCC fundamental phase voltages, represented by $\overline{s_{v,pcc}}$ in (3.22). Here, was used that one proposed in (MARAFÃO, DECKMAN, et al., 2005), that does not require frame-transformation.

$$\overline{s_{v,pcc}} = [s_{v_{A,pcc}}, s_{v_{B,pcc}}, s_{v_{C,pcc}}] \quad (3.46)$$

They are used to give a sinusoidal shape to the reference active current peak value. To compound output inverter current references, some compensation current vector, $\overline{i_{comp}}$, in (3.23) may be added to the reference.

$$\overline{i_{comp}} = [i_{comp,A}^*, i_{comp,B}^*, i_{comp,C}^*] \quad (3.47)$$

To have a properly current regulation in the main harmonic orders, a proportional multiresonant compensator is calculated to give the appropriate modulation index, $\overline{m_{inv}}$, represented in (3.24) for each phase of the inverter.

$$\overline{m_{inv}} = [m_{A,inv}, m_{B,inv}, m_{C,inv}] \quad (3.48)$$

Note that an additional feedback is implemented at the top of the diagram in the current figure. A unitary tridimensional vector is presented in (3.49). It is used in such representation to calculate possible instantaneous zero sequence component in the filter current measurement. Thus, a possible source of instability is avoided.

$$\overline{u} = [1, 1, 1] \quad (3.49)$$

The open-loop transfer function used to calculate the current compensator is presented in (3.50), where, $C_{i,f}(s)$ is the inverter’s current compensator, which is a proportional-resonant (PR) controller and $K_{i,f}$ is the current sensor gain.

$$OL \begin{bmatrix} i_{f,x}(s) \\ i_{f,x}^*(s) \end{bmatrix} = C_{i,f}(s) \frac{1}{C_{pk}} 2V_{dc}K_{i,f}G_{i_f}(s) \quad (3.50)$$
Such PR controller, for each phase, is presented in (3.51), whose parameters are shown in Table 3.2. Its open-loop bode diagram is shown in Figure 3.12. The PR controller was calculated for an open-loop frequency response of 1.2 kHz crossover frequency (i.e. around 1/10 switching frequency) and phase margin near to 60°.

\[
PR(s) = k_p + \frac{k_{i,h}}{n2\omega_b} \sum_h \frac{2\omega_b s}{s^2 + 2\omega_b s + \omega_h^2}
\]  

(3.51)

| Table 3.2 – Current Loop - PR controller |
|---|---|---|---|---|---|---|
|  | 1 | 5 | 7 | 11 | 13 | 19 |
| \(k_{i,h}\) | 94.82 | 89.12 | 83.42 | 66.30 | 54.90 | 9.268 |
| \(\omega_b\) | 6.28 rad | 6 | 1.8 |
| \(n\) | \(k_p\) |  

Figure 3.11 - Closed-loop control diagram - current regulation.

Figure 3.12 – Open loop frequency response for compensated and uncompensated grid-side-converter current loop.
### 3.5.1.2 Outer Control Loop (DC-bus voltage)

The closed-loop block-diagram for DC-bus is shown in Figure 3.13. The open-loop transfer function used to calculate the PI compensator is shown in (3.52), where $C_{v,dc}(s)$ represents its transfer function. For proper frequency response, a crossover frequency of 10 Hz and 60° phase-margin was considered in this calculation, giving the gains presented in Table 3.3.

$$ OL\left\{ \frac{\vec{v}_{dc}(s)}{\vec{v}_{dc}(s)} \right\} = -C_{v,dc}(s)K_{v,ac}V_{ac}^{pk}K_{v,dc} \frac{1}{K_{i,f}} G_{v,dc}(s) \quad (3.52) $$

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>proportional gain ($K_p$)</td>
<td>7.63</td>
</tr>
<tr>
<td>integral gain ($K_i$)</td>
<td>276.70</td>
</tr>
<tr>
<td>*Phase Margin (PM)</td>
<td>60°</td>
</tr>
<tr>
<td>*Cross-Over Frequency ($f_c$)</td>
<td>10Hz</td>
</tr>
</tbody>
</table>

*Open-loop frequency response parameters

Table 3.3 – DC-bus voltage PI controller parameters

![Figure 3.13 - Closed-loop control diagram – DC - bus voltage regulation.](image)

### 3.5.2 Generator-Side Control Loop

The generator side control loop is shown in Figure 3.14 and its parameters are presented in Table 3.4. A fast inner closed-loop regulates the stator currents. The mechanical speed loop is external and lower than the inner loop. It provides the necessary stator current reference to the inner control loop for torque variation and then speed regulation. Details of both closed-loops are describes in the following paragraphs.
Table 3.4 - Generator-Side Control Circuit Parameters

<table>
<thead>
<tr>
<th>Machine Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator series inductance ($L_s$)</td>
<td>5.1 mH</td>
</tr>
<tr>
<td>Stator series resistance ($R_s$)</td>
<td>1.6 Ω</td>
</tr>
<tr>
<td>Viscous damping coefficient ($B$)</td>
<td>$2.07 \times 10^{-3}$ N.m.s</td>
</tr>
<tr>
<td>*Total moment of inertia of the rotating mass ($J_{tg}$)</td>
<td>$5.64 \times 10^{-4}$ kg.m$^2$</td>
</tr>
<tr>
<td>Flux Linkage ($\Phi_{pm}$)</td>
<td>0.48 Wb</td>
</tr>
<tr>
<td>Number of pole pairs ($N_{pp}$)</td>
<td>4</td>
</tr>
<tr>
<td>Nominal rotation</td>
<td>1800 rpm</td>
</tr>
<tr>
<td>**Nominal terminal-voltage</td>
<td>220 V (rms)</td>
</tr>
<tr>
<td>Nominal Power</td>
<td>3kVA</td>
</tr>
<tr>
<td>Nominal Power Factor</td>
<td>0.8</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Converter Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switching frequency</td>
<td>$12$ kHz (simulation), $6$ kHz (experimental)</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>$12$ kHz</td>
</tr>
</tbody>
</table>

*Turine + generator
**Line-voltage

3.5.2.1 Inner Control Loop (Stator Currents)

An external loop regulates the mechanical speed through a proportional-integral (PI) controller. Then, its output generates the necessary stator reference peak current variation ($I^p k^+$). Subsequently, it is then multiplied by a symmetric three-phase set of signals of unit amplitude that are generated by the used estimator, which is the sliding mode observer (SMO), in this work. It is represented by the vector, $\bar{s}_\omega$, in (3.5), that is responsible for a sinusoidal synthesis.

$$\bar{s}_\omega = [s_{\omega_a}, s_{\omega_b}, s_{\omega_c}]$$ (3.53)

The vector $\bar{I}_s$, (3.54) gives the inner loop feedback, which makes the stator phase currents regulation through a PI controller, giving to the system an open loop response with cross-over angular frequency of $1/10$ switching frequency and PM of $60^\circ$.

Table 3.5 – Stator currents PI controller parameters

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>proportional gain ($k_p$)</td>
<td>1.9</td>
</tr>
<tr>
<td>integral gain ($k_i$)</td>
<td>76910</td>
</tr>
<tr>
<td>*Phase Margin (PM)</td>
<td>$60^\circ$</td>
</tr>
<tr>
<td>*Cross-Over Frequency ($f_c$)</td>
<td>$1$ kHz</td>
</tr>
</tbody>
</table>

*Open-loop frequency response parameters
The necessary three-phase modulation index signal, $\overrightarrow{m_{\text{rect}}}$, (3.53), is generated in order to build the converter switching pulses.

$$\overrightarrow{i_s} = [i_{s_a}, i_{s_b}, i_{s_c}]$$  \hspace{1cm} (3.54)

$$\overrightarrow{m_{\text{rect}}} = [m_{\text{rect}_a}, m_{\text{rect}_b}, m_{\text{rect}_c}]$$  \hspace{1cm} (3.55)

Figure 3.15 shows how the sinusoidal synthesis set of signals are performed. The estimated machine electrical angle $\hat{\theta}_e$ is used to an open control as a feed-forward signal. However, only an open loop control for the phase angle may not ensure good stability for the entire operation frequency range of the machine. Thus, quite little steady state error may appear between machine electromotive force and stator phase current angle. Therefore, a closed loop control must be added in order to give to the elements of $\overrightarrow{s_\omega}$, the necessary instantaneous electrical angle variation. Thus, the requirements described in the beginning of this section for torque control can be attended.

![Diagram](image)

Figure 3.14 – Closed-loop control diagram - stator phase currents regulation.

The following principle is used to make the feedback of the electrical angle closed loop control: each phase of induced armature electromotive force is orthogonal with respect to the permanent magnet field flux variation that induces itself. Therefore, those relative field flux variations proportional behavior may be represented as a three-phase set of unit amplitude signals, that is the vector $\overrightarrow{s_\omega}$. So, if each phase of the stator currents is exactly aligned with each phase of the armature electromotive force, the scalar product $\overrightarrow{i_s} \cdot \overrightarrow{s_\omega}$ must result a zero scalar.

The PI controller in the stator currents loop may give good reference tracking within the entire operation frequency range if well adjusted. However, some phase delay remains in a range close to the nominal frequency, what is not desired for a torque control. Depending on the phase-delay, some reluctance torque can attenuate the total electromagnetic torque, thus requiring a higher stator current to achieve the same value, what increases losses.
The stator currents closed-loop can be modeled as a sample delay, due to the digital control loop, followed by a low-pass filter transfer function that represents the approximate dynamics of the stator current closed-loop. Thus, a PI controller, $C_\theta(s)$, for the stator currents phase displacement can be calculated, according to the open loop transfer function in (3.56). Being, for linearity, the estimated instantaneous electrical angle, $\hat{\theta}_e$, a perturbation variable.

$$\frac{dp}{dp^*} = -C_\theta(s) \cdot I_{p}^{pk} \cdot \frac{1}{K_{i,s}} \left( \frac{\omega_{c,i}}{s + \omega_{c,i}} \right) e^{-s\tau_d} \cdot K_{i,s}$$

(3.56)

The cut-off angular frequency $\omega_{c,i}$ is the same of the stator current closed-loop. If its frequency response is plotted accordingly to Figure 3.17 for the calculated proportional and integral gains of $C_{i,s}(s)$, it is seen that the frequency at around -3dB (cut-off) is 1.4kHz, thus, $1400 \cdot 2 \cdot \pi \text{ rad} \cdot s^{-1}$ as indicated by arrows in Figure 3.16. Also it is indicated a phase-displacement, for the low-pass filter, of -2.5° at around the nominal electric frequency (60Hz). It is a good approximation for what is observed in simulation and experimental results. Thus, for (3.56) and (3.45), the calculated parameters of $C_\theta(s)$ are shown in Table 3.6. Those ones are suitable for a very stable angle estimation signal, which can be achieved in simulation. But, the use of the SMO algorithm in DSP, some undesired disturbances appear in the estimated electrical angle signal and may cause instability. However, such disturbances can be filtered by the PI controller $C_\theta(s)$, if a proper choice of its gains is made. The parameters of this controller are shown in Table 3.6.
3.5.2.2 Outer Control Loop (Mechanical Speed)

The open-loop transfer function used to calculate the PI compensator, \( C_\omega(s) \), for mechanical speed regulation is presented in (3.57) and its closed-loop in Figure 3.17. The parameters for this controller are presented in Table 3.7. The parameter \( K_\omega \) is the mechanical speed sensor gain.

\[
OL \left\{ \frac{\omega_m(s)}{\omega_m^*(s)} \right\} = C_\omega(s) \frac{K_\omega}{K_{ls}} \left( \frac{3N_{pp} \phi_{pm}}{2} \right) \left( \frac{1}{s_{tg} + B} \right) \quad (3.57)
\]

In Figure 3.18, it is shown how the mechanical speed reference is generated. Note that the high-frequency component of the pulsated DC-bus current is firstly filtered through a low pass filter of a bandwidth large enough to refine that signal and not slow down the control loop performance. After that, a moving average filter gives the properly power signal to the MPPT equation, shown in subsection 2.2, which gives the instantaneous mechanical speed reference value.

---

**Table 3.6 – Dot Product PI controller parameters**

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>proportional gain ((k_p))</td>
<td>0.01</td>
</tr>
<tr>
<td>integral gain ((k_i))</td>
<td>50</td>
</tr>
<tr>
<td>Phase Margin (PM)</td>
<td>90°</td>
</tr>
<tr>
<td>Cross-Over Frequency ((f_c))</td>
<td>89Hz</td>
</tr>
</tbody>
</table>

*[Open-loop frequency response parameters*}

Figure 3.16 – Frequency response: Compensated stator current closed-loop and its approximation by a first order low-pass filter.
3.6 Final Considerations

It concludes that the abc control approach may achieve similar results to conventional dq-frame approaches, in term of decoupling, since the proper implementation is adopted, as discussed herein, as: 1) to decoupling the phase quantities, it is required to comply with Kirchhoff’s Law, 2) to avoid introducing zero-sequence components through inaccuracy measurement, it is required to handle the variables as in (3.30) and (3.31).

From the generator-side control perspective, in order to ensure decoupling between torque and flux in abc-frame, a solution was described to allow torque control. The efficacy of the solutions and control design presented herein is evaluated in the next sections, through simulation and experimental results.

### Table 3.7 – Mechanical Speed PI controller parameters

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proportional gain ($k_p$)</td>
<td>0.08</td>
</tr>
<tr>
<td>Integral gain ($k_i$)</td>
<td>12.26</td>
</tr>
<tr>
<td>*Phase Margin (PM)</td>
<td>60°</td>
</tr>
<tr>
<td>*Cross-Over Frequency ($f_c$)</td>
<td>40Hz</td>
</tr>
</tbody>
</table>

*Open-loop frequency response parameters
CHAPTER 4

Simulation Results

4.1 Introduction

This chapter presents the whole system, modelled and implemented on Simulink and aims to analyse the performance of the proposed control system devised in abc-frame. For this purpose, variations on control references for both machine and grid-side are applied to.

As seen in chapter 3, the proposed torque control strategy in abcRF is developed in a way such that reluctance torque is regulated to zero value during PMSG operation. To achieve this, the stator currents vector must be controlled so that it is regulated in-phase with the back EMF. So, in the following sections the effectiveness of the proposed strategy for torque and flux decoupling is analysed for the operation of the PMSG.

The simulation diagram is shown in Figure 4.1. It is based on subsystems for wind turbine, B2B converter, load, grid, control loops and PWM control. It is important to mention that the homopolar terms of the generated non-active current reference are subtracted before being used at the reference input. Thus, some portion of load non-active current is not compensated. The way this subtraction occurs, is represented in Figure 4.2.

Figure 4.1 – Simulink block diagram for system simulation.
Figure 4.2 – Zero sequence component subtraction from compensation reference currents.

First result in subsection 4.2 shows the mechanical speed at the generator’s shaft while wind speed is under variation, then the MPPT tracking and mechanical speed loop are put to test. In subsection 4.3, firstly, the wind system is simulated in the nominal wind conditions showing two moments. In the first one, the power conditioner is injecting only the generators nominal power. In the second one, the compensation currents related to harmonic, load unbalance and reactive balanced compensation (non-active) take place with the active current, in order to verify the compensation action against those heavy grid disturbances.

It is worth to verify the system performance under hard fault conditions. So, in subsection 4.4, the system toward distorted and asymmetric grid voltage conditions is simulated under three-phase fault applied at the PCC. The load used in these simulations has a THD of 103.7%, 11.5% and 57.8% for phases A, B and C, respectively.

4.2 Reactive Power Variation

As it was mentioned in chapter 2, the power converter topology used herein provides decoupling between the grid-side and machine-side. This means that, if a properly sized capacitor at the DC-link is considered, any transient at the grid-side does not perturb the dynamic performance of the PMSG; and vice-versa.

The DC-capacitor of the power converter used in this work is well sized and some simulation result of reactive power variation at the grid-side can be useful to verify the decoupling between active and reactive power terms at the generator’s side. As shown in
subsection 3.2.1, a reactive current can be summed to the total reference current for the grid-side-control to inject reactive power into the grid. Then, as shown in Figure 4.3, a variation of reactive power is applied to the grid-side while active power is supplied to the grid. As shown in the bottom graphic of Figure 4.3, at 0.4s a degree of reactive power is applied to the grid-side and, at this moment, a perturbation on the active power at the grid-side is observed. Actually, this result highlights that there is a coupling between the power converter and the mains through the line impedance. On the other hand, the measured instantaneous active power at the machine-side does not suffer any disturbance, remaining unchanged during operation. Thus, this result clearly shows such a decoupling provided by the back-to-back topology, and proposed control strategy in abcREF.

![Figure 4.3 – Application of a reactive power variation.](image)

### 4.3 Wind Speed Variation

Wind speed variations disturb the mechanical power and then the generated electric power. Thus, the MPPT algorithm (explained in subsections 2.3 and 3.5.2) causes successive changes on the reference of the mechanical speed loop until the MPP is reached. Moreover, this variation also disturbs the mechanical torque at the machine’s shaft. This in turn means a perturbation that is applied on the mechanical speed control loop. So, a wind speed profile is...
applied herein in order to evaluate the performances of MPPT and mechanical speed control. As it is known, the mechanical speed closed-loop is external to the stator current one. Then, such a variation is interesting to test if the proposed control strategy for PMSG is really effective in cancelling the reluctance torque during operation. For this purpose, it is expected to visualize the vectors $\bar{t_{\delta}}$ and $\overline{s_{\omega}}$, which are explained in subsection 3.5.2.

Figure 4.4 shows the diagram block for wind turbine simulation. This structure is located inside “Wind Turbine + PMSG” subsystem shown in Figure 4.1. Figure 4.5 shows the simulation diagram related to the phase-displacement control loop presented in subsection 3.5.2. Further, the next figure depicts what is inside the subsystems of the previous block diagram. Figure 4.6a, shows in details the implementation of the phase-displacement control loop and in Figure 4.6b it is shown the generation of the stator reference currents.

In Figure 4.7, the wind speed varies from 0.5pu to 1pu in a 0.15s time interval. Note that the mechanical speed curve scale on such figure is reduced by a 0.5 factor. It is noted that a good reference tracking is reached for speed control. The dot product error, from the abcRF control, $dp_{error}$, has low peak-to-peak amplitude and zero mean value with low transitory variation. In addition, the dot product of $\bar{t_{\delta}}$ and $\overline{s_{\omega}}$ is also shown and presents similar dynamics, showing they are being maintained orthogonal. Note that the phase A signals of $\bar{t_{\delta}}$ and $\overline{s_{\omega}}$ are also shown. Note that from the dotted lines on this graphic, these signals are maintained 90° out of phase with each other during all operation. This would also be noted for the respective signals of phases B and C, as the orthogonally of such vectors was verified. These results indicate that the control is oriented and the active power is properly injected into the grid.

![Figure 4.4 – Simulink block diagram for wind turbine.](image)

![Figure 4.5 – Simulink block diagram for stator currents control.](image)
**Figure 4.6** – Block diagram for stator currents control: Phase-displacement control loop (a); stator reference currents generation (b).

**Figure 4.7** – System behavior under wind-speed variation.

### 4.4 Asymmetric and Distorted Grid Voltage

The main objective of this work is to implement the PMSG control in abcRF and to demonstrate its operation under ideal and non-ideal grid voltage conditions. Depending on the size of the capacitor considered in the DC-bus, non-ideal conditions at the grid-side can cause
oscillations in DC-bus voltage and, if the capacitor is sub-dimensioned, such oscillations may propagate to the mechanical torque of the generator, causing vibrations and reduction in efficiency. However, this last analysis related to the dimensioning of the capacitor is not addressed in this work, but only a study of correct system operation.

Refer to Figure 4.8 to verify the compensation current contribution despite of non-linear and unbalance loads at the PCC. Before the time instant of 0.4s, the grid currents, phases A and C, are highly distorted with a THD of 46% and 31.5%, respectively. The phase B is 5.6%. The power factor at the PCC is. After that time instant, the inverter supplies the non-active load current and the grid current presents low THD, 7% for phases A and C and 5% for phase B. Note that the grid current asymmetry is improved.

In the same figure, at the bottom, the DC-bus voltage presents a good tracking and disturbance rejection, but additional low order harmonic oscillation appears in its waveform during non-sinusoidal current supplying. It is worth of caution, since the MPPT is made through the mean power measured at the DC-bus, which gives the generator’s reference speed. On the other hand, it presents insignificant amplitude due to the capacitive filtering. Furthermore, the speed control loop has a bandwidth that is low enough to filter that disturbance.

![Figure 4.8 – Actuation of non-active current compensation at 0.4s.](image)
4.5 Three-phase Grid Fault

Figure 4.9 shows a three-phase instantaneous voltage sag of 0.30 p.u. due to a short circuit fault at the PCC from 0.35s to 0.37s. Such electric disturbance transient represents and external perturbation to the control scheme, highlighting the dynamic stiffness of the designed controllers. The transient increases the inverter output current and it reflects in the capacitor voltage at the DC-bus, however, the impact of this transient is practically eliminated at the generator side. It is possible to verify a disturbance rejection capability in the voltage control loop, in which the overshoot and undershoot reach 5%. The settling time of DC voltage control for this specific result takes approximately 200ms. Due to the oversizing of DC-capacitor, it is observed that this overshoot does not imply in significant torque oscillation in steady-state as shown in the bottom graphic of Figure 4.9. Moreover, the considered grid voltage is distorted.

Finally, it is noted that the inverter currents achieved about 2 p.u. values because no saturation limiter has been adopted herein. In practical applications it has to be taken into account.

![Diagram showing grid voltage and current responses](image.png)

Figure 4.9 – System behavior under grid voltage three-phase fault.
CHAPTER 5

Experimental Results

5.1 Test Bench

The used test bench is situated at LCCE and is part of the development of the master thesis in (SANTOS, 2015). It was proposed to be flexible and multi-use, so enabling this work implementation. It is composed basically by a VSR and a VSI, both connected in a back to-back configuration. Moreover, command, protection and signal conditioning circuitries, interconnect and control this system operation.

An eZdsp™ F28335 board, embeds a protection and command algorithm and also the code used to control the power converter.

5.1.1 Test Bench Start-up

A picture of the current bench is presented in Figure 5.1, where the main components for the system start-up are indicated.

The converter is connected to the grid through the contactor K2. In a first moment it is open and so is the contactor K1. In order to give pre-charge to the DC-bus capacitors, the DSP gives a signal command to K1 and it turns on. Thus, the converter is connected to the grid through the pre-charge resistors and a damped capacitor charging is performed through a resistor-inductor series connection. Such charging indicated in Figure 5.2. When the DC-bus voltage is 80% its nominal value, K1 turns off and K2 turns on. Then, a remote command is given to the DSP in order to start switching the IGBTs. In this step, the DC-bus voltage is controlled with a smooth reference ramp, driving it to its nominal regulation value, as indicated in the same figure.
The sampling and PWM processes may be properly synchronized so that it is possible to take advantage of switching frequency filtering. This approach is well established and brings advantage as the sampled signal is the average value of the control variable, avoiding to implement additional digital filters. On the other hand, some aliasing effect of low frequency will appear in the reconstructed signal if this sampling is not well aligned with modulation
signals. A DSP designed by power electronics applications may allow an analogic to digital conversion synchronized with the modulator (MATTAVELLI e BUSO, 2006), and so does the processor used in this work. Figure 5.3 shows the time instants that interruption sampling is triggered over an inductor current to perform analog to digital conversion.

![Figure 5.3 – Interruption sampling synchronization.](image)

### 5.2 Grid-Side Control

As mentioned before, the grid-side control starts at the last step of the start-up stage. The DC-bus voltage is then controlled with grid current reference variations that are necessary to provide appropriate instantaneous power balance. Thus, such voltage keeps stable and some power consumption due to parasitic impedances composes such energy exchange between DC and AC sides. This is depicted in Figure 5.4(a). Figure 5.4(b) shows reactive power generation with stable operation of DC-bus voltage.

![Figure 5.4 – DC-bus voltage regulation. (a) - No power reference; (b) - 2.75kVar reactive power reference. Line currents (CH1, CH2, CH3 - 5A/div) and DC-bus voltage (CH4 - 250V/div). Time scale: 5ms/div.](image)
The next figure depicts a result of the coupling between the three control loops that may occur in an experimental work. In Figure 5.5, the calculated $k_p$ for the 60Hz PR compensator is maintained in both cases. In letter “a”, the $k_i$ is 20. As shown, it is possible to note some distortion in the waveform, highlighted, and at the experimental set, some noise that does not appears in a proper operation can also be heard. For better reference tracking such gain might be increased but, little increasement, to 80, put the control loops in definitive unstable operation, as it can be seen in Figure 5.5(b). The source of this problem was not discovered during this work in a first moment and, what could be experienced is exactly what was studied and mentioned in (MATAKAS JÚNIOR, 2012).

It is worth to mention here, that the PR controller used in the experimental results was calculated as proposed in (YEPES, 2011), for an open-loop cross-over frequency of 600Hz (1 decade under switching frequency), thus giving a proportional gain ($k_p$) of 14 with a gain $k_i$ of 2000, as suggested in this mentioned reference.

At a moment, one could suppose that it could be feasible to use the lower gain ($k_i = 20$). However, it was observed that the simultaneous switching operation of the MS converter, summed by the high electromagnetic noise of the industrial environment where the laboratory is situated, bring additional electromagnetic interference (EMI) that is enough to trigger an unstable behaviour.

So, the decoupling suggested in the last mentioned reference was applied and then it was possible to use the desired gain $k_i=2000$ for proper reference tracking and stable control despite of EMI noises. This result is shown in Figure 5.6 for reactive power injection, with only the GS converter in work.

![No decoupling applied](image)

Figure 5.5 – Coupling between three-phase current loops (quadrature currents regulation). (a) - $k_i = 20$; (b) - $k_i = 80$. Line currents (CH1, CH2, CH3 - 5A/div) and phase A voltage (CH4 - 50V/div). Time scale: 5ms/div.
5.3 Machine-Side Stator Currents Control

In order to validate simulation results for the control loops presented in subsection 3.5, the synchronous generator was put under speed variation and the current tracking was tested for the entire variation range. I can be noted that the control loop that does not employ the dot product regulation (Figure 5.8), has this one variating during speed and current reference
variations, indicating great phase displacement related to the back-EMF. On the other hand, the results shown in Figure 5.9 depict the occurrence of a DT control by the addition of the phase-displacement regulation, as desired.

Figure 5.8 – Stator phase-currents control in abc-frame under speed variation.

Figure 5.9 – Stator phase-currents control in abc-frame with dot product regulation, under speed variation.
CHAPTER 6

Conclusions

This work proposed to study a three-phase WECS with its control system all devised in the “natural” abcRF connected to a low-voltage grid under distorted and asymmetric load and voltage conditions. From chapter 1, it is described in the literature that control scheme devised in stationary coordinates presents superior performance indices in term of power quality factor under non-ideal voltage condition avoiding using frame-transformation.

Regarding the power theories, they have been target of continuous effort from the scientific community in order to make them more suitable for modern grid constraints. The Conservative Power Theory was used in this work because it decomposes the instantaneous current and apparent power into many orthogonal current and power terms. These orthogonal terms bring flexibility and controllability to the PMSG system allowing the control scheme to track them independently.

From the bibliographic review in chapter 2, it was concluded that the PMSGs are applied to small power wind systems, and their efficiency is slightly superior to power systems based on induction generator. Moreover, it was concluded that a full-scale power converter used in PMSG-based systems may provide decoupling between grid-side and generator-side, what is suitable for a wind power system connected to an asymmetric and distorted grid (weak-grid).

Chapter 3 shown the requirements to implement a proper control scheme in abcRF in order to decoupling the phase quantities and avoid zero-sequence circulation. From the machine side control’s point of view, the required equation to calculate the torque and power in abcRF was presented, as well as a novel torque control in abcRF for PMSG. It was shown that in order to achieve torque control, some requirements must be pointed out, as: 1) stator current independently controlled; 2) constant field flux; 3) orthogonal spatial angle between the flux axis and the magnetomotive force, in order to avoid interaction of these variables. In fact, the stator current closed-control loop regulates the amplitude of stator phase-currents, matching the first requirement. The second requirement is met because of the permanent magnet poles of a PMSG. For the third condition, the closed-control loop presented in Figure 3.15 provides the
required field orientation. This is achieved by regulating the instantaneous dot-product of the $\overline{s_{\omega_1}}$ (which has unitary amplitude and is in phase with the sinusoidal field flux variation) and $\overline{I_s}$ vectors on zero mean value. So, it can be concluded that the machine-side control scheme was developed over these three requirements.

Simulation results in chapter 4 confirm the expected control performance for the PMSG, showing stator currents regulation with proper amplitude and dot-product regulation on zero (in-phase with the back EMF). Further, experimental results validated such proposal within a non-ideal environment. Such non-idealities may be, non-linearities in the real model, electromagnetical interferences present in experimental set and precision disturbances imposed by the sensorless algorithm. These results showed the stator currents amplitude regulation and the dot-product of $\overline{s_{\omega_1}}$ and $\overline{I_s}$ regulated on zero mean value during operation.

For the grid-side, simulation results were taken within a highly asymmetric and distorted load and grid-voltage condition. It was shown high improvement on total harmonic distortion currents trough load compensation using the CPT-based current decomposition. Then, it can be concluded that the natural frame-based control using the CPT is suitable for a TLTW VSI configuration. The experimental results in chapter 5, showed the difference between coupled and decoupled current control. For the coupled one, line currents presented distortion. On the other hand, for the decoupled one, line currents were presented purely sinusoidal. These experimental results confirmed the importance of those requirements for decoupling three-phase variables presented in chapter 3.

### 6.1 Future Works

The main goal of this work is study the PMSG in abcRF, and to confirm the proper operation of the power system. However, many other topics can be further throughout studies, and many questions are still not answered, such as:

- What is the best control scheme ($dq$-reference-frame, $\alpha\beta$-reference-frame, $abc$-reference-frame) applied to PMSG system under non-ideal load and voltage condition?
- What is the smallest size of DC-capacitor required to properly decoupling the grid side from generator side, and vice-versa, in order to avoid torque oscillation?

Then, as future works the following issues can be pointed out:
- An extension of this control can be performed to be suitable for induction machines.
- Investigating the machine control method proposed here for flux weakening approach.
- Investigating the proposed method in abc-frame in specific fields where analogic control may be required instead of digital one.
- An extensive comparison of the abc-frame machine control with $d$-$q$ and $\alpha$-$\beta$ frames.
- Experimental validation of three-phase load selective compensation, considering a load with three-wire and another load with four wire, but taking into account the limitation of the TLTW VSC, thus avoiding the compensation of homopolar terms.
- Experimental validation of a four-wire three-phase load selective compensation with a four-leg four-wire VSC, thus compensating homopolar terms.
- Experimental validation of the TLTW VSC in abc-frame control under voltage asymmetric sags. Also performing its comparison with $d$-$q$ and $\alpha$-$\beta$ based control under the same condition.

6.2 Publications in Conferences During the Masters


IEEE Brazilian Power Electronics Conference 2017 (COBEP 2017):

References


SINGH, M.; SANTOSO, S. *Dynamic Models for Wind Turbines and Wind Power Plants*. [S.l.].


DC-Bus Dynamic Modelling

The DC-bus dynamics may be modelled according to Figure A.1, where the generator plus its converter can be seen as a DC current source at the DC-side. It is made here as widely applied in literature, considering active power flow in a symmetric grid with unitary power factor (for simplicity).

Its regulation consists in maintaining the mean value of $i_{\text{dc},a}$ the same as $i_s$ so that all the generated active power is injected into the grid all the time. It is performed by maintaining the DC-bus mean-voltage at a constant nominal value.

![Figure A.1 – Equivalent circuit for DC-bus modeling.](image)

So, by the Kirchhoff Current Law in Laplace domain,

$$\dot{i}_s(s) - v_{\text{dc}}(s) \cdot sC = i_{\text{dc},a}(s) \quad (A.1)$$

And considering $i_s(s)$ a perturbation, by superposition,
\[
\frac{v_{dc}(s)}{i_{dc,a}(s)} = -\frac{1}{sC} \tag{A.2}
\]

But the VSI acts on grid-side quantities, thus variations in \(i_{dc,a}\) are performed indirectly, by directly actuating over the active current injected into the grid. By active power balancing, it can be taken into account.

\[
\overline{v_{dc}} \cdot \overline{i_{dc,a}} = 3v_{ac} \frac{i_{fa}^{pk}}{\sqrt{2}} \tag{A.3}
\]

Where \(v_{ac}\) is the phase-voltage rms value.

As the above quantities are time-variant, (A.3) is non-linear. Thus, linearizing it considering small signal variations around a quiescent point (ERICKSON e MAKSIMOVIC, 2007),

\[
\overline{v_{dc}} = V_{dc} + \overline{V_{dc}} \tag{A.4}
\]
\[
\overline{i_{dc,a}} = I_{dc,a} + \overline{I_{dc,a}} \tag{A.5}
\]
\[
v_{ac} = V_{ac} + \overline{v_{ac}} \tag{A.6}
\]
\[
i_{fa}^{pk} = I_{fa}^{pk} + \overline{i_{fa}^{pk}} \tag{A.7}
\]

substituting (A.4) to (A.7) in (A.3), then applying the Laplace transform and superposition in the resultant linear equation, brings the following relation, which is multiplied by (A.2) to achieve a transfer function of perturbations in DC-bus voltage related to perturbations in grid-current peak value.

\[
\frac{\overline{i_{dc,a}(s)}}{\overline{i_{fa}^{pk}(s)}} = \frac{3V_{ac}}{\sqrt{2}V_{dc}} \tag{A.8}
\]