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Comparative Analysis of Techniques to Emulate Synchronous Machines in Grid-connected Converters

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Belo Horizonte, January 13, 2020

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Dissertation presented to the Postgraduate Program in Electrical Engineering (PPGEE) of the Universidade Federal de Minas Gerais (UFMG), as a partial requirement for obtaining the degree of Doctor in Electrical Engineering.

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UNIVERSIDADE FEDERAL DE MINAS GERAIS
ESCOLA DE ENGENHARIA
PROGRAMA DE PÓS-GRADUAÇÃO EM ENGENHARIA ELÉTRICA

Belo Horizonte
January 13, 2020

F383c

Ferreira, Reginaldo Vagner.

Comparative analysis of techniques to emulate synchronous machines in grid-connected converters [recurso eletrônico] /Reginaldo Vagner Ferreira. - 2019.

1 recurso online (xxi, 116 f. : il., color.) : pdf.

Orientador: Sidelmo Magalhães Silva.

Coorientador: Danilo Iglesias Brandão.

Tese (doutorado) - Universidade Federal de Minas Gerais, Escola de Engenharia.

Anexos: f.97-116.

Bibliografia: f.83-96.

Exigências do sistema: Adobe Acrobat Reader.

1. Engenharia Elétrica - Teses. 2. Redes elétricas - Teses.
3. Microrredes - Teses. I. Silva, Sidelmo Magalhães. II. Brandão, Danilo Iglesias. III. Universidade Federal de Minas Gerais. Escola de Engenharia. IV. Título.

CDU: 621.3(043)

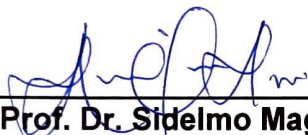
"Comparative Analysis of Techniques to Emulate Synchronous Machines in Grid-connected Converters"

Reginaldo Vagner Ferreira

Tese de Doutorado submetida à Banca Examinadora designada pelo Colegiado do Programa de Pós-Graduação em Engenharia Elétrica da Escola de Engenharia da Universidade Federal de Minas Gerais, como requisito para obtenção do grau de Doutor em Engenharia Elétrica.

Aprovada em 23 de outubro de 2019.

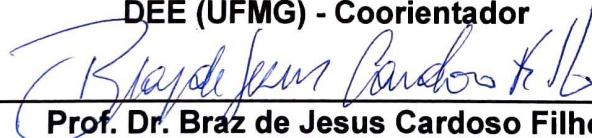
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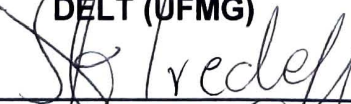
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Prof. Dr. Clodualdo Venício de Sousa
Instituto de Engenharia Elétrica (UNIFEI)

Acknowledgements

Throughout the writing of this dissertation I have received a great deal of support and assistance.

I would first like to thank God for my health and for surrounding me with good people.

I thank my wife Gabi for the support, companionship and incentive throughout this project. She spent many days explaining to Sophia that "Dad needed to study".

Thanks also to little Sophia, who although does not understand yet, gives great meaning to my life and my projects.

I would also like to thank my advisor, Dr. Sidelmo Magalhães Silva and my co-advisor Dr. Danilo Iglesias Brandão, for their valuable guidance.

I would like to acknowledge my colleagues from the Laboratory Tesla Power Engineering, especially Armando Guedes, Gideon Lobato, Fernando Venâncio, Rodrigo Rodrigues and Helio Antunes.

In addition, I would like to thank my parents, who taught me the main thing: moral values.

To Gabi, Sophia, Luíza
Rodrigues, Melissa and Luísa
Guedes.

Abstract

The increase in the application of electric microgenerators, mainly from renewable primary sources, such as wind and photovoltaic, has motivated the study of several issues related to the operation of these systems. Both in the context of connection to the grid and in its isolated operation, these microgenerators, which use power electronic converters as element of connection between primary source and grid, still present technological challenges, ranging from the converter topology to the implemented control techniques. Many control levels are presented in literature. The algorithm responsible for controlling the active and reactive power flow between converter and grid is called primary control. Among the most widespread primary control techniques are droop control, Virtual Synchronous Generator - VISMA, Virtual Synchronous Generator - VSG and synchronverter. The first is undoubtedly the simplest and most used, while the others have inherited advantages of the synchronous machine, such as virtual inertia and friction coefficient. The main difference between the VISMA, the VSG and the synchronverter techniques is that the VISMA is applied to current-controlled converters, while VSG and synchronverter operate as voltage-controlled converters. In this work, a literature review is carried out on the operation modes of converters in distributed generation and microgrids, hierarchical control levels, distributed generation requirements and primary control techniques. Then the focus turns specifically to a unified analysis of the operation of the synchronverter, the VSG and the droop-controlled converters by means of a small-signal model. In this step, we reached a high accuracy small-signal model when compared to the dynamic model. The comparison between the techniques was possible through the equivalence of gains, developed also in this work. Simulation and experimental results have shown that for a considerable range of gains, droop controlled converters and synchronverters have similar dynamic response, but the synchronverter allows a more extensive range of gains in the stable region. Continuing the search for contributions to the converters that emulate the synchronous machine, this dissertation addresses the resilience of the synchronverter to unbalanced sags and unbalanced grid voltages, presenting a proposal named positive-negative sequence synchronverter with enhanced fault ride-through capability for AC grids. Inspired by the Fortescue theorem, the technique allows to limit the output currents of the converter, to reduce power oscillations and to avoid unexpected converter shutdowns when subjected to unbalanced sags and voltage unbalance. The power flow control of single-phase converters emulating synchronous machine is also explored in this work. The proposal is an adaptation of the existing synchronverter, initially designed for three-phase applications. Simulation results illustrate the performance of the proposed technique in the grid-connected mode, including voltage sags conditions and harmonic distortions, in addition to the isolated mode.

Resumo

O crescente número de unidades microgeradoras de energia elétrica, sobretudo a partir de fontes primárias renováveis, tais como eólica e fotovoltaica, motivou o estudo de diversas questões relacionadas ao funcionamento destes sistemas. Tanto no contexto de conexão com a rede principal quanto em sua operação ilhada, estes microgeradores, que utilizam conversores eletrônicos de potência como elemento de conexão entre fonte primária e rede, ainda apresentam desafios tecnológicos que vão da topologia do conversor às técnicas de controle implementadas. Diversos níveis de controle são apresentados na literatura. Denomina-se controle primário o algoritmo responsável por controlar o fluxo de potências ativa e reativa entre conversor e rede. Dentre as técnicas mais difundidas de controle primário estão o controle por *droop*, o *Virtual Synchronous Machine* - VISMA, o *Virtual Synchronous Generator* - VSG e o *synchronverter*. O primeiro é sem dúvida o mais simples e mais utilizado, enquanto que os demais apresentam vantagens derivadas da máquina síncrona, tais como inércia virtual e coeficiente de fricção. A principal diferença entre o VISMA, o VSG e o *synchronverter* é que o primeiro se aplica a conversores controlados em corrente, ao passo que VSG e *synchronverter* operam como conversores fontes de tensão. Neste trabalho é realizada uma revisão bibliográfica sobre os modos de operação de conversores em geração distribuída e microrredes, os níveis hierárquicos de controle, os requisitos de geração distribuída e as técnicas de controle primário. Em seguida, o foco volta-se especificamente para uma análise unificada do *synchronverter*, do VSG e dos conversores controlados por *droop* por meio de um modelo de pequenos sinais. Nesta etapa chegou-se a um modelo de alta acurácia quando comparado com o modelo dinâmico. A comparação entre as técnicas foi possível através da equivalência de ganhos, desenvolvida também neste trabalho. Os resultados de simulação e experimentais mostraram que para uma considerável faixa de ganhos, conversores controlados por *droop* e *synchronverters* apresentam resposta dinâmica similar, porém o *synchronverter* permite uma faixa mais extensa de ganhos na região estável. Continuando a busca por contribuições aos conversores que emulam a máquina síncrona, a tese aborda o problema da resiliência do *synchronverter* trifásico frente a afundamentos desequilibrados e desequilíbrio de tensão na rede elétrica, apresentando uma proposta denominada *positive-negative sequence synchronverter with enhanced fault ride-through capability for AC grid* (um *synchronverter* de sequência positiva e negativa com resiliência aumentada frente a desequilíbrio de tensão). Inspirada no teorema de Fortescue, a técnica permite limitar a corrente de saída do conversor, reduzir oscilações nas potências e evitar desligamento inesperado do conversor quando submetido a afundamentos desequilibrados e desequilíbrio de tensão. O controle de fluxo de potência para conversores monofásicos emulando a máquina síncrona também é explorado no trabalho. A proposta é uma adaptação do já existente *synchronverter*, concebido inicialmente para aplicações trifásicas. Resultados de simulação ilustram o desempenho da técnica proposta no modo conectado à rede, incluindo condições de afundamento de tensão e distorções harmônicas, além do modo isolado.

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Notation

a	Scalar
$\vec{a}$	Vector
$\mathbf{A}$	Matrix
$\dot{a}$	Derivative term
$\vec{a}_i$	i -th column vector of matrix A
a_{ij}	Element from row i , column j of matrix A
$\tilde{a}$	variable with a small variation around an operating point
$\langle \cdot, \cdot \rangle$	Scalar (dot) product operator
A^T	Transpose of matrix A
A^*	Reference values

List of Acronyms

AC	Alternating Current
ANEEL	Agência nacional de energia elétrica
CCMG	Central Controller of the Micro-grid
DC	Direct current
DDSRF-PLL	Decoupled Dual Synchronous Reference Frame Phase-locked Loop
DG	Distributed generator
EPP	Electronic Power Processors
GDC	Generalized droop control
HIL	Hardware in the Loop
HVDC	High-voltage direct current transmission system
LC	Load Controller
LVRT	Low-voltage ride-through
MAS	Multi-agent system
MG	Microgenerator
MPPT	Maximum Power Point Tracking
PES	Primary Energy Source
PCC	Point of common coupling
PLL	Phase-Locked Loop
PI	Proportional-integral controller
PN	Positive-negative
PV	Photovoltaic
PWM	Pulse Width Modulation
RPI	Reactive Power Injection
SOC	State of charge
STATCOM	STATic synchronous COMPensator
UPS	Uninterruptible Power System

VISMA	Virtual Synchronous Machine
VSC	Voltage source converter
VSG	Virtual Synchronous Generator
VSI	Voltage source inverter

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1 Introduction

The interest in distributed generation and its integration with the power grid has grown significantly in the last decade (Brabandere et al. 2007b). This phenomenon is mainly associated with energetic crises, economic issues and governmental policies.

Renewable sources have been used to generate electric energy through grid-connected inverters (Borazjani et al. 2014). Among the called renewable sources, the most widely used around the world is the source based on wind, while the photovoltaic has been experiencing clear growth in recent years. The expectation is that these alternative sources will represent, each year, a more relevant part of the global energy matrix, as it has already been happening most strongly in some European countries (França, Castro, and Aredes 2015).

The use of small and medium size renewable power systems, connected to the distribution or transmission systems is named distributed generation. In general, the distributed generators (DGs) are constituted of primary energy source (PES), electronic power processors (EPPs) and output passive filters, controlled through specific control strategies. The introduction of these elements, mainly EPPs, can cause severe distortions and voltage fluctuations, therefore, voltage and frequency shall be maintained within the acceptable power quality limits (Guerrero et al. 2013).

In Brazil, in terms of regulation, the interaction between generation systems and electric grid had its first step after the publication of the Resolution 482/2012 (Energia Elétrica 2012), by means of which the Brazilian electrical system started to allow the integration of distributed generation sources into the grid, encouraged through an energy compensation system. This resolution was updated by the Agência Nacional de Energia Elétrica - ANEEL, when in 2015 was published the Resolution 687/2015 (Energia Elétrica 2015), improving significant changes to the microgeneration sector, such as new ratings for power ranges, adjustments in the process of grid accessibility and net measurements benefits (Alonso et al. 2017).

Coordinated grouping and control of DGs gave rise to the micro-grids. This term refers to a set of micro-sources and loads operating as a single-controllable system, which can act autonomously, dispensing when it is convenient, the connection to a main grid. Usually, the sources that make up the micro-grid are low power generators (up to 200kW) and must present reliability and low cost (Vechiu et al. 2011).

The micro-grid must operate in both islanded and grid-connected mode. The transition between these operation modes should be done smoothly and can be both intentional and unintentional, linked to an external event, such as, for example, the loss of a feeder circuit connected to the micro-grid bus (Xu, Li, and Tolbert 2012). As for regulation, there are no Brazilian laws or standards for the interaction between micro-grid and grid.

Some key issues for distributed generation and micro-grids are the challenges related to control techniques when there is a significant number of interconnected micro-generators (Piagi and Lasseter 2006). Each converter must be able to respond to load variations in a predetermined way, without

communication with other sources, allowing the insertion of each new source independently (Lasseter 2001).

In the context of micro-grids, the literature presents control methods based on a central controllers for the micro-grid (CCMG), controllers for each microgenerator (MG) and load controllers (LC). In the centralized operation, MG and LC receive a set-point from the CCMG, being the communication between these controllers a critical issue. In the decentralized case, control decisions are established locally through MG and LC only.

A conventional way of achieving decentralized control for the micro-grid without the need for complex communication systems is the droop control, which consists in controlling the active and reactive powers through small variations in the frequency and output voltage of the converter. In case of multiple sources, the power is shared among the generators, according to their characteristic curves.

This work presents power control algorithms for grid-connected converters emulating synchronous machines in distributed generation and micro-grid applications. Several techniques are studied and the deepening is carried out around the control of the single-phase synchronverters, the dynamic analysis of droop-controlled converters and synchronverters, and finally, the enhancement of fault-ride through in synchronverters subjected to unbalanced grids.

1.1 Motivation and Relevance

The power sharing control between DGs has been carefully studied, and consequently, many strategies have been published in literature, in which the droop control is the most consolidated, due to its simplicity, high reliability during the island operation and autonomous power sharing among generators.

Other three control strategies have been receiving attention from researchers, named: virtual synchronous machine (VISMA), synchronverter and virtual synchronous generator (VSG). In all these strategies, the interface converter between primary source (such as photovoltaic and wind) and power grid is regulated through a control scheme that emulates a synchronous generator.

The four strategies mentioned make it possible to operate the converter, both in the connected mode and in the islanded mode, thus avoiding control schemes that require important modifications from one mode of operation to another. An advantage of the VISMA, the Synchronverter and the VSG, raised by general literature, in comparison with the droop control is the possibility of choosing virtual parameters, such as inertia, friction coefficient, field inductance and mutual inductance, and even adjust these parameters during operation.

Evaluating the existent techniques, it is clear that there are still opportunities to improve them, which involve the effort to minimize the complexity of the algorithms, to reduce the problems resulting from grid connection transitions, to mitigate problems of parallelism of multiple converters, to improve fault-ride through, among others.

Specifically, this work attach the necessity of a deep understanding of the control strategies behavior, a single-phase synchronverter applied to residential PV power systems and a proposal to enhance fault-ride through in three-phase synchronverters based on the positive-negative sequence theory.

1.2 Objectives

The general objective of this work is to analyze the grid-connected converters emulating synchronous machines, and to propose improvements that contribute to the parallelism between micro generators, enhance reliability against disturbances and contribute to the stability of the system.

The specific objectives are:

- a comparative study of connected converters emulating synchronous machines: droop based converters, virtual synchronous generators, virtual synchronous machines and synchronverters;
- a single-phase synchronverter proposition for residential PV power systems;
- elaboration of an unified approach to model synchronverter and droop controlled converters;
- proposition of a positive-negative sequence synchronverter to improve fault-ride through in three-phase grid-connected converters.

1.3 Dissertation Contributions

The dissertation contributions are divided into three parts: the first one related to advances in the dynamic analysis of droop-controlled converters and synchronverters; the second concerning the proposed positive-negative sequence synchronverter to enhance fault-ride through in AC grids; and the last one relative to the achievements with the proposal of a single-phase synchronverter. The main contributions are:

1) From the dynamic analysis of droop control and synchronverter by means of a complete small signals model:

- An unified approach to model synchronverters, droop and other virtual synchronous machine control strategies for small-signal analysis;
- a parameter comparison of the droop control converter and the synchronverter, including both active and reactive control loops;
- an unified small-signal model framework to model the droop control, synchronverter and other control algorithms that mimic the operation of a synchronous generator, including the current dynamics;
- a closer view on the similarities and differences between the droop-controlled converter and synchronverter;
- presentation of the fact that for a given set of equivalent parameters, the synchronverter has a wider stability margin than the droop-controlled converter, considering variations on the voltage magnitude gain.

2) From the positive-negative sequence synchronverter technique:

- limiting synchronverter output currents during balanced or unbalanced voltage-sags;
- reducing of power oscillations during voltage sags;
- avoiding unexpected converter shutdown due to disturbances.

3) From the single-phase synchronverter:

- an adaption of the three-phase synchronverter for application in single-phase systems, avoiding major changes in the original control model.

1.4 Text Organization

After this introductory chapter, in the Chapter 2, a literature review on the operation modes of converters in distributed generation and micro-grids is realized, the classification of the converters as to the function performed is presented, the hierarchical control levels in micro-grids are explained, and a brief review of distributed generation requirements is performed for contextualization.

In Chapter 3 the most consolidated techniques for the control of active and reactive power between converter and grid distribution are presented. Namely, techniques called droop control, VISMA, Virtual synchronous generator and Synchronverter are exploited.

Chapter 4 presents a dynamic analysis of grid-connected droop-controlled converters and synchronverters. First, an unified small-signal model is proposed. From this, models for droop control and synchronverter are presented, as well as their parameters relations. Simulation and experimental results validate the small-signal models, and allow the dynamic behavior evaluation of the synchronverter and the droop-controlled converter.

In Chapter 5, a positive-negative sequence synchronverter are proposed, aiming to improve fault-ride through in AC grids.

Chapter 6 presents the single-phase synchronverter applied to residential PV power systems, including in addition to the theoretical description of the power and control parts, a set of simulation results that validate the proposal.

Finally, in chapter 7 are presented final remarks and future works.

2 Converters for Distributed Generation and Micro-grid Applications

The present chapter establishes some concepts to form a necessary understanding on converters in distributed generation and micro-grids. Initially, the operating modes of the converters in micro-grids are explained. Next, the converters are classified according to their function and a brief description and hierarchical control levels are presented. Finally, distributed generation requirements are introduced to clarify recommendations regarding cease to energize, control capability and other relevant features.

2.1 Introduction

The converters in DG needs to inject power following the grid voltage and can be used to support the grid, correcting power factor or contributing to the voltage regulation during voltage sags for example. On the other hand, in a MG application it is possible to have converters with the same characteristics cited above, but one or more converters need to assume the voltage and frequency control when a main grid is not present.

By referring to the "primary control of a grid-forming converter, operating in an islanded mode", for example, at least three concepts need to be known for the correct understanding of such converter: which are the operation modes, how the converters are classified as their function and what the control levels of the converters in DG and MG are composed by.

2.2 Operating Modes

The operating modes in micro-grids and, consequently, of its converters, can be found in the IEEE Standard 1547.4 (["IEEE Guide for Design, Operation, and Integration of Distributed Resource Island Systems with Electric Power Systems" 2011](#)). These operating modes are: islanded mode, grid-connected mode, reconnection mode and transition to and from islanded mode. The following subsections explain these concepts.

2.2.1 Islanded Mode

An islanded micro-grid may occur with either scheduled maintenance operations or switching incidents, such as a grid fault or power quality problems (Katiraei, Iravani, and Lehn 2005). It may also exist in cases where no primary grid is available because of geographic issues, for example.

In Brazil, the isolated systems are located mainly in the states of the North Region, specially where it is very difficult to deliver power to remote areas from centralized generation and the power system is characterized basically by the large number of diesel generating units (Energia 2013, *Sistema Interligado Nacional: Sistemas Isolados*). Also in the northeast of Brazil are found isolated systems, where the Fernando de Noronha island, in the state of Pernambuco (Eletrobrás 2009), deserves special mention.

In the islanded operation, the microgenerators are responsible for maintaining the integrity of the micro-grid without the aid from a main source (Vandoorn et al. 2011b) and the reference voltage is the voltage of an arbitrary node in the micro-grid. Islanded microgenerators have very different characteristics when compared to conventional electrical systems and therefore require different operating and control methods. Several papers present control strategies for the island operation (Guo et al. 2012, Lopes, Moreira, and Madureira 2006, Vandoorn et al. 2013a, Vandoorn et al. 2012, Vandoorn et al. 2011a, Shao, Wei, and Nie 2011). The main challenges of this operation mode are (Mahmoud 2014):

- voltage and frequency control;
- power balance between supply and demand;
- quality of power;
- communication between microgenerators (depending on the control strategy).

Since microgeneration units are mostly interfaced by electronic power devices, the control of the micro-grid depends on the control of the inverter, which can be divided basically into two types: PQ (active and reactive power) control and Vf (voltage and frequency) control. The PQ control allows each unit to inject power into the micro-grid in response to set-points in order to obtain rational use of the various resources that make up the system. On the other hand, the Vf control consists on regulating the voltage and frequency in the micro-grid, once in the island mode the main grid with its stabilizing capacity is not present. During the islanded operation, at least one of the converters must operate in Vf mode and must be sized to a power range that allows the voltage magnitude and frequency to be maintained close to their nominal values.

In the islanded operation, the power balance can have two different scenarios:

- over-generation - in this case, the extra energy generated by renewable sources is stored in battery systems, depending on their state of charge. If the over-generation lasts for a long time, the power generated by renewable sources must be reduced to meet the actual load power consumption, according to the distance-based power sharing criterion.
- under-generation - this condition is more critical than the previous one. The power generated within the micro-grid is not sufficient to maintain the loads and therefore, additional power must be provided by battery systems according to their state of charge. Obviously, the stored energy will be dispatched for a limited duration of time and after this interval it may be necessary to reject loads or to start back-up generators at least until the states of charge of the storage systems are restored (Tenti et al. 2013).

2.2.2 Grid-connected Mode

In the grid-connected mode, the microgenerators exchange power with the grid. The inputs to the controller are basically the set-points for the output power and local bus voltages. In order to obtain frequency control, detection and synchronization algorithms are used. They are called Phase-Locked Loops (PLLs) and are widely used in DG applications. Generators connected to the grid mostly operate in the grid-feeding function, with the possibility of being designed to operate as grid-support. These definitions related to the function of the DGs are presented in section 2.3.

2.2.3 Re-connection Mode

The re-connection process requires some care, especially regarding the previous voltage synchronism between converters and grid. Voltage, frequency and phase angle between the two systems must meet acceptable limits before the transition is started, for example the values specified in the IEEE Std 1547-2018 ([“IEEE Standard for Interconnection and Interoperability of Distributed Energy Resources with Associated Electric Power Systems Interfaces” 2018](#)), transcribed to the Table 2.1. If there are multiple "islands," some scheduling strategy must be adopted to avoid simultaneous re-connection of multiple sources in parallel.

Table 2.1: Synchronization parameter limits for synchronous interconnection to an EPS, or an energized local EPS to an energized Area EPS Source: IEEE Std 1547-2018.

Aggregate rating of DR units (kVA)	Frequency (Hz)	Voltage (%)	Phase angle (°)
0 – 500	0.3	10	20
> 500 – 1 500	0.2	5	15
> 1 500 – 10 000	0.1	3	10

2.2.4 Transition to Islanded Mode

As already mentioned, the transition to island mode may result from a scheduled or unintentional event. Knowledge of operating conditions prior to islanding facilitates smooth transition, for example in response to the occurrence of disturbances in the main grid.

During the transition it is necessary that the available converters have the ability to maintain voltage and frequency in the micro-grid. There should also be protection relays that ensure the integrity of the system. If sufficient microgenerators are not present, then the loads are shedded. Other issues to be evaluated are the balance between supply and demand, power quality, communication between components of the micro-grid, and specific issues of each micro-generator, such as lack of inertia and response time (Bhaskara and Chowdhury 2012).

The converter control must avoid transients that may trip the protections and consequently cause the disconnection of some loads and even of the whole micro-grid.

2.3 Classification of Converters

Power electronics converters can be classified according to the functions they perform in the micro-grid, as grid-forming, grid-feeding or grid-support. Figure 2.1 illustrates the schematic diagram representation of these converters.

2.3.1 Grid-forming Converter

When the micro-grid operates connected to the grid, voltage and frequency are imposed by the grid itself. In this case, the grid-forming must inject power into the system, following the grid references, as the other converters. In the islanded operation, however, the grid-forming is controlled to operate as an ideal voltage source, imposing fixed amplitude and frequency for the micro-grid and for this purpose it must be provided with energy storage systems. In addition, it must be able to respond to load variations and manage the transition from the connected mode to the isolated.

The grid-forming converter has low output impedance and, therefore, requires a precise synchronization system to operate with other grid-forming converters. The power sharing among parallel-connected grid-forming converters is a function of their output impedances (Rocabert et al. 2012). Figure 2.1(a) illustrates the simplified representation of this converter.

2.3.2 Grid-feeding Converter

The grid-feeding converter is controlled as a current source, according to Figure 2.1(b) and presents high output impedance. It is widely used in the connected mode, because instead of determining the voltage and frequency at its terminals, this converter follows the magnitude and phase of the voltage measured at the connection point. In this way, the grid-feeding usually operates with unit power factor and at the maximum power point tracking (MPPT), with example of the photovoltaic generators. In order to achieve the MPPT, methods such as constant voltage, current sweep, incremental conductance or the so-called perturb and observe are used, all of these applicable to photovoltaic systems.

In the islanded mode, this converter is also applicable as long as it operates in parallel with at least one grid-forming converter, a grid-support converter or a local synchronous generator.

2.3.3 Grid-support Converter

The main purpose of the grid-support converter is to participate in the voltage and frequency regulation of the AC grid through the active and reactive power control delivered to the grid. This converter can be of two types, as illustrated by Figures 2.1 (c) and (d):

- Grid-support converter controlled as a current source: in this case, the main objective is not to supply the load connected to the micro-grid, but to contribute to the voltage and frequency regulation, both on the grid side and on the micro-grid. It has as fundamental characteristic the high output impedance and, for this reason, its control is naturally more stable than the grid-support converter controlled as a voltage source.
- Grid-support converter controlled as a voltage source: by presenting low impedance output, this converter is connected to the grid through a series impedance, which can be a physical component or a virtual element emulated in the current control loop. In the context of the isolated micro-grid, this type of converter can be used to regulate the voltage and frequency, dispensing with the need for a grid-forming converter.

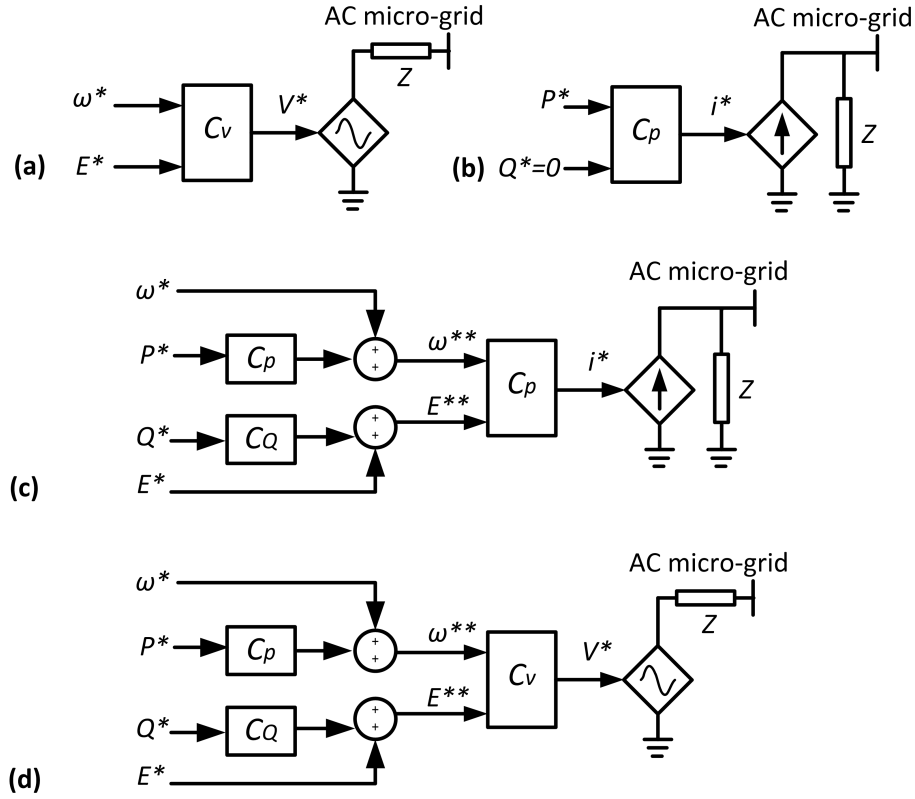


Figure 2.1: Representation of converters: (a) grid-forming converter; (b) grid-feeding converter; (c) grid-support controlled as a current source and (d) grid-support controlled as a voltage source (Adaptated from Rocabert et al. 2012).

2.4 Hierarchical Control of Micro-grids

Before explaining the hierarchical control levels to which the converters are subjected in micro-grid applications, it should be clarified that such stratification can vary if the control strategy is centralized or decentralized.

A centralized control system depends on a communication structure between the converters and a central controller, which performs calculations and determines the control actions for the generating units (Olivares et al. 2014). On the other hand, in the decentralized approach, each drive is controlled by its local controller, which does not need to know all system variables, but receives only local information.

Centralized approaches are typically multi-agent systems (MASs), which essentially integrate the grid operator, a central micro-grid controller and local controllers, either microgeneration or loads (Dimeas and Hatziargyriou 2005).

The main advantages of the decentralized control strategy are the reduction of losses, the ease of the micro-grid expansion, greater reliability, economic issues (cost of installation and maintenance of the data communication and data storage) and the possibility of bidirectional power flow (Borazjani et al. 2014).

The control systems in micro-grids are presented in three hierarchical levels (Bidram and Davoudi 2012, Brandão 2015, Haiyun et al. 2013), in addition to the PWM level, as it is explained in the following sections.

2.4.1 PWM Level - Voltage and Current Control

The inner control loops of a microgeneration converter are commonly named zero level control, usually implemented through PQ (Li, Vilathgamuwa, and Loh 2004) or voltage control methods. Figure 2.2 shows the voltage control in a block diagram. A cascade control is observed, in which the output of the voltage controller provides the reference for the current control loop and this, in its turn, produces the references for the PWM, which switches the static power devices. A passive filter (Gabe 2008) is required to provide the coupling impedance with the AC grid.

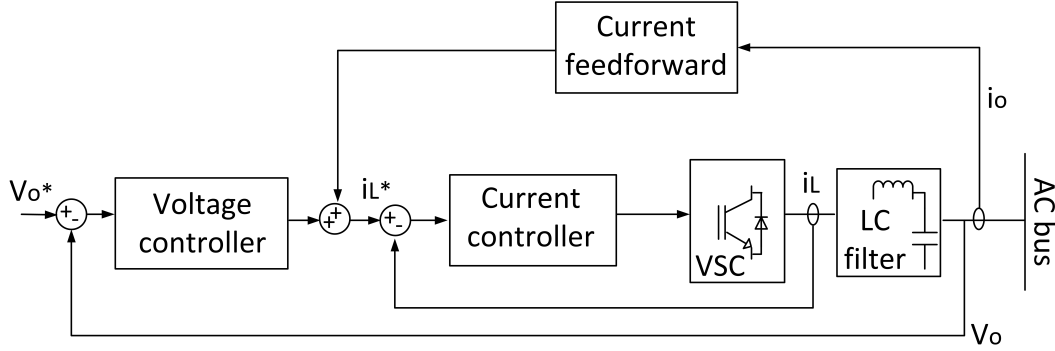


Figure 2.2: PWM level: voltage and control loops.

The PWM level control is often performed through a proportional-integral controller (PI) in the voltage loop in cascade with a proportional controller (P) for the current loop. Repetitive control, dead-beat, and sliding modes for the voltage loop are also reported in the literature and, in addition, the hysteresis control in the current loop (Miveh et al. 2015). In three-phase converters, the choice of technique depends, among other things, on the adopted references: synchronous ($dq0$), stationary ($\alpha \beta 0$) or natural (abc).

2.4.2 Primary Level

The primary control is designed to meet the following requirements: stabilize voltage and frequency; attenuate currents that may cause overload on static power devices and avoid damaging bus DC capacitors. This control level provides the references for the voltage and current control loops of the converters (Bidram and Davoudi 2012). Islanding detection and algorithms for maximum power point tracking (MPPT) are also treated as the primary control level (Brandão 2015).

In the decentralized approach, the primary control is also intended to control the active and reactive power sharing among the microgenerators in the presence of linear and non-linear loads. In the grid-connected mode or operating as grid-feeding in isolated mode, this level receives the active and reactive power references from the upper level. In island mode, operating as a grid-forming converter, it establishes fixed voltage and frequency for nominal load values. The droop control method is often used at this level to emulate physical behaviors that make the system stable and damped (Guerrero et al. 2011). More complex and sophisticated than droop control are the techniques that emulate more completed synchronous machine models, which will also be covered in the next chapter.

2.4.3 Secondary Level

When voltage and frequency deviations exceed permitted limits due to important load variations, the voltage and frequency references need to be adjusted to ensure the power quality of the system (Haiyun

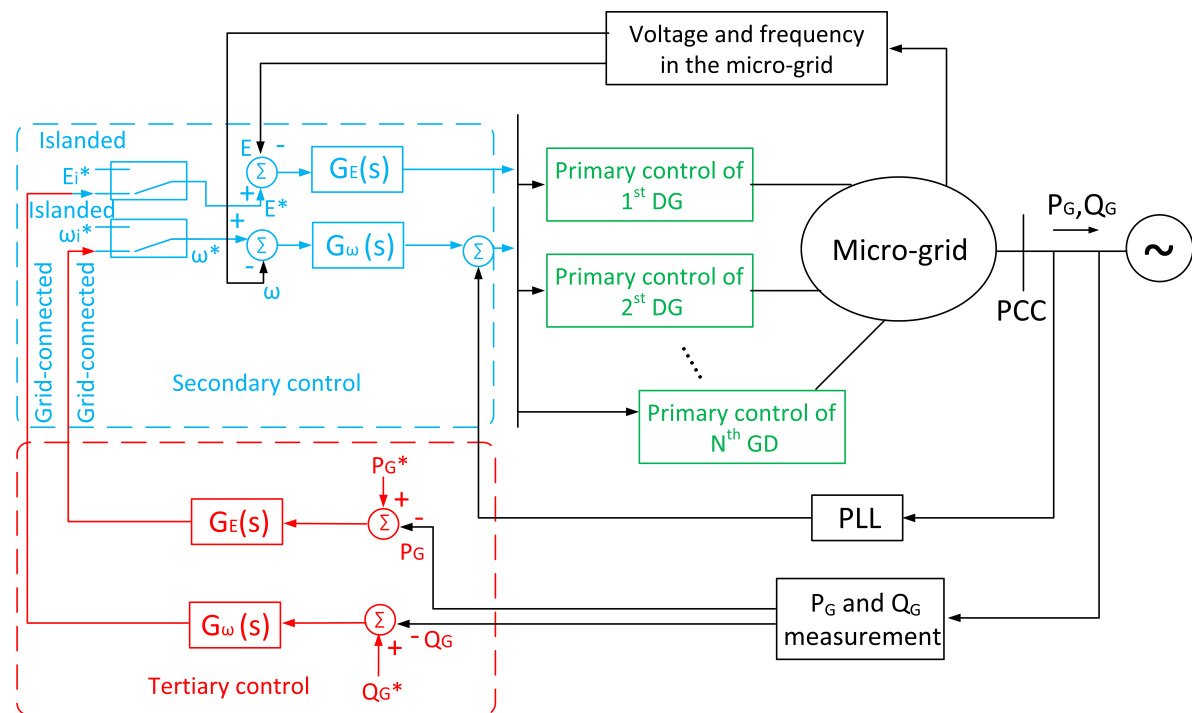


Figure 2.3: Block diagram of control levels in micro-grids.

et al. 2013). Thus, the secondary control is a control with slower dynamic response, which compensates deviations originating in the primary control.

In the centralized approach, however, the secondary level is responsible for controlling power sharing between generator and micro-grid.

2.4.4 Tertiary Level

Tertiary control is the latest and slowest control level, which considers economic issues in the optimal operation of the micro-grid and manages the power flow between micro-grid and main grid. In the grid-connected mode, the tertiary control can manage the power flow between the main and the micro-grid by means of voltage and frequency adjustments of the individual microsources (Bidram and Davoudi 2012). Figure 2.3 illustrates the integration between the three control levels in a block diagram.

Table 2.2 presents a synthesis of the functions assigned to each control level in centralized and decentralized architectures (Brandão 2015).

2.5 Distributed Generation Requirements

This section aims to present the main distributed generation requirements based on the IEEE 1547 Standard, a recommendation for Interconnection and Interoperability of Distributed Energy Resources with Associated Electric Power Systems Interfaces.

2.5.1 Cease to Energize Performance requirement

The "Cease to energize" definition is: the cessation of active power delivery under steady-state and transient conditions and limitation of reactive power exchange. This does not necessarily imply, nor

Table 2.2: Control levels and their functions in micro-grids.

Control level	Decentralized	Centralized
PWM	Voltage and current control	Voltage and current control
Primary	Inverter output control, voltage and frequency stability, Islanded detection, MPPT, State of charge control (SOC) Power sharing	Inverter output control, voltage and frequency stability, Islanded detection, MPPT, State of charge control (SOC)
Secondary	Compensation of voltage and frequency deviations	Power sharing
Tertiary	PCC power flow control and power quality	PCC power flow control and power quality

exclude disconnection, isolation, or trip. In this situation, however, limited reactive power exchange may continue as specified, e.g., through filter banks.

For reactive power exchange, there are two different rules for power rating levels:

- for local electrical power systems (EPS) with aggregate distributed energy resources (DER) rating less than 500 kVA, the reactive power exchange shall be less than 10% of nameplate DER rating;
- for local EPS with aggregate DER rating ≥ 500 kVA, the reactive power limit is 3% of nameplate DER rating.

In both cases, the reactive power shall exclusively result from passive devices.

Cease to energize requirements include import of active and reactive power exchange only for continuity of supply to DER, housekeeping and auxiliary loads.

2.5.2 Control Capability Requirements

Control capability requirements can be classified as the capability to disable permit service, capability to limit active power and execution of mode or parameter changes. These three requirements can be summarized as:

- capability to disable permit service: it is the DER capability of cease to energize the Area EPS and trip in no more than 2s.
- capability to limit active power: the DER shall limit its active power output to equal or less than the active power limit set point in no more than 30s or in the time it takes for the primary energy source to reduce its active power output to achieve that limit, whichever is greater.
- execution of mode or parameter changes: transitions between modes shall need to start until 30s after the mode setting change is received, and changes of control functional modes need to be executed in a smooth transition over a time period between 5 and 300s.

2.5.3 Reactive Power Capability and Voltage/Power Control Requirements

This section specify the requirements applied to the continuous operation region when the voltage is between 0.88 and 1.1 times the nominal voltage.

For reactive power capability, the DER shall be capable of injecting reactive power (over-excited) and absorbing reactive power (underexcited) for active power output levels greater than or equal to the minimum steady-state active power capability, or 5% of rated active power of the DER, whichever is greater.

The DER shall operate in one of the following modes: constant power factor mode, voltage-reactive power mode, active power-reactive power mode and constant reactive power mode. The constant power factor mode operating with unity power factor is the default mode. The maximum DER response time to maintain constant power factor shall be 10s or less. In the voltage-reactive power mode, the DER controls the reactive power output as a function of voltage in accordance with parameter values specified in the recommendation.

In the active power-reactive power mode, the DER controls the reactive power output as a function of the active power output following a target piecewise linear active power-reactive power characteristic. Finally, in the constant reactive power mode the DER maintains a constant reactive power according the specified by the Area EPS operator and its nominal characteristics.

2.5.4 Response to Grid Abnormal Conditions

Regarding to abnormal conditions, the IEEE Std 1547-2018 defines mandatory voltage and frequency tripping requirements, and voltage and frequency disturbance ride-through requirements.

The tripping requirements are specified by tables for each DER category. The description of each category is explained below, as transcribed from the IEEE recommendation.

- Abnormal operating performance Category I is based on essential bulk power system (BPS) stability/reliability needs and reasonably attainable by all DER technologies that are in common usage nowadays.
- Abnormal operating performance Category II covers all BPS stability/reliability needs and is coordinated with existing reliability standards to avoid tripping for a wider range of disturbances of concern to BPS stability.
- Abnormal operating performance Category III is based on both BPS stability/reliability and distribution system reliability/power quality needs and is coordinated with existing interconnection requirements for very high DER penetration.

Table 2.3 summarizes the tripping requirements for under and overvoltage for DER of abnormal operating performance category I. For category II and III, there are some different values, but the tables are not presented here to avoid repetitive tables.

Table 2.3: DER response (shall trip) to abnormal voltages for DER of abnormal operating performance Category I.

Shall trip—Category I				
Shall trip function	Default settings		Ranges of allowable settings	
	Voltage (p.u.)	Clearing time (s)	Voltage (p.u.)	Clearing time (s)
OV2	1.20	0.16	fixed at 1.20	fixed at 0.16
OV1	1.10	2.0	1.10 - 1.20	1.0 - 13.0
UV1	0.70	2.0	0.0–0.88	2.0–21.0
UV2	0.45	0.16	0.0–0.50	0.16–2.0

Voltage ride-through requirements are shown in Table 2.4 and define if the operation mode in a given voltage range is "cease to energize", "permissive operation", "mandatory operation" or "continuous operation". One more time it was chosen category I for illustrate the recommendation.

Table 2.4: Voltage ride-through requirements for DER for abnormal operating performance Category I

Voltage range (p.u.)	Operating mode/ response	Minimum ride-through time (s)	Maximum response time (s)
$V > 1.20$	Cease to Energize	N/A	0.16
$1.175 < V \leq 1.20$	Permissive Operation	0.2	N/A
$1.15 < V \leq 1.175$	Permissive Operation	0.5	N/A
$1.10 < V \leq 1.15$	Permissive Operation	1	N/A
$0.88 \leq V \leq 1.10$	Continuous Operation	Infinite	N/A
$0.70 \leq V < 0.88$	Mandatory Operation	T_{VRT}	N/A
$0.50 \leq V < 0.70$	Permissive Operation	0.16	N/A
$V < 0.50$	Cease to Energize	N/A	0.16

The term T_{VRT} is a linear curve presented by:

$$T_{VRT} = 0.7s + \frac{4s}{1p.u.} (V - 0.7p.u.) \quad (2.1)$$

There are also mandatory frequency tripping requirements. When the system frequency is in a given range, and the fundamental-frequency component of voltage on any phase is greater than 30% of nominal, the DER shall cease to energize and trip within a clearing time as indicated in Table 2.5.

Table 2.5: DER response (shall trip) to abnormal frequencies for DER of abnormal operating performance Category I, II, and III

Shall trip function	Default settings		Ranges of allowable settings	
	Frequency (Hz)	Clearing time (s)	Frequency (Hz)	Clearing time (s)
OF2	62.0	0.16	61.8–66.0	0.16–1000
OF1	61.2	300.0	61.0–66.0	180.0–1000
UF1	58.5	300.0	50.0–59.0	180.0–1000
UF2	56.5	0.16	50.0–57.0	0.16–1000

Regarding to the frequency ride-through, the operating mode in a full-range of frequency variation can be classified in "continuous operation", "mandatory operation" and "no ride-through requirements apply", according to a range of values specified in the Table 2.6.

Table 2.6: Frequency ride-through requirements for DER of abnormal operating performance Category I, II, and III

Frequency range (Hz)	Operating mode	Minimum time (s)
$f > 62.0$	No ride-through requirements apply to this range	
$61.2 < f \leq 61.8$	Mandatory Operation	299
$58.8 \leq f \leq 61.2$	Continuous Operation	Infinite
$57.0 \leq f < 58.8$	Mandatory Operation	299
$f < 57.0$	No ride-through requirements apply to this range	

2.5.5 Power Quality

The Power Quality section defines limits of DC injection, where the DER shall not inject DC current greater than 0.5% of the full rated output current, voltage fluctuations induced by the DER, current distortion and overvoltage contribution. These limits are explained in detail in the IEEE 1547-2018, but not discussed here for simplification.

2.6 Active and Reactive Power Injection Strategies for Grid-connected Converters during voltage sags

In the section 2.5, distributed generation requirements were presented, including rules for reactive power injection and response to abnormal conditions. These rules demands power injection strategies that can be divided into: constant average active power, constant active current and constant peak current.

In the normal operation mode, the active power reference is the output of the maximum power point tracking (MPPT) algorithm P_{mppt} , and the system shall operate at unity power factor. When a voltage sag is detected, the DG runs into the LVRT operating mode, and the system is required to withstand the voltage drop for a specific period of time. Simultaneously, the converter injects a specific amount of reactive power to support the grid voltage recovery. Figure 2.4 shows the relationship between the voltage drop and reactive current injection, split into three operating ranges, as

$$\begin{cases} (I_{rm}/I_N) = 0 & V_{p.u.} > 0.9 \\ (I_{rm}/I_N) = k \cdot (1 - V_{p.u.}) & 0.9 \geq V_{p.u.} > 0.5 \\ (I_{rm}/I_N) = 1 & V_{p.u.} \leq 0.5 \end{cases} \quad (2.2)$$

where (I_{rm}/I_N) is the ratio of peak value of reactive current injection per m-phase to the nominal current of PES reflected to ac side, $I_N = 2.P_N / (3.V_N)$; $V_{p.u.}$ is the per-unit value of the grid voltage during the disturbance from its nominal value, per phase; and k is a constant given by $k = 2$. Then, three different operating ranges are set: 1) the normal operation from 0.9 to 1.1 p.u.; 2) the second range is set between 0.5 and 0.9 p.u.; and 3) the third range for voltage values lower than 0.5 p.u.

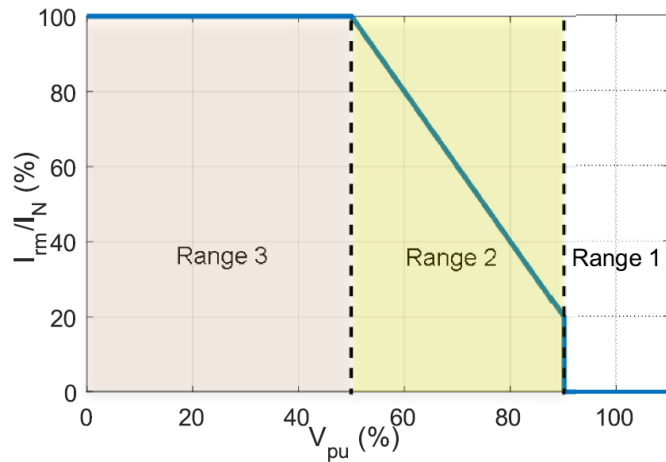


Figure 2.4: Required reactive current injection during the three different operating ranges: Range 1 with absolute value of voltage higher than 0.9 p.u.; range 2 with voltage between 0.5 and 0.9 p.u.; and range 3 with voltage value lower than 0.5 p.u.

2.6.1 Constant Average Active Power

To maximize the active power injection of DGs, this strategy keeps the average active power constant under the disturbance. Then, the references of active (i_{am}^*) and reactive (i_{rm}^*) current during the voltage sag are given as following. Such that the variables v_{pllm} and $\hat{v}_{pllm}$ shape the current reference in-phase and in-quadrature to the m-phase voltage, once the inverter control scheme is devised in abc-frame.

Range 1 - $0.9 \text{ p.u.} \leq V_m < 1.1 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} \frac{1}{V_{pu}} v_{pllm} \\ i_{rm}^* = 0 \end{cases} \quad (2.3)$$

Range 2 - $0.5 \text{ p.u.} \leq V_m < 0.9 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} \frac{1}{V_{pu}} v_{pllm} \\ i_{rm}^* = k.(1 - V_{pu}) \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.4)$$

Range 3 - $V_m < 0.5 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} \frac{1}{V_{pu}} v_{pllm} \\ i_{rm}^* = \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.5)$$

2.6.2 Constant Active Current

The objective of this control strategy is provide constant active current during voltage sag. Then, the references of active and reactive current are given as:

Range 1 - $0.9 \text{ p.u.} \leq V_m < 1.1 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} v_{pllm} \\ i_{rm}^* = 0 \end{cases} \quad (2.6)$$

Range 2 - $0.5 \text{ p.u.} \leq V_m < 0.9 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} v_{pllm} \\ i_{rm}^* = k.(1 - V_{pu}) \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.7)$$

Range 3 - $V_m < 0.5 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} v_{pllm} \\ i_{rm}^* = \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.8)$$

2.6.3 Constant Peak Current

This control strategy establishes constant magnitude of the injected current during the voltage sag. Then, the references of active and reactive current are given as following:

Range 1 - $0.9 \text{ p.u.} \leq V_m < 1.1 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \frac{2(P_{mppt}/3)}{V_N} v_{pllm} \\ i_{rm}^* = 0 \end{cases} \quad (2.9)$$

Range 2 - $0.5 \text{ p.u.} \leq V_m < 0.9 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = \sqrt{\left(\frac{P_{mppt}/3}{P_N}\right)^2 - k^2(1 - V_{pu})^2} \cdot \frac{2P_N}{V_N} v_{pllm} \\ i_{rm}^* = k.(1 - V_{pu}) \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.10)$$

Range 3 - $V_m < 0.5 \text{ p.u.}$:

$$\begin{cases} i_{am}^* = 0 \\ i_{rm}^* = \frac{2P_N}{V_N} \hat{v}_{pllm} \end{cases} \quad (2.11)$$

2.6.4 Simulation results

A three-phase four-wire inverter is simulated in Matlab/Simulink to demonstrate the performance of the strategies above described under symmetrical/asymmetrical voltage sag condition. The parameters of the power circuit are shown in Table 2.7.

Table 2.7: Parameters for simulation of active / reactive power injection strategies.

Three phase voltage grid	220 V / 60Hz
DC link voltage	400 V
Grid impedance	$L_{gm} = 60 \mu H$; $R_{gm} = 0.12 \Omega$
Output filter capacitor	$C_{fm} = 5 \mu F$; $R_{fm} = 0.01 \Omega$
Output filter inductor	$L_{im} = 1.5 mH$; $R_{im} = 0.05 \Omega$
Switching frequency	$f_{sw} = 12 \text{ kHz}$
Nominal power of PES	3x4 kW (12 kW)
Rated power of inverter	3x10 kVA (30 kVA)

A. Symmetrical voltage sags

During symmetrical voltage sags, the reactive current injection shall be set according to the sag magnitude following the ranges 1 to 3. The inverter voltage and current waveforms for the three strategies under each operating range are shown in Figures 2.5-2.7. The grid voltage in the simulation varies according the following: voltage range 2 = 0.75 p.u. at $t = 0.168\text{s}$ and voltage range 3 = 0.45 p.u. at $t = 0.234\text{s}$).

For the three strategies, before the voltage sag, i.e., range 1, the inverter injects only active power and the system operates with unity power factor. Once the voltage sag is detected, i.e., range 2 and 3, under constant average active power control, the injected active current increases and the system starts, simultaneously, providing reactive current. It can be seen from the amplitude risen of the injected current and the fact that currents are shifted from the phase voltages, as in Figure 2.5. Under constant

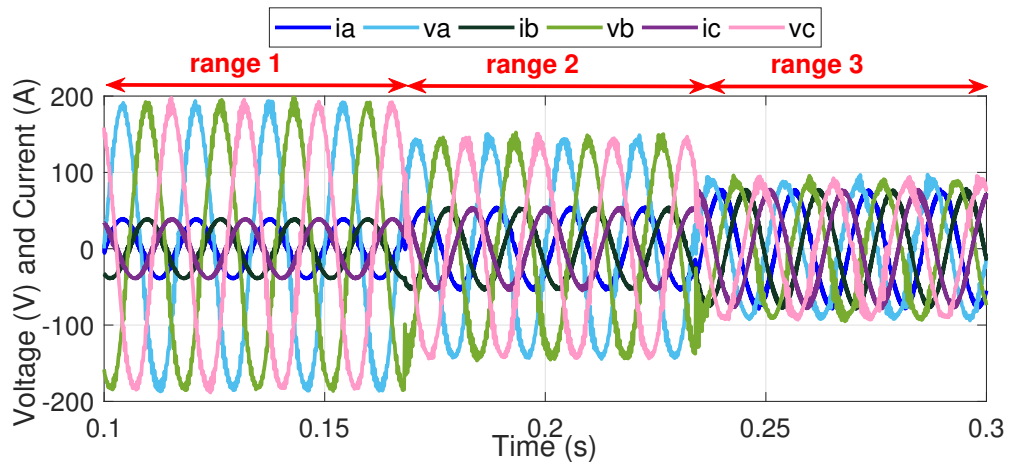


Figure 2.5: Constant average active power strategy.

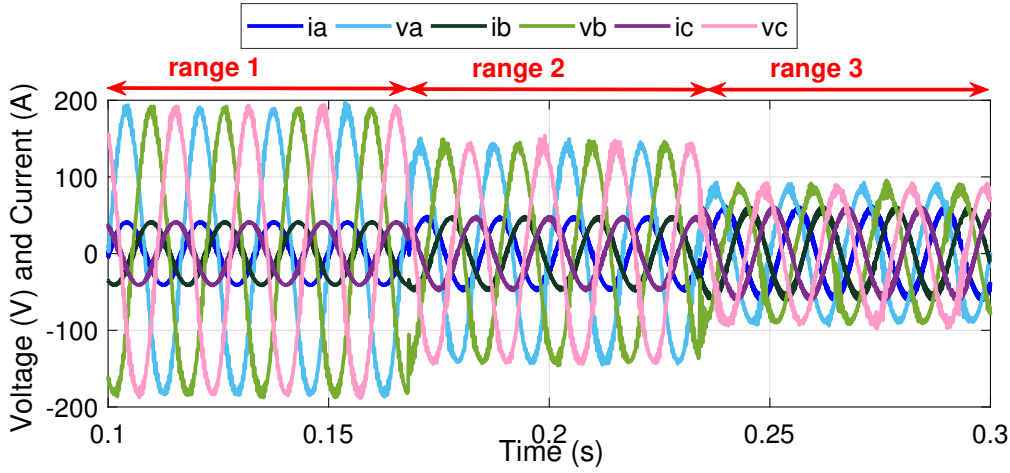


Figure 2.6: Constant active current strategy.

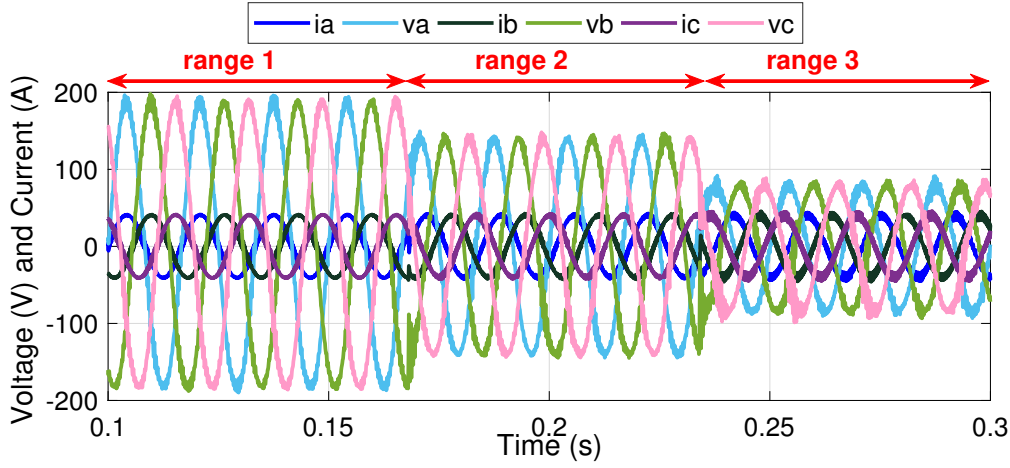


Figure 2.7: Constant peak current strategy.

active current control strategy, the injected currents show the same behavior as before for constant average active power control, but with smaller amplitude increase, as shown in Fig. 2.6. From Fig. 2.7, it is observed that the current magnitudes are kept constant during the voltage sag, since constant peak current control is applied. However, currents are shifted from phase voltages indicating the injection of reactive power into the grid. Consequently, the injected active current must be reduced to guarantee constant peak current.

Figures 2.8-2.10 evaluates the proposed algorithms showing the behavior of the controlled quantities for each injection strategy, as following: 1) constant average active power, 2) constant active current, and 3) constant peak current. Observe that each strategy keeps its controlled quantity constant during the symmetrical voltage sags. Figure 2.8 demonstrates that full active power is injected under voltage sags with constant average active power control. The other two strategies must reduce their active power to ensure constant active current and constant peak current while complying with (2.2). Figure 2.9 shows the magnitude of the collective value of active currents, while Figure 2.10 shows the magnitude of the collective value of peak currents, for all strategies. From Figures 2.8-2.10, it is clear that there exists a trade-off between the active power injection and the current magnitude increasing during the voltage sags, which must be safely controlled to avoid overcurrent.

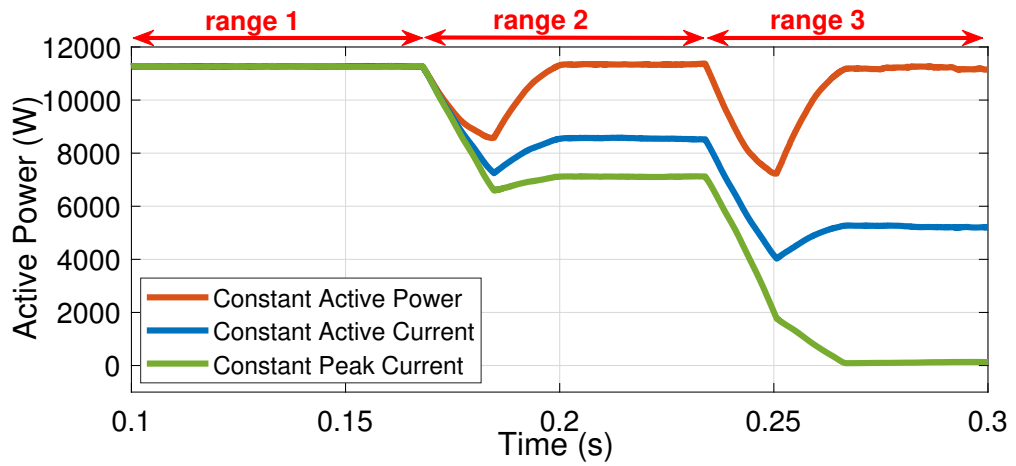


Figure 2.8: Active power behavior for the three active/reactive power injection strategies (voltage range 2 = 0.75 p.u. at $t=0.168$ s and voltage range 3 = 0.45 p.u. at $t=0.234$ s).

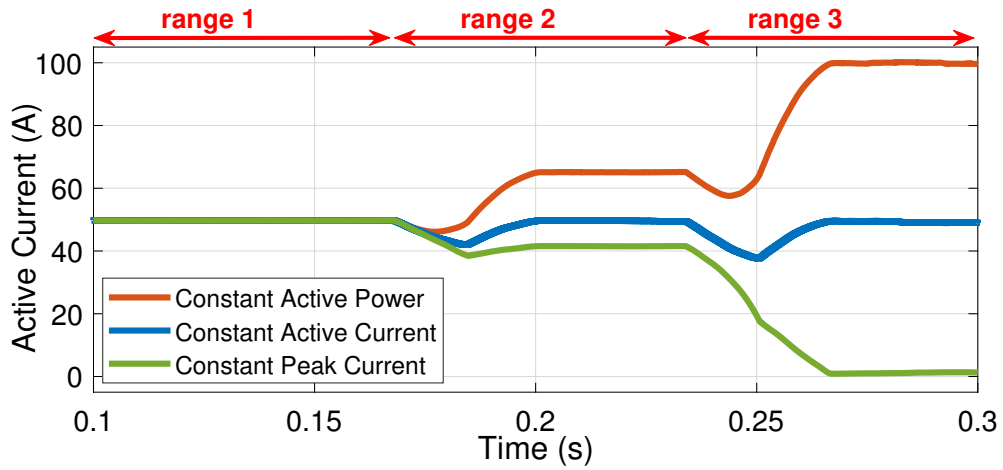


Figure 2.9: Active current behavior for the three active/reactive power injection strategies.

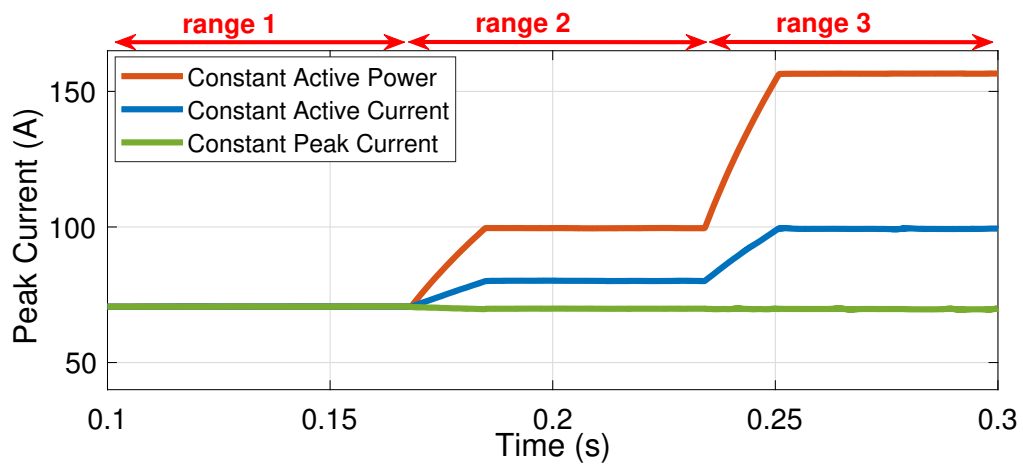


Figure 2.10: Peak current behavior for the three active/reactive power injection strategies.

B. Asymmetrical voltage sags

Figures 2.11-2.13 show the response of the DG system when submitted to a type E voltage sag, two phases (phases b and c) drop to 50% of the nominal voltage, while the other one remains at nominal condition. By definition, the reactive current is injected only into the phases that suffer from voltage sag. It requires the three-phase four-wire inverter be properly designed to operate unbalanced avoiding heavy DC-link voltage oscillation.

The system behaviour under asymmetrical voltage sag is quite similar to that observed under symmetrical voltage sags, since the control scheme is devised in abc-frame and for the three-phase four-wire system, the grid phases are decoupled from each other. The exception is the DC-link regulation, in which the DC-voltage oscillates because of instantaneous active and reactive power oscillation at the grid side due to unbalanced currents.

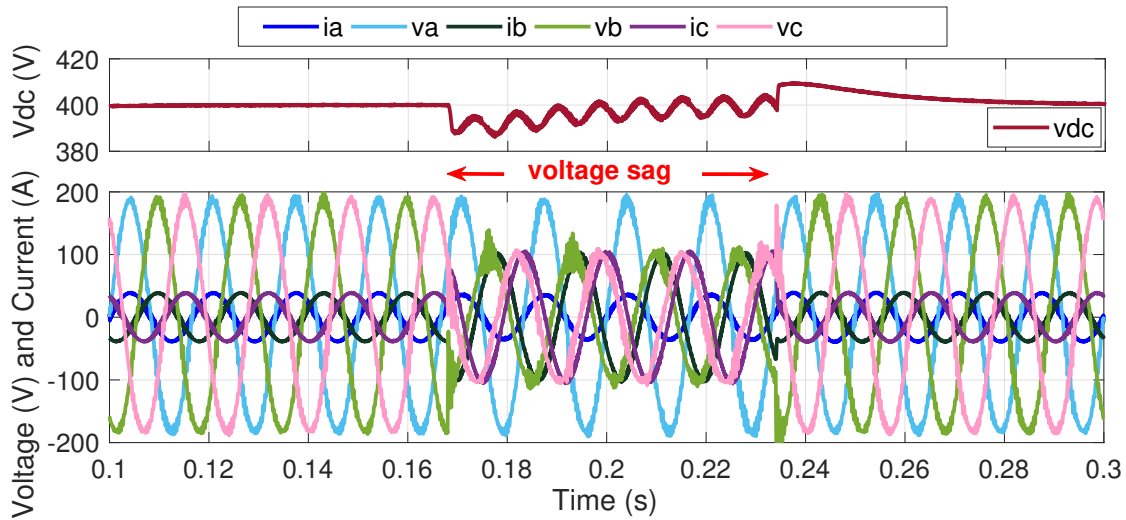


Figure 2.11: Constant average active power strategy under unbalanced voltage sag.

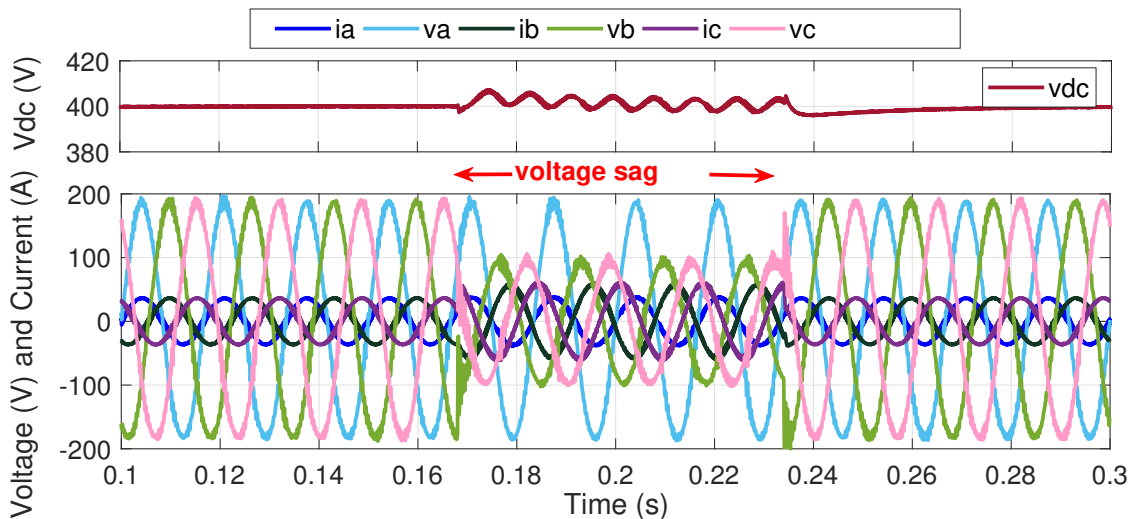


Figure 2.12: Constant active current strategy under unbalanced voltage sag.

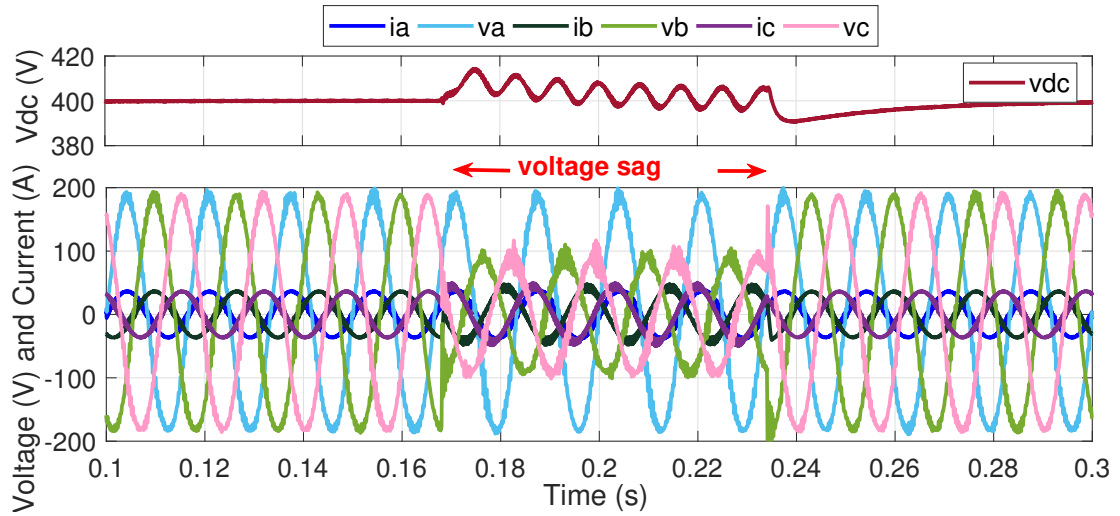


Figure 2.13: Constant peak current strategy under unbalanced voltage sag.

2.7 Conclusions

In this chapter, fundamental concepts were presented to the understanding of the power converters operation applied to distributed generation and micro-grids, which will be frequently used in the following chapters.

The operation modes clarified the control challenges related to the absence of a robust main source in isolated mode and the need for grid synchronization in the connected mode. The concern with mode transitions is also pointed out in the text.

The classification of the converters according to the function allowed to understand that the control methods can vary, for example if the converter is a grid-forming or a grid-feeding.

The explanation of the hierarchical control levels in micro-grids served to contextualize that the subject are not limited to one or two control loops, but a set that includes from the voltage and current control to management decisions related to the dispatch of power.

The distributed generation requirements was a literature review based on IEEE 1547-2018 and help us to understand for example the acceptable limits for voltage and frequency variations to which the converter must continue to operate, among other features, such as control capability and reactive power capability requirements.

In the next chapter, active and reactive power control techniques for grid-connected converters emulating the synchronous machines are presented.

3 Power Control Techniques Based on Synchronous Machines Emulation

The present chapter aims to study control techniques for power converters that emulate synchronous machines. The droop control is explained by being a widespread and easy implementation technique. Then, the strategies that reproduce more completed models of the synchronous machine are presented. A comparison between the techniques discussed in the chapter is performed through an analysis of mathematical models and simulations.

3.1 Introduction

Figure 3.1 illustrates the connection between two generators in parallel and the phasor diagram relating the electrical voltages and the resulting current. The complex power flowing between two sources separated by an inductive reactance can be divided into a real and an imaginary part, and corresponds to the conjugate complex of the current multiplied by the voltage at the sources, according to equation 3.1. Using the voltage magnitudes (V_r and V_s) and the phase angle δ , the equation (3.1) becomes (3.2), where the real part of the complex power S is known as the active power P and the imaginary part is called reactive power Q .

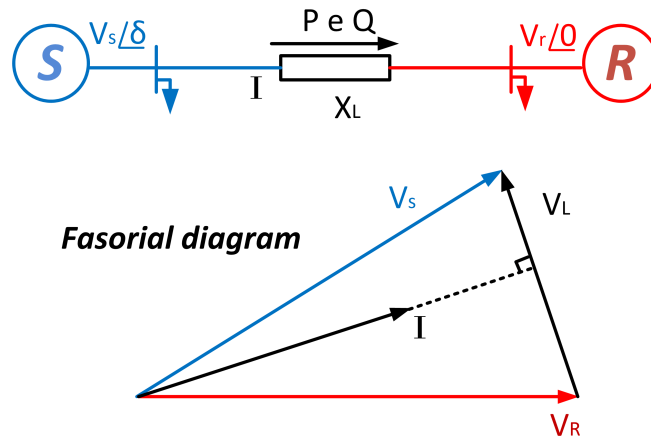


Figure 3.1: Generators connected through a coupling impedance.

$$S = P + jQ = \vec{V}_r \cdot \vec{I} \quad (3.1)$$

$$S = \frac{V_r V_s \sin(\delta)}{X_L} + j \frac{V_r V_s \cos(\delta) - V_r^2}{X_L} \quad (3.2)$$

The power control, therefore, can be obtained by adjusting the relative angle between the voltages or by small variations in the magnitude of the voltage in one of the sources. To date, the layout, properties, operating strategies of the grid, as well as its dynamic behavior are closely linked to the static and dynamic properties of the electromechanical synchronous machines (Beck and Hesse 2007). This statement has supported the development of control techniques that seek to make interface converters between primary and secondary power sources to have identical or at least similar behavior as synchronous machines and this concept has become a very active research subject in recent years (Hogan et al. 2016).

But why try to emulate a synchronous machine with electronic converter? First, because the control variables take physical meaning and consequently, the effect of the variation of these parameters on the electrical quantities becomes predictable. In addition, the possibility of virtual parameter adjustments, in favor of the converter stability and the minimization of transients due to disturbances in the system, represents an advantage over other techniques. Finally, it is understood that the parallelism of generators based on electronic power converters becomes, from the point of view of control, similar to the model so consolidated in electric systems with synchronous generators. Thus, in a future perspective, systems based on converters emulating a synchronous machine are recognized as an effective method to solve stability problems related to decentralized generation (D'Arco and Suul 2014).

In this line, Virtual Synchronous Machine (VISMA), Virtual Synchronous Generator (VSG) and Synchroverter are the techniques found. The present section sets out to briefly explain each of the three techniques.

3.2 Droop Control

The droop control is considered the most widespread (Wandhare, Thale, and Agarwal 2014) method for implementing the active and reactive power sharing control among inverters that integrate an autonomous micro-grid (Palizban and Kauhaniemi 2013). By means of this technique, which eliminates the need of communication systems between the micro generators, the magnitude and frequency of the output voltage are controlled by the reactive and active power injection respectively, as described in (3.3) and (3.4):

$$f = f^* - k_p \cdot P - m_d \frac{dP}{dt} \quad (3.3)$$

$$V_n = V^* - k_q \cdot Q - n_d \frac{dQ}{dt} \quad (3.4)$$

where f^* and V^* are the frequency and magnitude of the output voltage, k_p e k_q are the droop coefficients, P and Q the average active and reactive powers, and m_d and n_d are the derivative droop coefficients. The derivative terms absorb abrupt changes and preserve the stability of the power sharing (Oureilidis and Demoulias 2013). In many references, however, this derivative term is not used (Ding and Lo 2012 Dobakhshari, Azizi, and Ranjbar 2011 Dong et al. 2014 Bevrani and Shokoochi 2013 Barklund et al. 2008 Alotibe, Chandrasena, and Rajakaruna 2012 Brabandere et al. 2007a Luo et al. 2012 Guerrero et al. 2009 Lopes, Moreira, and Madureira 2006 Mohamed and El-Saadany 2008 Quesada et al. 2014 Lu and Chu 2013), restricting the algorithm to the proportional action. This simplified approach is considered throughout this work.

Proportional gains k_p e k_q can be calculated according to equations 3.5 e 3.6.

$$k_p = \frac{\Delta\omega}{P_n} \quad (3.5)$$

$$k_q = \frac{\Delta V}{Q_n} \quad (3.6)$$

where, P_n e Q_n are active and reactive powers and $\Delta\omega$ and ΔV are frequency and voltage variations, which will produce those nominal powers.

The active and reactive powers are calculated in three-phase natural coordinates or in dq references, obtained from three-phase output voltage and current measurements (Lee, Chu, and Cheng 2013). In grid-connected mode, the control strategy consists of controlling the power to be injected at the coupling point to ensure maximum power transfer (Vandoorn et al. 2013b). In the islanded mode, the control acts in order to maintain fixed voltage and frequency, and the injected power is dictated by the load impedance characteristics. Figure 3.2 illustrates droop control for the reference voltage determination.

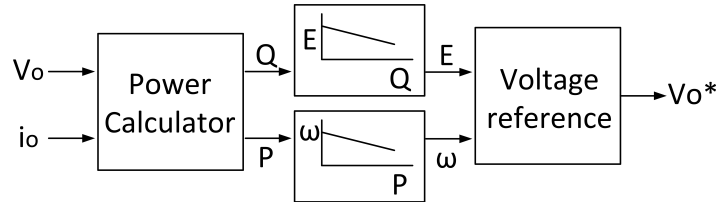


Figure 3.2: Control of the reference voltage in the voltage control mode, through the droop control.

For applications in low voltage grids, which are typically resistive in nature, it is recommended to use the inverted droop control, where the magnitude of the voltage is associated with the active power control and the frequency regulation to the reactive power (Basak et al. 2009). Besides that, the slope of the droop curves are positive for resistive grids.

After the first approach, many improvements were developed, focusing on better performances in transient time, enhance fault-ride through during grid-disturbances conditions and to increase reliability of the converters, either in connected mode or stand-alone. In the traditional droop control, the equations link variations of frequency with the injected active power and variations of voltage magnitude with the injected reactive power control. This approach is feasible when the line impedance is predominantly inductive, while in resistive lines, frequency is associated with reactive power and voltage is associated with active power. For X/R near to 1, however, these simplest approaches are not effective in transient time. This problem is addressed in Brabandere, where power angle and voltage magnitude influences the active and reactive power simultaneously. Equation (3.7) presents the transformation matrix where P' and Q' are the new power references, and R , X and Z are the resistance, the reactance and the impedance estimated from the grid.

$$\begin{bmatrix} P' \\ Q' \end{bmatrix} = \begin{bmatrix} \sin \theta & -\cos \theta \\ \cos \theta & \sin \theta \end{bmatrix} \begin{bmatrix} P \\ Q \end{bmatrix} = \begin{bmatrix} \frac{X}{Z} & -\frac{R}{Z} \\ \frac{R}{Z} & \frac{X}{Z} \end{bmatrix} \begin{bmatrix} P \\ Q \end{bmatrix} \quad (3.7)$$

Another issue studied throughout the droop control evolution is the addition of a virtual impedance, first introduced to shape desired output impedances in uninterruptible power systems (Guerrero et al. 2005). The idea is that a proper design of the virtual output impedance in droop-controlled inverters can reduce the impact of the line-impedance unbalance.

Virtual output impedance loop is considered an intermediate stage that adjust the output impedance of the converter, by subtracting a portion of the output current from the voltage reference (Guerrero et al. 2007). Equation (3.8) represents the described alternative.

$$v^* = v_{ref} - Z_v(s).i_o \quad (3.8)$$

in which, v^* is a new voltage reference with the virtual impedance, v_{ref} is the voltage reference generated from the power control loop, Z_v is a virtual output impedance function and i_o is the measured output current.

Figure 3.3 illustrates the block diagram of the closed-loop system with the virtual output impedance. The $Z_v(s)$ function can adopt resistive behavior if implemented as a constant or assume a reactive behavior, depending on the chosen transfer function. Another possibility is to use a discrete Fourier transform to program different impedance over the current harmonics. Many papers present variations of the virtual impedance implementation (Guerrero et al. 2007; Vasquez et al. 2009; Wang, Blaabjerg, and Chen 2012; Ye et al. 2017).

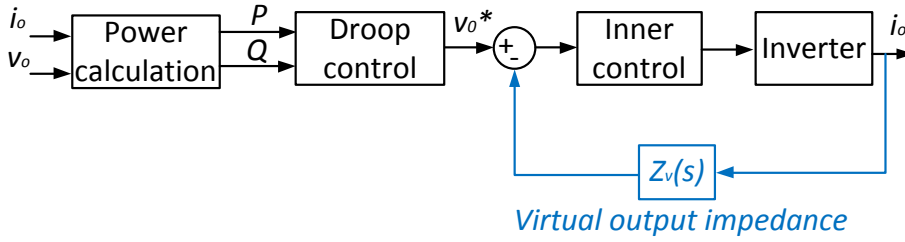


Figure 3.3: Block diagram of the closed-loop system with the virtual output impedance.

A generalized droop control (GDC) for a grid-supporting inverter is presented in (Meng, Liu, and Liu 2019) and is created based on the comparison between VSG control and traditional droop control. Figure 3.4 shows the power control of GDC scheme. It is possible to observe that the reactive power loop is identical to the traditional droop control. On the other hand, three new gains are introduced in the active power and are clearly inspired by the VSG: G_{PLC} is the controller in the forward channel, which can be designed flexibly and not be limited to a droop coefficient with a first-order low-pass filter, G_{CPF} is the secondary controller and H_{CPF} is the compensator in the active power feedback channel. Compared with the VSG control, the GDC can suppress the oscillation and reduce the settling time of the output active power when the power reference changes. In the stand-alone mode, compared to droop control, the GDC can provide inertia and damping properties. In summary, GDC can be equivalent to either the traditional droop or the VGS controlled converters, depending on the design of the control parameters.

A variation of the droop control technique presented in this section can be found in DC micro-grids, named voltage droop control, wherein the voltage in the DC micro-grid is regulated by adjusting the current flowing through the converters. In this approach, inspired by conventional droop control, the implementation is also accomplished through a proportional controller. Voltage droop control is reported to be the most commonly used technique in DC multi-terminal grids (Rouzbehi et al. 2013).

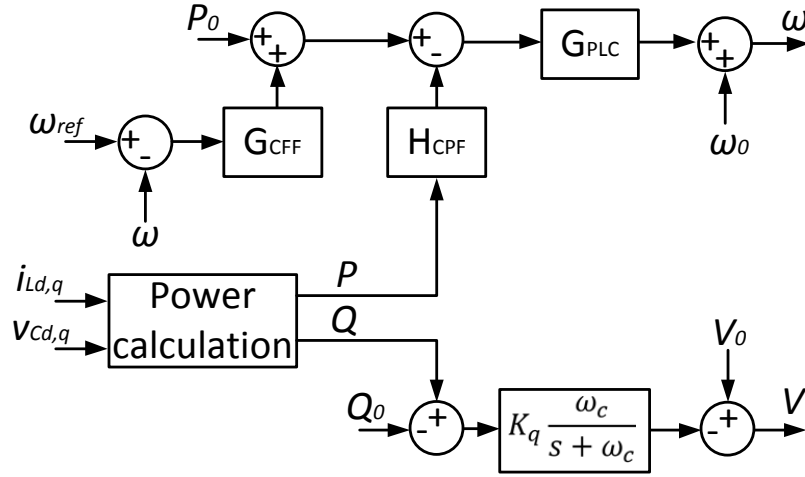


Figure 3.4: Power control loop of GDC scheme.

3.3 VISMA

The virtual synchronous machine, or simply VISMA (Beck and Hesse 2007), is the first control design that combines the advantages of electronic power converter technologies with the static and dynamic characteristics of electromechanical synchronous machines. If there is a storage system connected to the DC side, the VISMA can operate in all four quadrants, with the AC voltage side corresponding to the stator output of a synchronous machine.

The reproduction of the mathematical model of the synchronous machine or, in other words, a complete VISMA unit has three sub-processes: voltage measurement at the coupling point with the grid, real-time machine current calculation and current feedback. The machine parameters, which in this case are virtual, can be freely adjusted and modified in real-time as long as they guarantee system stability. The calculated machine currents are the reference values for the current that the inverter must inject into the grid.

The block diagram of the three-phase VISMA and its corresponding control algorithm can be seen in Figure 3.5 (Chen et al. 2011). The variable J is the virtual moment of inertia, k_d is the virtual friction coefficient, ω is the angular frequency, θ is the rotation angle, T_e and T_m are respectively the electric torque and the mechanical torque. The induced electromotive force (e) is given by the product between an adjustable amplitude (E_p) and three sinusoidal components displaced 120° from each other, calculated from the angle θ .

As can be observed, the control implements the equations of the synchronous machine, and the output of the algorithm represents the reference for the current control, or in other words, the current required to produce the desired power. It is also necessary an internal control loop, which generates the references for inverter switching, which is commonly performed through PWM, or alternative solutions such as hysteresis control (Frack, Mercado, and Molina 2015).

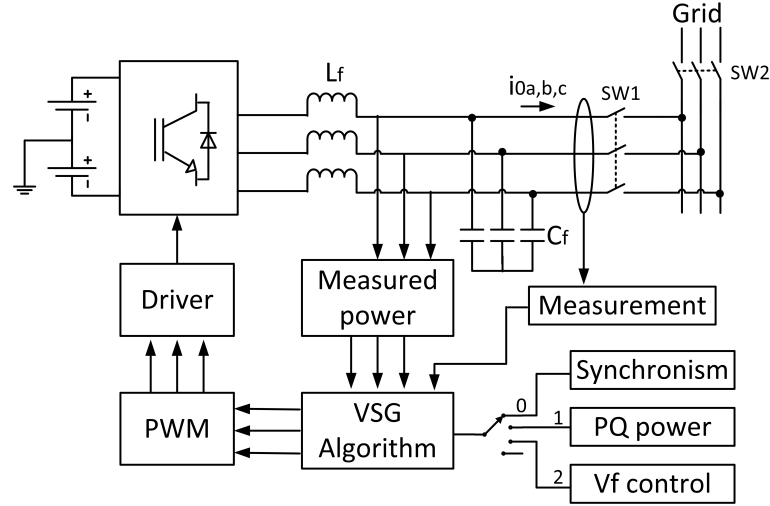


Figure 3.6: VSG scheme.

allow the closing of the switch between the two systems. This functionality prevents the occurrence of transients that shall trip protections or damage inverter components.

The PQ control is used when one desires to dispatch power to the grid according to active and reactive power set-points, applicable to grid-feeding converters. The Vf control is used to regulate the voltage and frequency. Figure 3.7 illustrates the PQ and Vf controls in block diagram.

Vf control:

The Vf algorithm is applicable to two situations: for a grid-forming VSG, in the concept of a micro-grid with a master DG and other slave generators, or to feed priority loads in case of grid disconnection, operating as an uninterruptible power source.

For the Vf control, the goal of the first control loop of Figure 3.7 is to regulate the output frequency by means of a PI controller, when the switch B_{p1} is closed. The constants K_f and T_{fi} are respectively the proportional and integral gains, and LTP is a limiter, so that the power (P_m) is not greater than the maximum output power of the inverter. The voltage control of the VSG is implemented by the second control loop of the Figure 3.7, with the switch B_{p2} closed, being analogous to the excitation control of the synchronous machine. The RMS excitation voltage of the synchronous generator is proportional to the excitation current I_f , the angular frequency ω_m and the excitation coefficient G_{af} , as shown in eq. (3.9).

$$E_0 = G_{af} \cdot I_f \cdot \omega_m \quad (3.9)$$

In Figure 3.7, the variable V^* is the voltage reference, V_c is the measured voltage at the output of the VSG, and I_{ff}^* is the current required to establish the voltage excitation of the VSG in no-load conditions. Typically $I_{ff}^*=1$. The K_v , T_{ui} and K_{vff} gains are, respectively, the proportional gain, the integral gain and the feedforward gain. The output of this control loop (E_0) together with the output variable of the frequency loop (P_m) are the inputs for the VSG algorithm to calculate the reference voltages for the PWM.

An intentional variation in the virtual inertia is proposed in (Alipoor, Miura, and Ise 2013), with the aim of reducing oscillations in frequency and power, by controlling the acceleration / deceleration term of the machine. The derivative of the angular speed represents this acceleration.

The implementation of the VSG in the context of distributed generation is fundamentally justified by the following characteristics:

- prevent grid instability and possible blackouts due to large variations caused by decentralized generation;
- favor the intentional islanding of areas with decentralized generation, in order to maintain the functioning of social structures;
- maintain system safety in case of power fault (Driesen and Visscher 2008).

Single-phase versions of the VSG are found in (Torres and Baier 2015), for CSI converters, and in (Zeng et al. 2013 and Hirase et al. 2014), for VSI converters, both operating connected to the grid.

3.5 Synchronverter

The synchronverter was first introduced in (Zhong and Weiss 2009) named static synchronous generator and two years later was called synchronverter. Synchronverters are inverters that mimic synchronous generators and consists of a power part and a control part (Zhong and Weiss 2011). They are modeled for three-phase systems, as shown in Figure 3.9, where a primary power source is connected to the DC bus through an auxiliary converter, and this in turn, to the three-phase converter that controls the power flow between generator and grid. For the study of the synchronverter, it is assumed that the voltage on the DC bus is constant (Aouini et al. 2014).

A passive LC output filter is required for the connection of converter to the grid, setting the current ripple and attenuating the harmonic distortions from the converter. The intrinsic output resistance of the filter influences the stability of the synchronverter, which can be improved by using a virtual resistance added to the synchronverter control model (Piya and Karimi-Ghartemani 2016).

A control switch is used to ensure that the generation system and the grid are connected only after they are synchronized, avoiding transients that may cause protection actuation or component damage.

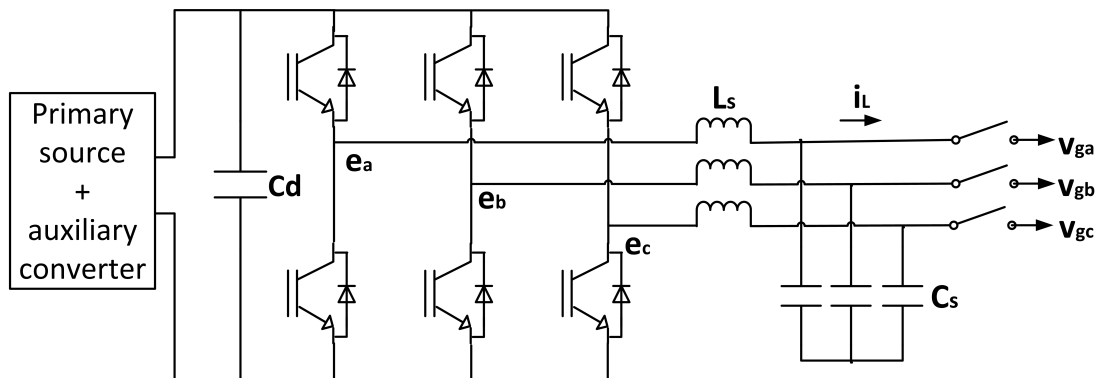


Figure 3.9: Schematic diagram of the power part of the synchronverter.

The control part of the synchronverter can be understood as a system that emulates the behavior of a synchronous generator, as shown in Figure 3.10. The block diagram shows that the set-points are the active power (P^*) and reactive power (Q^*) references, while the output quantity is the voltage needed to achieve the desired powers.

The active power delivered to the grid is obtained by adjusting the frequency through a control loop called frequency droop which can be implemented in the synchronverter by comparing the virtual angular speed ω with the angular frequency reference ω_r , which is the grid frequency in the connected mode or a fixed set-point value during the islanded operation. When the active power reference is increased, the speed of the synchronous generator is reduced due to the increase of the electromagnetic torque, T_e .

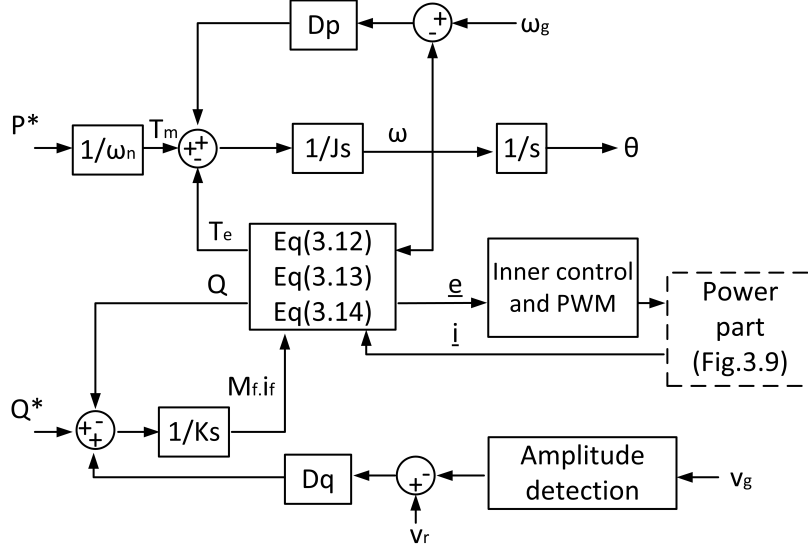


Figure 3.10: Control diagram of the synchronverter.

Equations (3.12), (3.13) and (3.14) are, respectively, the electromagnetic torque, the output voltage and the reactive power. Underlined quantities indicate that these are three-phase vectors.

$$T_e = M_f i_f \langle \underline{i}, \underline{\text{sen}\theta} \rangle \quad (3.12)$$

$$\underline{e} = \omega M_f i_f \underline{\text{sen}\theta} \quad (3.13)$$

$$Q = -\omega M_f i_f \langle \underline{i}, \underline{\text{cos}\theta} \rangle \quad (3.14)$$

where, M_f represents the mutual inductance, i_f represents the rotor excitation current, $\langle \cdot, \cdot \rangle$ denotes the internal product in R3, $\underline{\text{sen}\theta}$ and $\underline{\text{cos}\theta}$ are:

$$\underline{\text{sen}\theta} = \begin{bmatrix} \text{sen}\theta \\ \text{sen}(\theta - 2\pi/3) \\ \text{sen}(\theta - 4\pi/3) \end{bmatrix}, \underline{\text{cos}\theta} = \begin{bmatrix} \text{cos}\theta \\ \text{cos}(\theta - 2\pi/3) \\ \text{cos}(\theta - 4\pi/3) \end{bmatrix} \quad (3.15)$$

The variable D_p can be interpreted as the damping factor of the synchronous machine, while J is the virtual moment of inertia. Parameters D_p and J can be obtained through (3.16) and (3.17), respectively.

$$D_p = -\frac{\Delta P}{\omega_n \Delta \omega} \quad (3.16)$$

$$J = D_p \tau_f \quad (3.17)$$

where ΔP is the total active power variation, it can also be understood as the nominal power of the converter, $\Delta\omega$ is the variation of the needed angular frequency for the active power to vary from zero to its nominal value and τ_f is the time constant of the frequency droop.

The reactive power Q is similarly controlled. The difference between the voltage reference (V_r) and the measured voltage amplitude (V_g) represents the amplitude tracking error, that is multiplied by the voltage droop coefficient (D_q). This value is then added to the reactive power error. The resulting signal is integrated with a $1/K$ gain, to generate the virtual flux linkage ($M_f.i_f$). Equation (3.18) allows the calculation of coefficient D_q :

$$D_q = -\frac{\Delta Q}{\Delta V} \quad (3.18)$$

where, ΔV is the voltage variation necessary to generate a given reactive power variation ΔQ .

Although the way of calculating the nominal value for the virtual inertia has been made explicit in this section, the variation of this parameter, as mentioned for the VSGs (Alipoor, Miura, and Ise 2013), may be beneficial to minimize oscillations under transient conditions. In (França, Castro, and Aredes 2015) a proposal of a virtual inertia adjustment is presented as a function of the rate of change of the active power. This relationship is such that, as the virtual inertia value is reduced, its active power injection is faster and the voltage regulation on the DC side becomes less robust.

The amplitude and phase detection of the grid can be performed by a three-phase phase-locked loop (PLL) in the original proposal. A pulse width modulation (PWM) technique is used to operate the static converter switches by connecting the synchronverter power part to its power control algorithm. Regarding the transient behavior, the active and reactive powers are coupled during the transient state, but this effect is not observed in steady state. Concerning voltage and frequency, the voltage magnitude is increased to provide the desired power, while the frequency suffers small variations in the transition intervals.

As can be seen, the synchronverter is similar to the classic droop control, having in common the control of power through adjustments in frequency and voltage. The main difference is easily verified by the presence of the virtual inertia of the synchronverter, which restricts frequency and voltage fluctuations when the converter is subjected to disturbances. The synchronverter can not only support local loads in island mode, but also inject power into the grid in the connected mode (Mo et al. 2015).

An enhancement of the synchronverter is proposed in (Zhong et al. 2014a). To avoid dynamic problems related to PLLs, the initial version of synchronverter is changed to not depend on units dedicated to synchronization. Other works also deal with stability analysis and propose small changes to improve the dynamic performance of the synchronverter (Wei, Jie, and Chunying 2015a, Konstantopoulos et al. 2015, Zhang et al. 2013a and Lu and Chu 2015).

Synchronverter applications in distributed generation are presented in the literature in wind systems (Ma 2012) and photovoltaic systems (Ming and Zhong 2014, Ferreira et al. 2016a). In addition, there are publications using synchronverters to improve the stability of HVDCs (Dong, Chi, and Li 2016) and STATCOMs (Nguyen et al. 2012).

Control of PWM rectifiers is also possible (Ma, Zhong, and Yan 2012a) through modifications in the original synchronverter, so that the converter operates as a virtual synchronous motor. For this, the reactive power, Q_{ref} , is controlled to track a reference through an integrator whose output is $M_f.i_f$, so that the power factor can be controlled. If Q_{ref} is brought to zero, the rectifier operates at unit power factor. A DC bus voltage control loop is added and also implemented through a PI controller whose output is the virtual torque T_m .

According to the author who coined the term synchronverter, what makes this model a unified interface for integration in Smart Grids is the possibility to function as a synchronous generator in applications of distributed generation and to operate as a synchronous motor through changes in its mathematical model, in applications involving loads connected to the grid by active rectifiers (Zhong 2016).

3.6 Comparative Analysis Between Droop Control, VSG and Synchronverter

Virtual synchronous machines and control schemes based on droop control have been studied independently, although they have a strong similarity. However, there are already works that propose to compare these techniques through mathematical formulations and simulations that try to reproduce the same operating conditions (Liu, Miura, and Ise 2016).

The droop control operation is also treated by some authors as an attempt to mimic the synchronous generators connected in conventional grids (Wandhare, Thale, and Agarwal 2014). In fact, the two approaches are equivalent under certain conditions (D'Arco and Suul 2014). This section performs a comparative analysis between the droop control, the VSG and the synchronverter, by manipulating the equations that govern each of the techniques. The reasoning is inspired by other works, but it is not identical to any of the approaches found in the current literature.

Initially, synchronverter equations are developed, in order to obtain the relations between angular frequency and active power. Figure 3.11 shows the interaction between the variables.

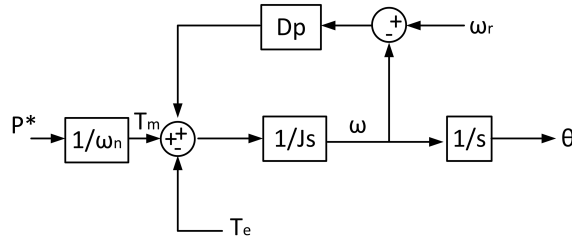


Figure 3.11: Active power control loop of the synchronverter.

From the block diagram of Figure 3.11, one can extract (3.22), according to the following development:

$$\omega = \frac{1}{Js} \cdot \left(\frac{P^*}{\omega_n} + (\omega_r - \omega) \cdot D_p - T_e \right) \quad (3.19)$$

$$\omega \cdot Js = \frac{P^*}{\omega_n} + (\omega_r - \omega) \cdot D_p - T_e \quad (3.20)$$

$$\omega \cdot (Js + D_p) = \frac{P^*}{\omega_n} + \omega_r \cdot D_p - T_e \quad (3.21)$$

$$\omega = \frac{\frac{P^*}{\omega_n} + \omega_r \cdot D_p - T_e}{(Js + D_p)} \quad (3.22)$$

As the goal is to compare droop control and synchronverter, and knowing that the presence of a virtual moment of inertia in the synchronverter is notorious, a simplification becomes necessary, making $J=0$, resulting in (3.23):

$$\omega = \frac{\frac{P^*}{\omega_n} - T_e}{D_p} + \omega_r \quad (3.23)$$

At this point, the droop control equation is derived, from the observation of Figure 3.12, leading to (3.24).

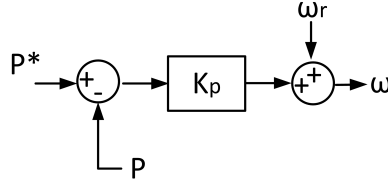


Figure 3.12: Droop control loop, active power part.

$$\omega = (P^* - P) \cdot k_p + \omega_r \quad (3.24)$$

Evaluating preliminarily, it is not possible to conclude whether (3.23) and (3.24) are equivalent, mainly because the equations for calculating D_p (of the synchronverter) and K_p (of the droop control) were not considered. This procedure is then performed, obtaining (3.26) as a relation between the two gains.

$$k_p = \frac{\Delta\omega}{P_n} \quad \text{and} \quad D_p = \frac{P_n}{\omega_n \Delta\omega} \quad (3.25)$$

$$D_p = \frac{1}{\omega_n \cdot k_p} \quad (3.26)$$

Taking (3.26) in the (3.23):

$$\omega = \frac{\frac{P^*}{\omega_n} - T_e}{D_p} + \omega_r \rightarrow \omega = \frac{\frac{P^*}{\omega_n} - T_e}{\frac{1}{\omega_n \cdot k_p}} + \omega_r \rightarrow \omega = \left(\frac{P^*}{\omega_n} - \frac{P}{\omega_n} \right) \cdot \omega_n \cdot k_p + \omega_r \quad (3.27)$$

$$\omega = (P^* - P) \cdot k_p + \omega_r \quad (3.28)$$

Since (3.28) is identical to (3.24), we can conclude that the synchronverter with $J=0$ behaves like a droop control (Liu, Miura, and Ise 2016) in terms of active power control.

For the VSG, a similar analysis is developed, beginning with (3.29), which represents the control of the virtual machine.

$$P_{in} - P_{out} = J \cdot \omega \frac{d\omega_m}{dt} + D_{(pu)} \cdot S_{base} \cdot \frac{\omega - \omega_r}{\omega_0} \quad (3.29)$$

where, P_{in} is the virtual power in the shaft, P_{out} is the measured power, S_{base} is the nominal power of DG, J is the virtual inertia, D is the damping coefficient, ω is the virtual angular frequency, ω_r is the angular frequency of the point where V is measured (or reference frequency in the islanded case) and ω_0 is the nominal angular frequency.

The equation of the governor is:

$$P_{in} = P^* - k_{p(pu)} \cdot S_{base} \cdot \frac{\omega - \omega_r}{\omega_0} \quad (3.30)$$

in which, P^* is the set-point of active power, $k_{p(pu)}$ is the droop coefficient, and the (pu) indicates value in p.u, thus:

$$k_p = \frac{k_{p(pu)} \cdot S_{base}}{\omega_0} \quad e \quad D = \frac{D_{(pu)} \cdot S_{base}}{\omega_0} \quad (3.31)$$

Taking (3.30) in (3.29), one obtains:

$$\omega = \frac{P^* + (k_p + D) \cdot \omega_r - P_{out}}{k_p + D + J\Delta\omega} \quad (3.32)$$

Note that (3.32) is therefore a general equation for the VSG, so that by making $D=0$ and $J=0$, we obtain (3.33), which again shows a behavior equivalent to that of the droop control.

$$\omega = \frac{(P^* - P)}{k_p} + \omega_r \quad (3.33)$$

The same methodology adopted for the active power is then applied to the reactive power. From the synchronverter, through the Figure 3.13, we get (3.34), which relates the magnitude of the voltage to the reactive power.

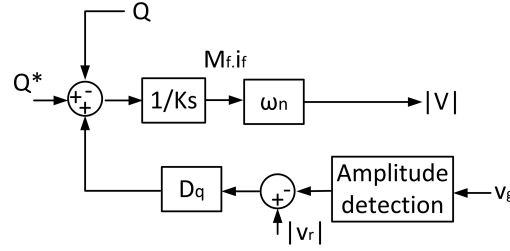


Figure 3.13: Reactive power control loop of the synchronverter.

$$V = \frac{Q^* - Q + D_q \cdot V_r}{\frac{k}{\omega_n} s + D_q} \quad (3.34)$$

Similarly to what has been pointed out in relation to the virtual moment of inertia, the friction coefficient of the synchronverter also does not exist in the droop control, thus, making $k=0$, (3.34) is transformed in:

$$V = \frac{Q^* - Q}{D_q} + V_r \quad (3.35)$$

From the droop control, in Figure 3.14, we obtain:

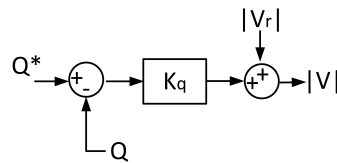


Figure 3.14: Droop control loop, reactive power part.

$$V = (Q^* - Q) \cdot k_q + V_r \quad (3.36)$$

The gains k_q of the droop control and D_q of the synchronverter are expressed, one as the inverse of the other, as (3.37).

$$k_q = \frac{\Delta V}{Q_n} \text{ and } D_q = \frac{Q_n}{\Delta V} \rightarrow D_q = \frac{1}{k_q} \quad (3.37)$$

Taking the (3.37) into (3.35), we have:

$$V = \frac{Q^* - Q}{D_q} + V_r \rightarrow V = (Q^* - Q) \cdot k_q + V_r \quad (3.38)$$

The equivalence between (3.36) and (3.38) shows that, with respect to reactive power control, the synchronverter behaves as the droop control, when $K=0$.

As for the VSG, its reactive power control consists of a proportional-integral controller that tries to zero out the steady state error of this variable, based on the adjustment of the terminal voltage of the converter. At this point, VSG and droop control do not differ, also because some references on the droop use PI controllers, so no mathematical deduction is necessary.

Table 3.1 shows a qualitative comparison of the techniques addressed in the present chapter, having as evaluated characteristics the presence or absence of the virtual inertia parameter, the implementation complexity and the control mode of the inverter (voltage or current controlled). For better understanding of the implementation complexity, Table 3.2 shows the number of math operations for each control method.

Table 3.1: Qualitative synthesis of active and reactive power control techniques.

Technique	Virtual inertia	Implementation	Controlled
Droop control	no*	very simple	voltage
VSG	yes	simple	voltage
Synchronverter	yes	simple	voltage
VISMA	yes	simple	current

Table 3.2: Number of math operations for each control method.

Control method	Implementation			
	Sum / Sub.	Multiplication	Integration	Trigonometric fcn.
Droop control*	5	5	1	3
Synchronverter	10	14	3	6
VSG	9	9	3	3
VISMA	6	10	3	3

*Considering the absence of low-pass filters in the calculation of active and reactive powers.

3.7 Simulation Results

In the previous section, a comparative analysis was presented, based on the mathematical manipulation of transfer functions and on qualitative characteristics of power control techniques. This section presents simulation results to allow the visualization of the dynamic behavior of the droop, VSG and synchronverter control algorithms.

A standard simulation scenario was established to show the three techniques at the same conditions, including grid-distribution characteristics, coupling impedance and power set-points. To avoid loss of generalization, the converters were replaced by ideal controlled voltage sources. Table 3.3 illustrates the parameters used in the simulation.

Table 3.3: Grid and converter parameters used in the simulation.

Grid parameters				
RMS voltage	Number of phases	Frequency	Lg	Rg
220V	3	60Hz	10.68 μ H	3m Ω
Converter parameters				
Active power	Reactive power	Number of phases	Ls	Rs
20kW	10kVAr	3	450 μ H	13.5m Ω

The control gains of each technique were obtained from (3.25) and (3.37), presented in the previous section. Variations in the active (P^*) and reactive (Q^*) power references are performed in the simulation as follows:

- t=0s: $P^*=0$ W and $Q^*=0$ VAr;
- t=1s: an active power step equal to $P^*=20$ kW is applied;
- t=1.5s: a reactive power step equal to $Q^*=10$ kW is applied;
- t=2s: P^* returns to 0W;
- t=2.5s: Q^* returns to 0VAr;

Figures 3.15 and 3.16 illustrate the dynamic behavior of the droop, VSG and synchronverter control algorithms, in their active and reactive control loops respectively. To make sense of the comparison, virtual inertia and virtual friction coefficient of the VSG and synchronverter were zeroed.

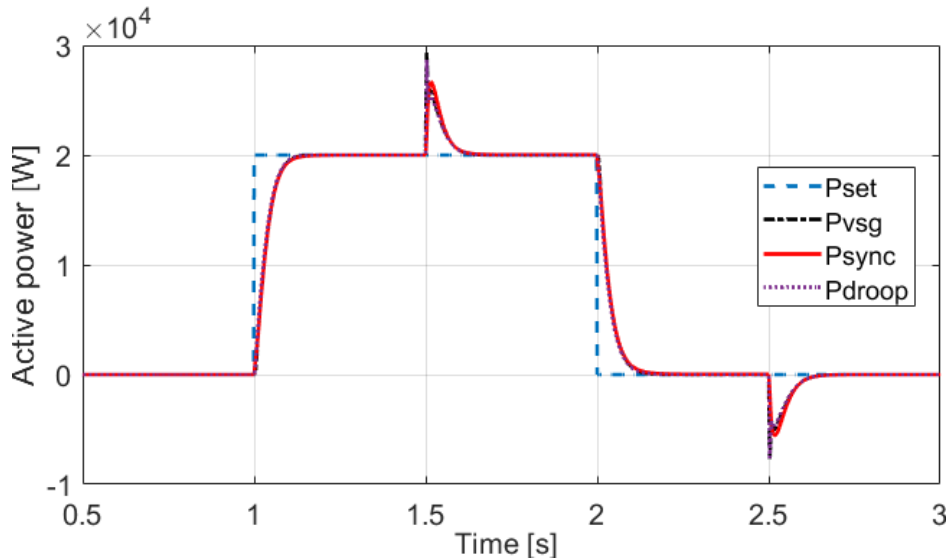


Figure 3.15: Dynamic behavior of the active power loops of the droop control, VSG and synchronverter.

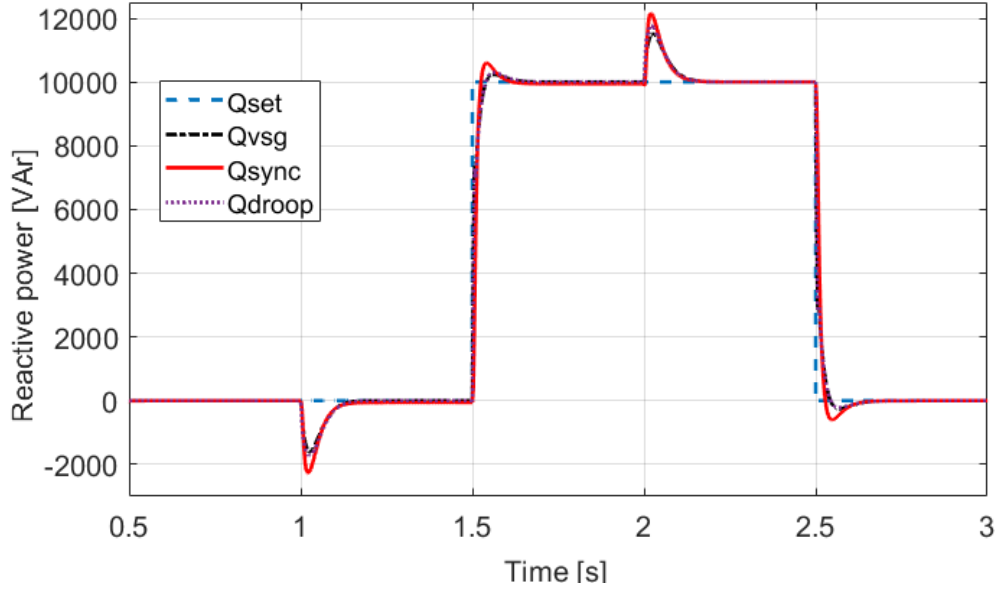


Figure 3.16: Dynamic behavior of the reactive power loops of the droop control, VSG and synchronverter.

In the Figure 3.15 it is possible to verify a practically identical dynamic behavior between droop control and VSG, whereas the synchronverter behaves slightly different during transients. In principle, it is believed that this difference is attributed to the coupling between the P and Q control loops in the synchronverter by means of the portion corresponding to the flux linkage ($M_f \cdot I_f$), present in the electromagnetic torque calculation (T_e).

In Figure 3.16, again the droop control and the VSG show practically identical behavior, while the synchronverter dynamics are notably different.

3.8 Conclusions

In this chapter, the characteristics of control algorithms emulating the synchronous machine were presented. The simplicity of the implementation of droop control and the similarity between synchronverter and VSG, which conceptually seek to emulate the synchronous machine, are observations that can be highlighted. These latter methods allows the selection of parameters such as the virtual moment of inertia and coefficient of friction, which contribute to the stability of the power system.

The VISMA also presents such virtual parameters, but has as main difference the fact of being controlled in current, while in the other techniques studied the converter is voltage controlled.

In the comparative analysis of power control techniques, the conclusion that droop control can be considered a particular case of the techniques that emulate the synchronous machine, demystifies that they are absolutely different strategies and allows an appropriate choice and implementation. In the simulation results, we saw a difference in the synchronverter dynamics, which deserves to be better investigated.

4 Dynamic Analysis of Grid-connected Droop-controlled Converters, Synchronverters and VSGs.

This chapter presents a dynamic analysis for grid-connected droop-controlled converters, synchronverters and virtual synchronous generators. First, a detailed modeling of the three methods is described in the synchronous reference frame. Then, a unified small-signal modeling is developed to allow representing not only the mentioned control algorithms, but also to model other control methods in which the connected converter is controlled as a voltage source. The presentation of droop, synchronverter and VSG parameters relations allows the evaluation of these methods with equivalent control gains. The small-signal models are validated through simulation and experimental results and are used to access information on the dynamic operation of the systems.

4.1 Introduction

Among the main reasons for the widespread use of control techniques that emulate the operation of synchronous machines are:

- Well proven capability of parallel connected synchronous machines;
- Possibility to independently control the injection of reactive and active power;
- Possibility to operate parallel machines with different rated parameters;
- Easy power sharing capability.

Although these main characteristics are known, the deep understanding of their dynamic behavior demands a detailed modelling and stability analysis. A unified small-signal model can be developed from the schematic diagram represented in Figure 4.1. The simplified schematic representation illustrates a micro-source connected to a power bus through a RL series impedance. For the sake of simplicity, the bus is assumed to have a rated power much greater than the micro-source, such that the operation of this last is not capable of changing the grid frequency (ω_0) or grid voltage magnitude (v_0) of the bus. In addition, the micro-source is modeled initially as an ideal voltage-controlled source. Then the power converter with its high-frequency dynamics due to the PWM (pulse-width modulation) switching, internal filters and voltage and current control loops are included to take into account any influence of these elements on the active and reactive power flow control.

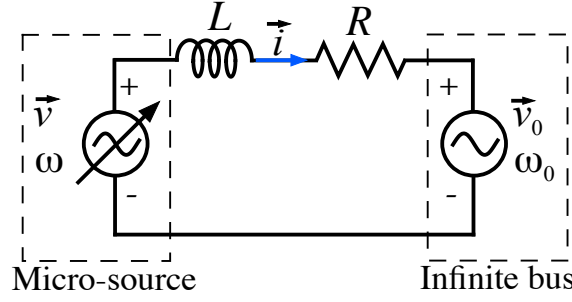


Figure 4.1: Schematic diagram of the micro-source connected to an infinite bus.

A small-signal model for the system with each control architecture is developed and used to allow a better understanding of the dynamic behavior of the alternatives under parameter variations and disturbances. The small-signal models are validated through simulations using Matlab/Simulink and Typhoon HIL, with the micro-source control running in a Texas Instruments TMS320F28335 micro-controller. Once the small-signal models are validated, they are used to access information on the dynamic operation of the systems under more general operating conditions. The unified approach used to develop the small-signal models can also be used to model other algorithms that mimic the synchronous machine operation. It is also shown that the models developed could be used even for large signal excursions with acceptable errors.

4.2 Unified Small-signal Model

In this section, small-signal models for the droop-controlled converter, the synchronverter and the VSG connected to an infinite bus are developed. A similar approach based on a synchronously rotating reference frame is used for both systems, aiming an easier comparison of the small-signal models.

4.2.1 Droop-controlled Converter

The micro-source considered in this study is assumed to operate with a frequency ω and a voltage magnitude v , which are based on the droop control strategy (Brabandere et al. 2005; Chandorkar, Divan, and Adapa 1993; Coelho, Cortizo, and Garcia 2002; Guerrero et al. 2007; Kawabata and Higashino 1988; Lee et al. 2009; Liu et al. 2016; Majumder et al. 2010; Mendoza-araya 2014; Moslemi and Mohammadpour 2015; Tayab et al. 2017; Wen et al. 2016).

As illustrated in Figure 4.2, the droop-controlled micro-source varies its frequency and voltage magnitude based on the active (p) and reactive (q) power at its terminals, with a voltage droop gain (K_q) and a frequency droop gain (K_p). The idea behind the droop control is the emulation of a synchronous machine connected to a power bus through a purely inductive line. In the present work, the R/L relation is considered very small and the coupling and the transformations of the measured powers are not considered.

In Figure 4.2, P^* is the active power that the micro-source has to inject when operating at frequency Ω^* , and Q^* is the reactive power to be injected when operating with a voltage magnitude equal to V^* . The Sinusoidal Waveform Generation block is responsible for generating the three-phase voltages, based on the voltage vector angle (θ) and magnitude (v).

Using a vector representation for the three-phase variables (Novotny and Lipo 1996), the vector quantities of Figure 4.2 can be written as:

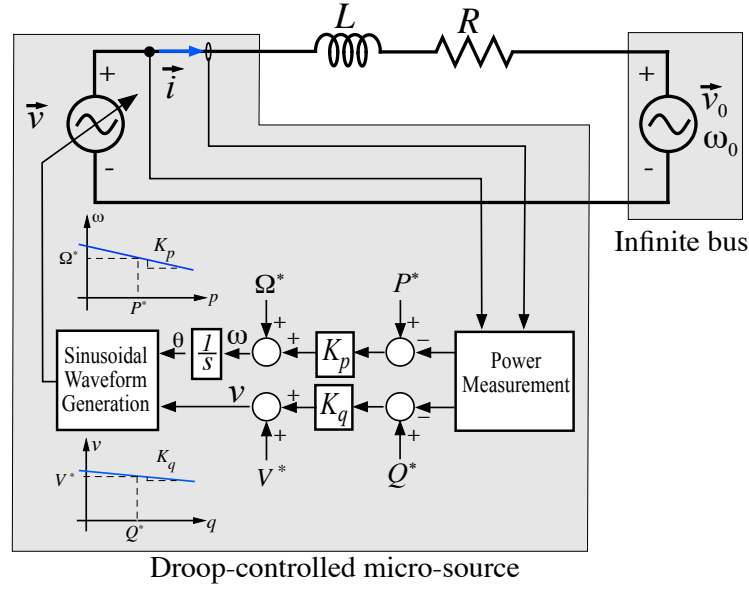


Figure 4.2: Schematic diagram of the droop-controlled micro-source.

$$\vec{v} = v_d + jv_q \quad (4.1)$$

$$\vec{v}_0 = v_{0d} + jv_{0q} \quad (4.2)$$

$$\vec{i} = i_d + ji_q \quad (4.3)$$

where subscripts d and q stand for the direct and quadrature axis components of the vectors quantities in a synchronous rotating frame.

The circuit equation for the system in the vector form is:

$$\vec{v} - \vec{v}_0 = [(s + j\omega)L + R]\vec{i} \quad (4.4)$$

In the vector form, the term $j\omega$ has to be added to the derivative Laplace symbol s to represent the cross-couplings between the d and q axes variables.

The magnitude and angle of the micro-source voltage vector are, respectively:

$$v = |\vec{v}| = \sqrt{v_d^2 + v_q^2} \quad (4.5)$$

$$\theta = \arctan\left(\frac{v_q}{v_d}\right) \quad (4.6)$$

In the time domain, (4.4) can be rewritten for the d -axis and q -axis components as:

$$\frac{di_d}{dt} = \frac{1}{L}(v_d - v_{0d} - Ri_d + \omega Li_q) \quad (4.7)$$

$$\frac{di_q}{dt} = \frac{1}{L}(v_q - v_{0q} - Ri_q - \omega Li_d) \quad (4.8)$$

The active and reactive power at the micro-source terminals can be expressed as:

$$p = \frac{3}{2}(v_d i_d + v_q i_q) \quad (4.9)$$

$$q = \frac{3}{2}(v_q i_d - v_d i_q) \quad (4.10)$$

A first-order low-pass filter with a cut-off frequency equal to ω_f is commonly included in the power measurement of the droop-controlled micro-source, as expressed by:

$$p_f(s) = \frac{\omega_f}{s + \omega_f} p(s) \quad (4.11)$$

$$q_f(s) = \frac{\omega_f}{s + \omega_f} q(s) \quad (4.12)$$

From Figure 4.2, the frequency and voltage droop equations can be written as:

$$\omega = \Omega^* - K_p(p_f - P^*) \quad (4.13)$$

$$v = V^* - K_q(q_f - Q^*) \quad (4.14)$$

Building upon the traditional droop control presented here, many different variations have been proposed to improve the operation of the droop-controlled converter, including Moslemi and Mohammadpour 2015; Rowe and Summers 2013; Skjellnes, Skjellnes, and Norum 2002; Zhang et al. 2014. While the methodology developed here can also be used to deal with different control strategies with minor modifications, these techniques are considered to be out of the scope of this work, in order to maintain the tight focus on the more general model.

In (Coelho, Cortizo, and Garcia 2002), a small-signal model of a droop-controlled source was developed. However, it was assumed that the frequency variations were negligible and the effect of these variations on the line impedances was not considered. It is important to emphasize that although frequency variation in power systems are usually small, the effect of these variations must be taken into account for a more complete understanding of the dynamics of the system, as it will be shown bellow.

If the frequency variations in (4.7) and (4.8) are accounted, these equations become no-linear due to the current-frequency product.

Linearizing (4.7) and (4.8), one can obtain:

$$\frac{d\tilde{i}_d}{dt} = \frac{1}{L}(\tilde{v}_d - \tilde{v}_{0d}) - \frac{R}{L}\tilde{i}_d + \Omega\tilde{i}_q + I_q\tilde{\omega} \quad (4.15)$$

$$\frac{d\tilde{i}_q}{dt} = \frac{1}{L}(\tilde{v}_q - \tilde{v}_{0q}) - \frac{R}{L}\tilde{i}_q - \Omega\tilde{i}_d - I_d\tilde{\omega} \quad (4.16)$$

where the accent $\sim$ is used to denote variables with a small variation around an operating point. In addition, throughout this work, upper-case letters are used for constant and stand-still values of the variables in the operating point.

From (4.15) and (4.16), it is possible to draw the linearized DQ equivalent circuit of the micro-source connected to an infinite bus, as shown in Figure 4.3. Observe the inclusion of a term that accounts for the effect of the frequency variations on the currents of the system. This term is essential for a more complete modeling of the micro-source.

Linearizing (4.5), one can obtain:

$$\tilde{v} = n_d\tilde{v}_d + n_q\tilde{v}_q \quad (4.17)$$

where:

$$n_d = \frac{V_d}{\sqrt{V_d^2 + V_q^2}} = \frac{V_d}{V} \quad (4.18)$$

$$n_q = \frac{V_q}{\sqrt{V_d^2 + V_q^2}} = \frac{V_q}{V} \quad (4.19)$$

The linearization of (4.6) leads to:

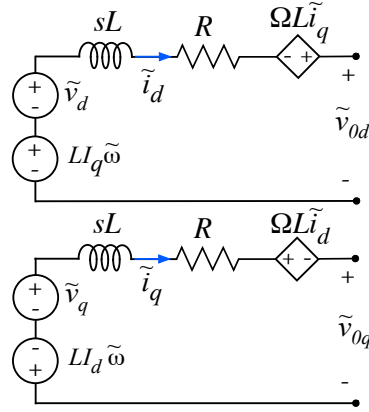


Figure 4.3: Linearized DQ equivalent circuit of the micro-source connected to an infinite bus.

$$\tilde{\theta} = m_d \tilde{v}_d + m_q \tilde{v}_q \quad (4.20)$$

where:

$$m_d = \frac{-V_q}{V_d^2 + V_q^2} = \frac{-V_q}{V^2} \quad (4.21)$$

$$m_q = \frac{V_d}{V_d^2 + V_q^2} = \frac{V_d}{V^2} \quad (4.22)$$

Differentiating (4.17) and (4.20), one can obtain:

$$\dot{\tilde{v}} = n_d \dot{\tilde{v}}_d + n_q \dot{\tilde{v}}_q \quad (4.23)$$

$$\dot{\tilde{\theta}} = \dot{\tilde{\omega}} = m_d \dot{\tilde{v}}_d + m_q \dot{\tilde{v}}_q \quad (4.24)$$

For small-signal quantities, (4.13) and (4.14) become:

$$\tilde{\omega} = \omega - \Omega^* = -K_p(p_f - P^*) = -K_p \tilde{p}_f \quad (4.25)$$

$$\tilde{v} = v - V^* = -K_q(q_f - Q^*) = -K_q \tilde{q}_f \quad (4.26)$$

The use of (4.11) and (4.12) in (4.25) and (4.26) yields:

$$\frac{d\tilde{\omega}}{dt} = -\omega_f \tilde{\omega} - K_p \omega_f \tilde{p} \quad (4.27)$$

$$\frac{d\tilde{v}}{dt} = -\omega_f \tilde{v} - K_q \omega_f \tilde{q} \quad (4.28)$$

Using (4.17) - (4.28) and after manipulations, it is possible to obtain:

$$\dot{\tilde{v}}_d = -V_q \tilde{\omega} - \frac{\omega_f V_d^2}{V^2} \tilde{v}_d - \frac{\omega_f V_d V_q}{V^2} \tilde{v}_q - \frac{\omega_f K_q V_d}{V} \tilde{q} \quad (4.29)$$

$$\dot{\tilde{v}}_q = V_d \tilde{\omega} - \frac{\omega_f V_d V_q}{V^2} \tilde{v}_d - \frac{\omega_f V_q^2}{V^2} \tilde{v}_q - \frac{\omega_f K_q V_q}{V} \tilde{q} \quad (4.30)$$

Linearization of (4.9) and (4.10) leads to:

$$\tilde{p} = \frac{3}{2} (V_d \tilde{i}_d + I_d \tilde{v}_d + V_q \tilde{i}_q + I_q \tilde{v}_q) \quad (4.31)$$

$$\tilde{q} = \frac{3}{2} (V_q \tilde{i}_d + I_d \tilde{v}_q - V_d \tilde{i}_q - I_q \tilde{v}_d) \quad (4.32)$$

In a matrix form, (4.15) and (4.16) can be re-written as:

$$\begin{bmatrix} \tilde{v}_d \\ \tilde{v}_q \end{bmatrix} - \begin{bmatrix} \tilde{v}_{0d} \\ \tilde{v}_{0q} \end{bmatrix} = \begin{bmatrix} (sL + R) & -\Omega L \\ \Omega L & (sL + R) \end{bmatrix} \begin{bmatrix} \tilde{i}_d \\ \tilde{i}_q \end{bmatrix} + \begin{bmatrix} -LI_q \\ LI_d \end{bmatrix} [\tilde{\omega}] \quad (4.33)$$

In (4.33) the large-signal impedance matrix of the system is:

$$\mathbf{Z}_0 = \begin{bmatrix} (sL + R) & -\Omega L \\ \Omega L & (sL + R) \end{bmatrix} \quad (4.34)$$

where boldface characters represent matrix quantities.

From (4.33), if the micro-source voltage and the bus voltage variations are set to zero, one can obtain:

$$\mathbf{Z}_0 \begin{bmatrix} \tilde{i}_d \\ \tilde{i}_q \end{bmatrix} = \begin{bmatrix} LI_q \\ -LI_d \end{bmatrix} [\tilde{\omega}] \quad (4.35)$$

From (4.35), solving for $\tilde{i}_d$ and $\tilde{i}_q$ yields:

$$\tilde{i}_d(s) = -\frac{1}{\Omega L} [(sL + R)\tilde{i}_q + LI_d\tilde{\omega}] \quad (4.36)$$

$$\tilde{i}_q(s) = \frac{1}{\Omega L} [(sL + R)\tilde{i}_d - LI_q\tilde{\omega}] \quad (4.37)$$

Using (4.36) and (4.37) in (4.35), it is possible to obtain a relation that express the impact of the frequency variation on the circuit currents, as:

$$\frac{\tilde{i}_d(s)}{\tilde{\omega}(s)} = \frac{-(s + R/L)I_q - \Omega I_d}{[(s + R/L)^2 + \Omega^2]} \quad (4.38)$$

$$\frac{\tilde{i}_q(s)}{\tilde{\omega}(s)} = \frac{(s + R/L)I_d - \Omega I_q}{[(s + R/L)^2 + \Omega^2]} \quad (4.39)$$

In a matrix form, (4.38) and (4.39) are expressed as:

$$\begin{bmatrix} \tilde{i}_d(s) \\ \tilde{i}_q(s) \end{bmatrix} = \begin{bmatrix} \frac{-(s+R/L)I_q - \Omega I_d}{[(s+R/L)^2 + \Omega^2]} \\ \frac{(s+R/L)I_d - \Omega I_q}{[(s+R/L)^2 + \Omega^2]} \end{bmatrix} [\tilde{\omega}] = \mathbf{G}_{i\omega} [\tilde{\omega}] \quad (4.40)$$

From (4.31) and (4.32), a matrix representation for the active and reactive powers can be written as:

$$\tilde{S} = \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} = \frac{3}{2} \begin{bmatrix} V_d & V_q \\ V_q & -V_d \end{bmatrix} \begin{bmatrix} \tilde{i}_d \\ \tilde{i}_q \end{bmatrix} + \frac{3}{2} \begin{bmatrix} I_d & I_q \\ -I_q & I_d \end{bmatrix} \begin{bmatrix} \tilde{v}_d \\ \tilde{v}_q \end{bmatrix} \quad (4.41)$$

For the sake of simplification, we define matrices:

$$\mathbf{V}_{dq} = \frac{3}{2} \begin{bmatrix} V_d & V_q \\ V_q & -V_d \end{bmatrix} \quad (4.42)$$

$$\mathbf{I}_{dq} = \frac{3}{2} \begin{bmatrix} I_d & I_q \\ -I_q & I_d \end{bmatrix} \quad (4.43)$$

From (4.17) and (4.20) and after manipulations, one can write:

$$\tilde{v}_d = \frac{V_d}{V} \tilde{v} - V_q \tilde{\theta} \quad (4.44)$$

$$\tilde{v}_q = -\frac{V_q}{V} \tilde{v} + V_d \tilde{\theta} \quad (4.45)$$

Recalling that $\tilde{\omega} = s\tilde{\theta}$, (4.44) and (4.45) can be written in a matrix form as:

$$\begin{bmatrix} \tilde{v}_d \\ \tilde{v}_q \end{bmatrix} = \begin{bmatrix} -\frac{V_q}{s} & \frac{V_d}{V} \\ \frac{V_d}{s} & -\frac{V_q}{V} \end{bmatrix} \begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} = \mathbf{G}_{rp} \begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} \quad (4.46)$$

The frequency variations can be also represented in a matrix form, based on the definition developed above, as:

$$\begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} = \mathbf{G}_\omega \begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} \quad (4.47)$$

Writing the droop gains and the low-pass filter in a matrix form leads to:

$$\begin{bmatrix} \tilde{\omega} \\ \tilde{\theta} \end{bmatrix} = \begin{bmatrix} -K_p \frac{\omega_f}{(s+\omega_f)} & 0 \\ 0 & -K_q \frac{\omega_f}{(s+\omega_f)} \end{bmatrix} \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} = \mathbf{G}_d \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} \quad (4.48)$$

Once the small-signal model for the droop-controlled micro-source is now available, it is possible to include the dynamics of the control loop of the LC filter of the converter. Figure 4.4 presents the block diagram of the voltage and current control loops of the converters. In the voltage loop a proportional+integral (PI) controller is used. For the current control loop a proportional (P) controller is applied. In this Figure, G_v is the voltage sensor gain, G_i is the current sensor gain, K_{pv} and K_{iv} are the proportional and integral gains of the voltage loop, K_{pi} is the proportional gain in current loop, G_{inv} is the converter gain, L_f , R_f and C_f are the inductance, inductor resistance, and capacitance of the LC output filter. Figure 4.4 also presents the output currents of the micro-source ($i_{o,dq}$), which act as disturbance for the control loops, and the voltage references v_{dq}^* from the droop, synchronverter, or VSG control algorithms, which act as set points for the voltage control loop.

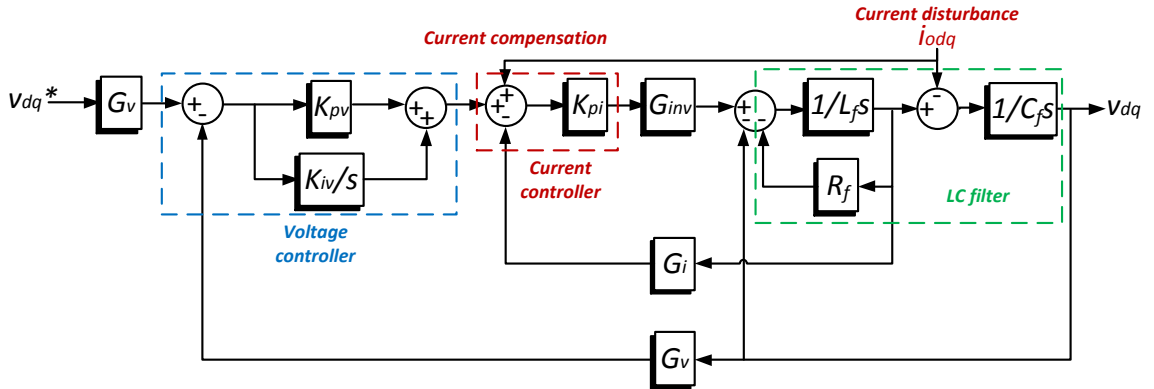


Figure 4.4: Block diagram of inverter inner control, including the LC filter behavior.

From the block diagram of Figure 4.4, it is possible to obtain the two transfer-functions relating the filter output voltages with the set point and disturbance of the control system, as presented in (4.49) and (4.50). Both contributions are added to obtain v_{dq} as illustrated in Figure 4.5. Eq. (4.51) presents these relations in a matrix form.

It is important highlight that G_c is a ratio between the output voltage and the voltage reference to the inverter. So, this equation results in a dq output voltage to be converted to $\tilde{\omega}, \tilde{\theta}$, using matrix (4.52).

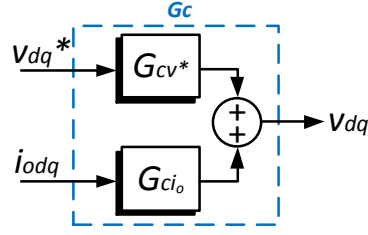


Figure 4.5: DQ small-signal representation of the control of the output filter.

$$\tilde{v}_{dq_{v*}} = \frac{\frac{G_v \cdot K_{pv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f} \cdot s + \frac{G_v \cdot K_{iv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f}}{s^3 + \left(\frac{R_f}{L_f} + \frac{K_{pi} \cdot G_{inv} \cdot G_i}{L_f}\right) \cdot s^2 + \frac{G_v \cdot K_{pv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f} + \frac{G_v \cdot K_{iv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f}} \cdot \tilde{v}_{dq}^* = \mathbf{G}_{cv*} \cdot \tilde{v}_{dq}^* \quad (4.49)$$

$$\tilde{v}_{dq_{io}} = \frac{-L_f s^2 - R_f s}{s^3 + \left(\frac{R_f}{L_f} + \frac{K_{pi} \cdot G_{inv} \cdot G_i}{L_f}\right) \cdot s^2 + \frac{G_v \cdot K_{pv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f} + \frac{G_v \cdot K_{iv} \cdot K_{pi} \cdot G_{inv}}{C_f L_f}} \cdot \tilde{i}_{o_{dq}} = \mathbf{G}_{ci_o} \cdot \tilde{i}_{o_{dq}} \quad (4.50)$$

$$\tilde{v}_{dq} = \begin{bmatrix} G_{cv*} & G_{ci_o} \end{bmatrix} \begin{bmatrix} \tilde{v}_{dq}^* \\ \tilde{i}_{o_{dq}} \end{bmatrix} \quad (4.51)$$

$$\mathbf{G}_{pr} = \frac{1}{V_d^2 - V_q^2} \begin{bmatrix} V_q \cdot s & V_d \cdot s \\ V_d \cdot V & V_q \cdot V \end{bmatrix} \quad (4.52)$$

Using the matrix representation of the developed small-signal models, it is possible to obtain a block diagram representation for the complete system, as shown in Figure 4.6.

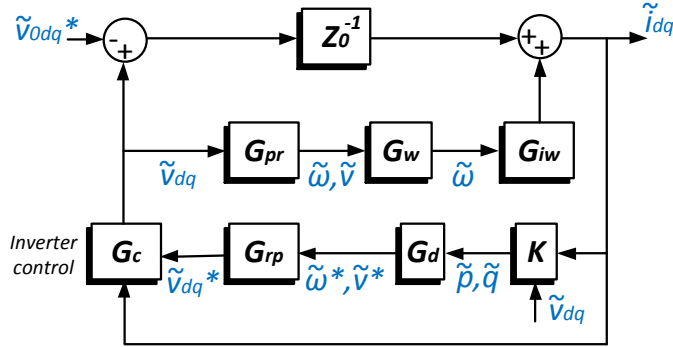


Figure 4.6: DQ small-signal block diagram of the micro-source.

From Figure 4.6, the transfer function relating small variations in the bus terminal voltage with the variations in the micro-source currents can be obtained. This relation is the small-signal output impedance of the droop-controlled micro-source, given by:

$$\frac{\tilde{v}_{o_{dq}}(s)}{\tilde{i}_{dq}(s)} = G_{vi}(I + Z_0 G_{pr} G_{\omega} G_{i\omega}) - Z_0 \quad (4.53)$$

where:

$$\mathbf{G}_{vi} = \frac{G_{cv*}G_{rp}G_dV_{dq} + G_{ci_o}}{I - G_{cv*}G_{rp}G_dI_{dq}} \quad (4.54)$$

$$\mathbf{Z}_{ss}(s) = \frac{\tilde{v}_{dq}(s)}{\tilde{i}_{dq}(s)} = \begin{bmatrix} Z_{dd}(s) & Z_{qd}(s) \\ Z_{dq}(s) & Z_{qq}(s) \end{bmatrix} \quad (4.55)$$

Equation (4.55) gives the small-signal impedances from channels d and q to channels d and q .

4.2.2 Synchronverter

The concept of the synchronverter was firstly introduced by Zhong, Member, and Weiss 2011 in an attempt to mimic the dynamic behavior of synchronous machines connected to the power grid, through the use of power electronic converters. Since then, many efforts have been developed to better understand the synchronverter behavior and to improve its dynamic response (Ma, Zhong, and Yan 2012b; Zhang et al. 2013b; Zhong et al. 2014b), in addition to adaptations for single-phase applications (Ferreira et al. 2016b; Mishra et al. 2016a). Figure 4.7 shows a schematic diagram of the synchronverter. While it appears to be different from the schematic diagram proposed in (Zhong, Member, and Weiss 2011), the model presented herein uses the variables represented in the DQ reference frame, for the sake of simplicity and model unification. This approach enables a straight forward comparison with the droop control scheme, presented in Figure 4.2.

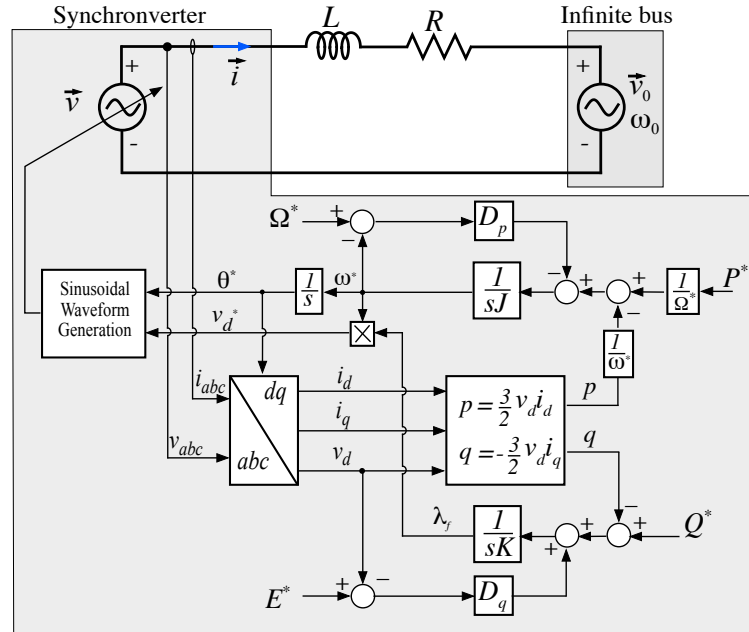


Figure 4.7: Schematic diagram of the synchronverter represented in DQ reference frame.

In Figure 4.7, J is the emulated moment of inertia, D_p is the emulated damping coefficient, D_q is the voltage droop coefficient and K is a constant used to calculate the flux linkage (λ_f) of the virtual machine. Also in the Figure, one can note the presence of a block ($abc - dq$) responsible for transforming the three-phase voltages and currents to a synchronous rotating reference frame, aligned with the d -axis voltage component of the generator voltage. By doing this, the q -axis voltage (v_q) component will be zero and the active and reactive power will be functions of the direct (i_d) and quadrature (i_q) current components, respectively, besides the d -axis voltage (v_d) component. The function of the Sinusoidal Waveform Generator block is the same as for the droop control algorithm, shown in Figure 4.2.

In (Wei, Jie, and Chunying 2015b), a small-signal model of a synchronverter was developed. However, in the adopted approach the droop gain to regulate the voltage magnitude was not taken into account. A modeling based on an *abc* reference also did not contribute very much to the understanding of the similarities and differences between the droop-control and the synchronverter.

Most of the previous development presented here for the small-signal model of the droop-controlled converter can also be applied to obtain the small-signal model of the synchronverter. Specifically, if one looks at Figure 4.6, it is possible to note that, excluding the Control Matrix (G_d) of the droop-control, all the other elements can also be used to model the synchronverter connected to the infinite bus, since they deal with the model of the system without its control. The analysis of Figure 4.7 allows us to obtain a relation between the small-signal active and reactive powers and the small-signal frequency and voltage variations for the synchronverter as:

$$\begin{bmatrix} \tilde{\omega} \\ \tilde{v} \end{bmatrix} = \begin{bmatrix} -\frac{1}{\Omega} \frac{1}{(sJ+D_p)} & 0 \\ \frac{\Lambda_f}{\Omega} \frac{1}{(sJ+D_p)} & -\frac{1}{(s\frac{K}{\Omega}+D_q)} \end{bmatrix} \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} = \mathbf{G}_s \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} \quad (4.56)$$

By the use of matrix $\mathbf{G}_s$ as the Control Matrix in Figure 4.6, one can obtain a small-signal model for the synchronverter. It is worth noting that (4.53)-(4.55) can also be used to calculate the small-signal impedances and/or admittance of the synchronverter, if matrix $\mathbf{G}_d$ is replaced by $\mathbf{G}_s$.

4.2.3 Virtual Synchronous Generator

Using the same methodology presented for droop control and synchronverter, it is possible to draw the virtual synchronous generator in DQ reference frame, as can be seen in Figure 4.8. The red-highlighted parts clarify the difference in VSG implementation from synchronverter. Basically, in the active power loop, the G_p gain presented in the governor equation modify the algorithm. On the other hand, the rective power control is implemented in the VSG by the use of a proportional-integral controller, unlike synchronverter.

One more time, the small-signal presented in this chapter can be used, through the substitution of G_d from droop control for the G_v shown in (4.57). It is important to observe that in both the figure and matrix we do not have the coupling between active and reactive power loop, different of the synchronverter algorithm.

$$\begin{bmatrix} \tilde{\omega} \\ \tilde{v} \end{bmatrix} = \begin{bmatrix} -\frac{G_p+D}{(G_p+D)Js+\Omega} & 0 \\ 0 & -(\frac{G_{qi}}{s} + G_{qp}) \end{bmatrix} \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} = \mathbf{G}_v \begin{bmatrix} \tilde{p} \\ \tilde{q} \end{bmatrix} \quad (4.57)$$

where G_p is the proportional gain from the governor, D is the damping factor, G_{qi} and G_{qp} are the integral and proportional gains of the reactive power loop.

4.3 Synchronverter, Droop and VSG Parameter Relations

Comparing the non-zero elements of matrices $\mathbf{G}_d$, $\mathbf{G}_s$ and $\mathbf{G}_v$ it is possible to determine the parameter values for the droop-control and the synchronverter that would give them a closer dynamic response. For the frequency droop gain, one can obtain:

$$K_p = \frac{1}{D_p \Omega} \rightarrow D_p = \frac{1}{K_p \Omega} \quad (4.58)$$

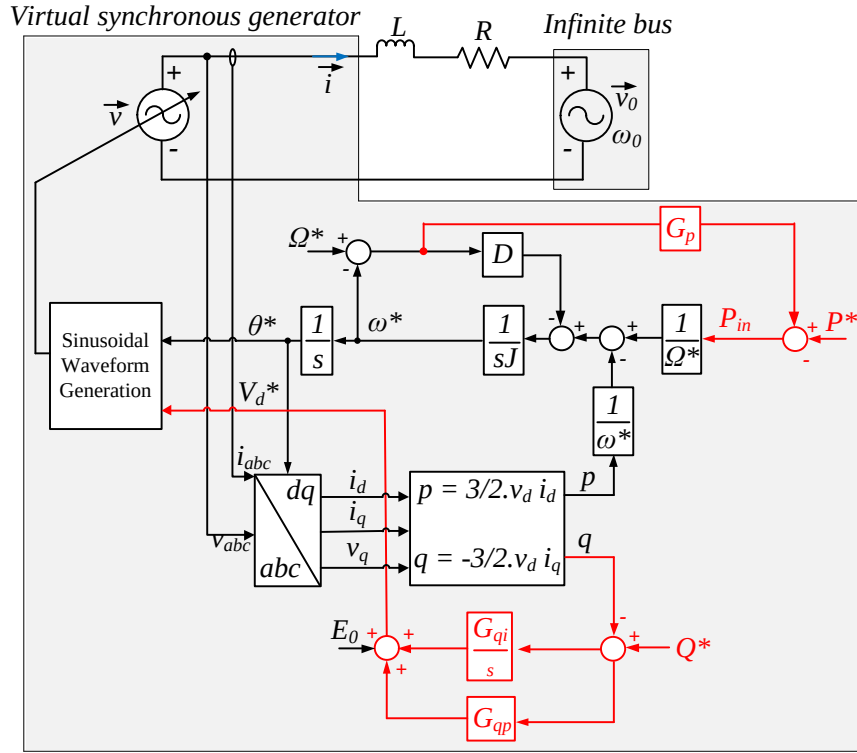


Figure 4.8: Schematic diagram of the virtual synchronous generator micro-source.

For the cut-off frequency of the low-pass filter we get:

$$\frac{1}{\omega_f} = \frac{J}{D_p} \rightarrow J = \frac{D_p}{\omega_f} \quad (4.59)$$

$$\frac{1}{\omega_f} = \frac{K}{\Omega D_q} \rightarrow K = \frac{\Omega D_q}{\omega_f} \quad (4.60)$$

For the magnitude droop gain, it is obtained:

$$K_q = \frac{1}{D_q} \quad (4.61)$$

According to Zhong, Member, and Weiss 2011, J and K are calculated through:

$$J = \tau_f D_p \quad (4.62)$$

$$K = \tau_v \omega_n D_q \quad (4.63)$$

where τ_f and τ_v are the frequency and voltage droop time constants, respectively. From the previous equations, it is possible to express:

$$\tau_f = \tau_v = \frac{1}{\omega_f} \quad (4.64)$$

For the sake of simplicity, in this work these time constants are considered equal and the filters in the measurement of the active and reactive powers of the droop control algorithm have the same time constants. It is worth noting that the low-pass filter commonly used in the power measurement of droop-controlled converters adds the effect of a virtual inertia to this control strategy, which could also be varied for increased flexibility.

Regarding the VSG, the comparison of $\mathbf{G}_v$ with $\mathbf{G}_d$ and $\mathbf{G}_s$ allows to conclude that the VSG has a more explicit equivalence with the synchronverter in the active power control loop and it has a closest implementation of the droop control in the reactive power control loop. The parameter relations are shown in Table 4.1. In addition, the coupling between active and reactive power control loops observed in the synchronverter is not used in the VSG.

Table 4.1: Droop, synchronverter and VSG parameter relations.

Control Loop	Droop	Synchronverter		VSG
Frequency	K_p	$D_p = \frac{1}{\Omega K_p}$	$J = \tau_f D_p$	$G_p + D = \frac{1}{K_p}$
	ω_f	$\tau_f = \frac{1}{\omega_f}$		$J_{VSG} = J_{synch}$
Voltage	K_q	$D_q = \frac{1}{K_q}$	$K = \tau_v \Omega D_q$	* $G_{qp} = K_q$
	ω_f	$\tau_v = \frac{1}{\omega_f}$		

* In connected mode there is a need to use the integral term, omitted here to make the comparison feasible.

Another interesting remark when comparing matrices $\mathbf{G}_d$ and $\mathbf{G}_s$ is that the droop-controlled micro-source can be seen as a particular case of the synchronverter or VSG when element G_{s21} is equal to zero and the equivalence of parameters of Table 4.1 is considered.

4.4 Validation of the Small-signal Models

To validate the small-signal models of droop-controlled micro-source, synchronverter and the VSG, dynamic models were implemented in Matlab/Simulink. The parameters used in the simulations are presented in table 4.2. A set of simulations was conducted using the small-signal and the dynamic models with these parameters.

Figures 4.9-4.12 show the transient response of the droop-controlled micro-source, the synchronverter and the VSG from the dynamic and small-signal models, after a variation of the infinite bus voltage. The d -axis component of the bus voltage was increased by 5% for 50ms. The frequency droop gain (K_p) was set to 0.05%, the voltage droop gain (K_q) was set to 0.01% and the cut-off frequency of the low-pass filter (ω_f) was set to 6Hz. The parameters of the synchronverter and the VSG were calculated from these values using the expressions listed in Table 4.1. It is important to note that the behavior of the low-pass filter of the droop control is present in the VSG through the virtual inertia J , but the same is not observed in the reactive power loop.

Figure 4.9 shows the results for the droop controlled-converter, Figure 4.10 presents the results for the synchronverter and Figure 4.11 shows the results for the VSG. In Figure 4.12 it is possible to compare the results for the three control strategies. For all the cases, Figure (a) shows the variation of the micro-source frequency ($\tilde{\omega}$), and Figure (b) the variation of the micro-source d -axis current ($\tilde{i}_d$).

As shown in Figures 4.9, 4.10 and 4.11, there is a good agreement in the results of the dynamic models and the small-signal models for the three control strategies. Many other simulations with different

Table 4.2: Base parameters for simulations of synchronverter and droop controlled converter.

Electrical Parameters	Value	Units
System frequency (f_0)	60	Hz
Infinite bus voltage magnitude (V_0)	$127\sqrt{2}$	V
Infinite bus voltage angle	0	rad
Line resistance (R)	15	$m\Omega$
Line inductance (L)	710	μH
Micro-source voltage magnitude (V)	$127\sqrt{2}$	V
Micro-source voltage angle	0.01	rad
Capacitance of the filter (C_f)	411	μF
Inductance of the filter (L_f)	171	μH
Resistance of the filter (R_f)	6.4	$m\Omega$
Control Parameters	Value	Units
Frequency droop gain (K_p)	0.05	%
Voltage droop gain (K_q)	0.01	%
Frequency of the low-pass filter (ω_f)	6	Hz
Proportional voltage gain (K_{pv})	1.9	-
Integral voltage gain (K_{iv})	10	-
Proportional current gain (K_{pi})	0.13	-
Switching frequency (f_{sw})	6k	Hz

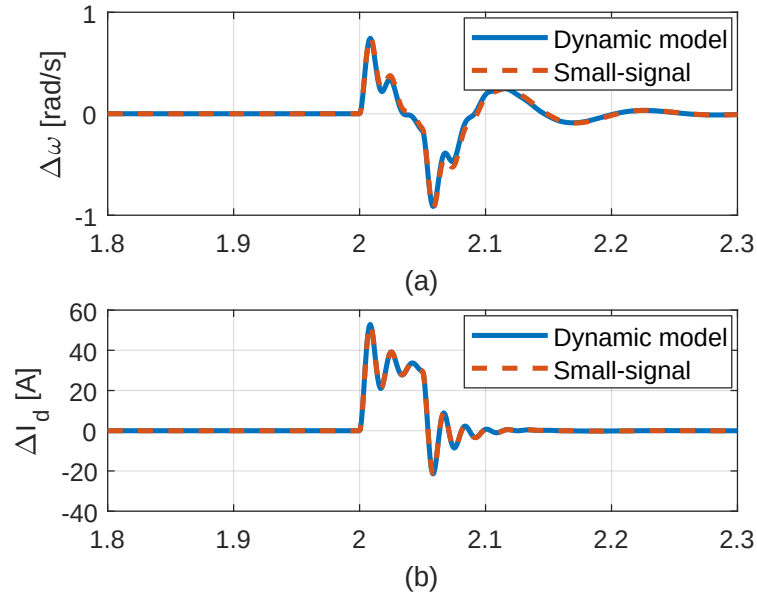


Figure 4.9: Dynamic response of the droop-controlled micro-source under a 5% disturbance in the d -axis component of the bus voltage for 50ms.

gains were conducted in order to validate the small-signal models. The results obtained closely match the results from the dynamic models and to avoid repetitive figures are not shown here.

Figure 4.13 presents the maximum percentage error of the small-signal models compared with the dynamic models for the droop-controlled micro-source, synchronverter and VSG. As illustrated, the small-signal models can reasonably well predict the behavior of the dynamic systems, even in case of large signal variations. Still in Figure 4.13, one can see that for a variation in the d -axis component of the bus voltage equal to 100%, the maximum errors in the small-signal variables are around 12% for

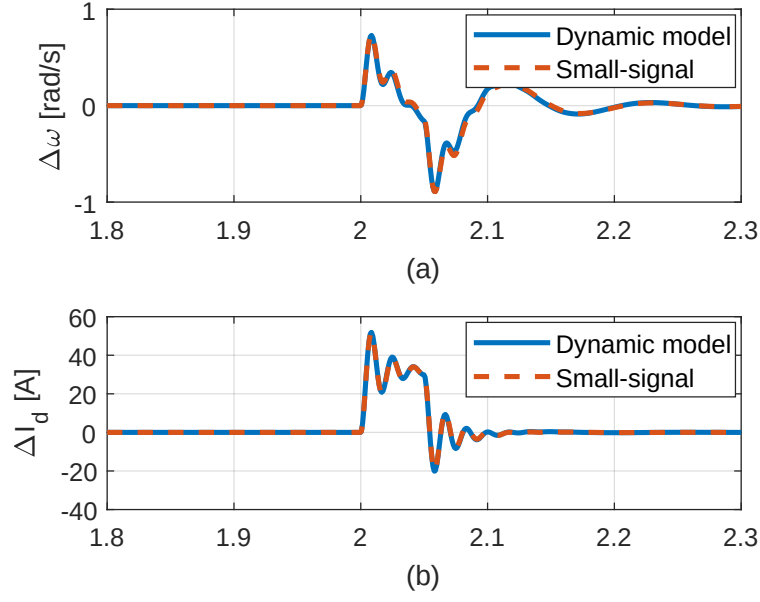


Figure 4.10: Dynamic response of the synchronverter under a 5% disturbance in the d -axis component of the bus voltage for 50ms.

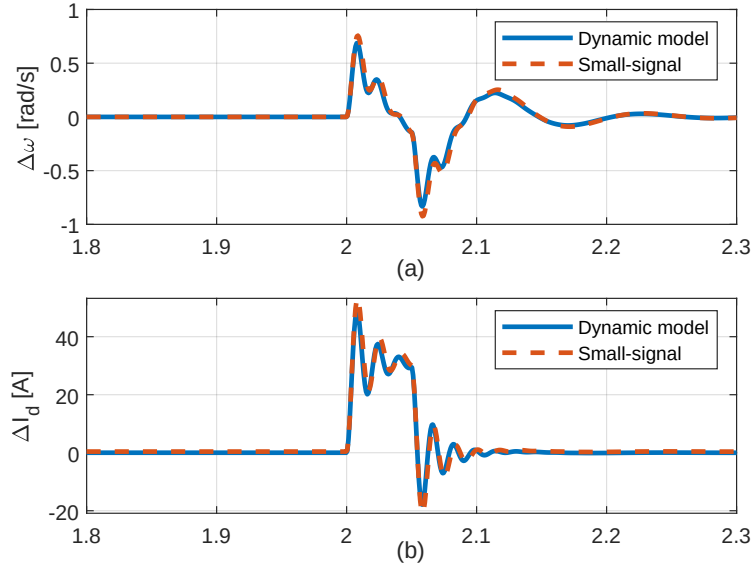


Figure 4.11: Dynamic response of the VSG under a 5% disturbance in the d -axis component of the bus voltage for 50ms.

frequency and 6% for currents for droop control and synchronverter. The accuracy of the VSG for the frequency presented the worst case, in which the maximum error was about 26%.

4.5 Small-signal Dynamic Behavior of the Synchronverter and the Droop-controlled Converter

Figure 4.14 shows the closed-loop poles of the droop-controlled micro-source, synchronverter and VSG in the complex plane for the same parameters as used in Section 4.4. As illustrated, the closed-loop poles

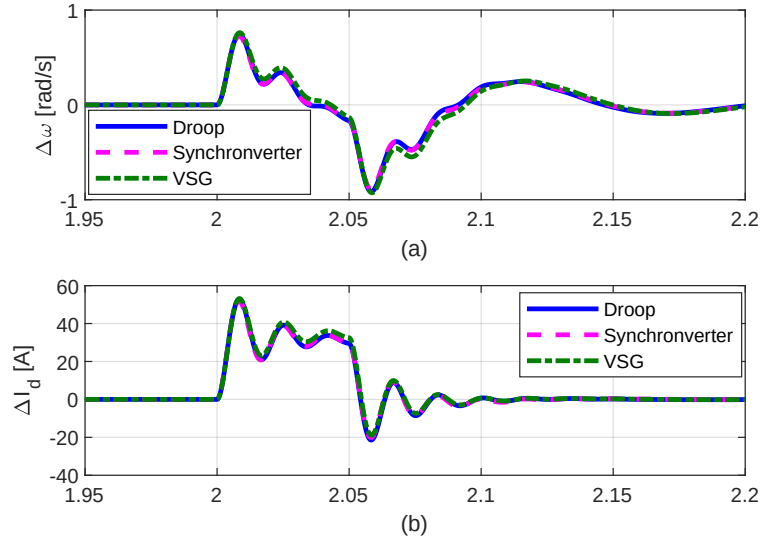


Figure 4.12: Dynamic responses of the droop-controlled micro-source, the synchronverter and the VSG under a 5% disturbance in the d -axis component of the bus voltage for 50ms.

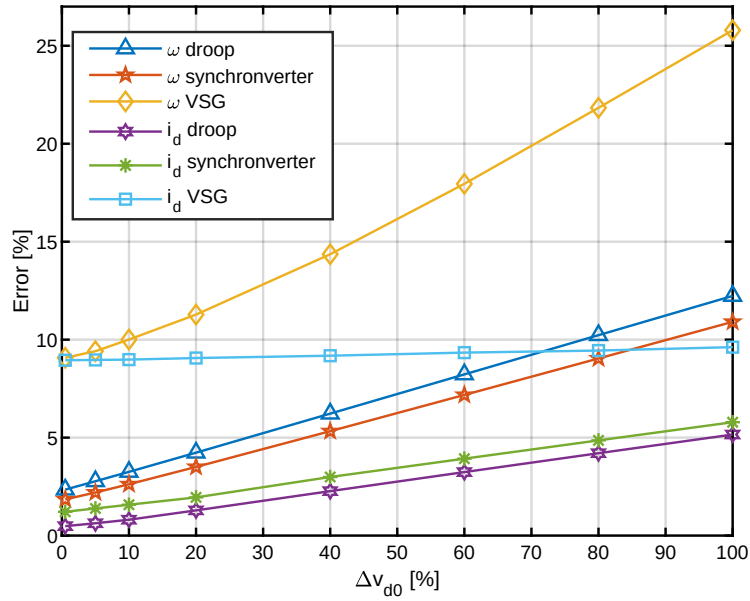


Figure 4.13: Percentage errors of the small-signal models compared with the dynamic model. Droop-control, synchronverter and VSG.

of the droop and synchronverter are very similar for the considered parameters, while the VSG presents on pole closest to the $j\omega$ axis due to the absence of the dynamics of the w_f in the reactive power control loop.

Figure 4.15 shows the variation of the closed-loop poles on the complex plane for the droop-controlled converter, the synchronverter and the VSG, as a function of the increase on gain K_p (0.01 – 0.9%). This range of variation on K_p leads the droop-controlled converter and the VSG into instability, illustrating the fact that these control algorithms presents a narrower stable range of values for K_p , compared to the synchronverter. For the present conditions, the critical value of K_p that leads the droop-controlled converter and the VSG into instability is $K_p = 0.6\%$, while for the synchronverter is $K_p = 1.4\%$. This dynamic difference in favor of the synchronverter is due to the coupling between its power control loops

through the term λ_f .

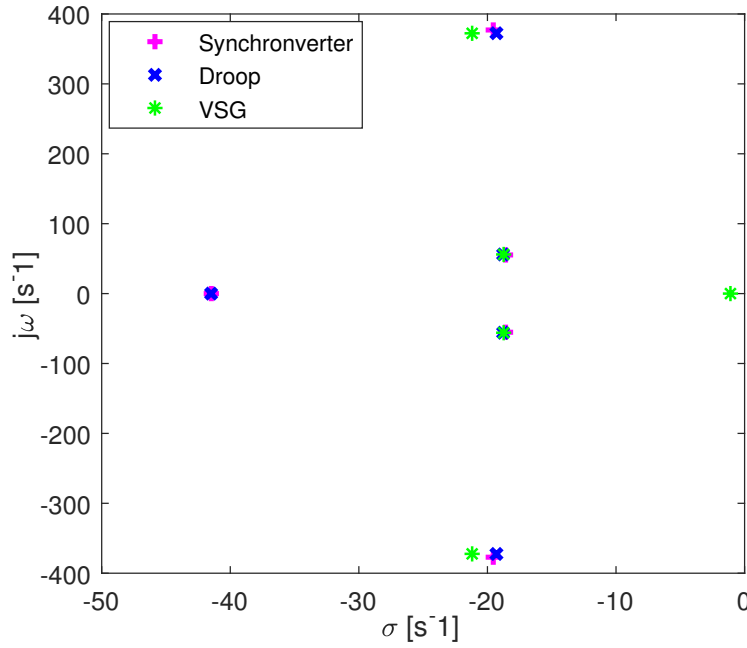


Figure 4.14: Poles of the DQ small-signal model on the complex plane ($K_p = 0.05\%$ and $K_q = 0.01\%$). Droop-control, synchronverter and VSG.

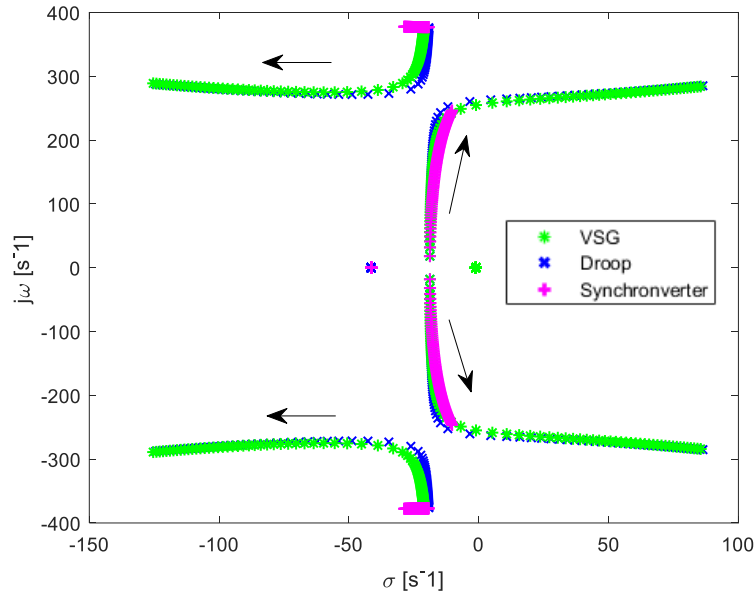


Figure 4.15: Poles of the DQ small-signal model on the complex plane as a function of the gain K_p ($0.01 - 0.9\%$), $K_q = 0.01\%$. Droop-control, synchronverter and VSG.

Figure 4.16 shows the variation of the closed-loop poles in the complex plane for the droop-controlled converter, the synchronverter and the VSG, as a function of the increase on gain K_q ($0.01 - 0.2\%$). As one can see, an increase on gain K_q affects almost identically the closed-loop poles of the droop-controlled

converter and the synchronverter. The behavior of the VSG poles, however, shows that the absence of the w_f in the reactive power control loop, besides adding a more oscillating pole, reduces the displacement of the poles when K_v are increased.

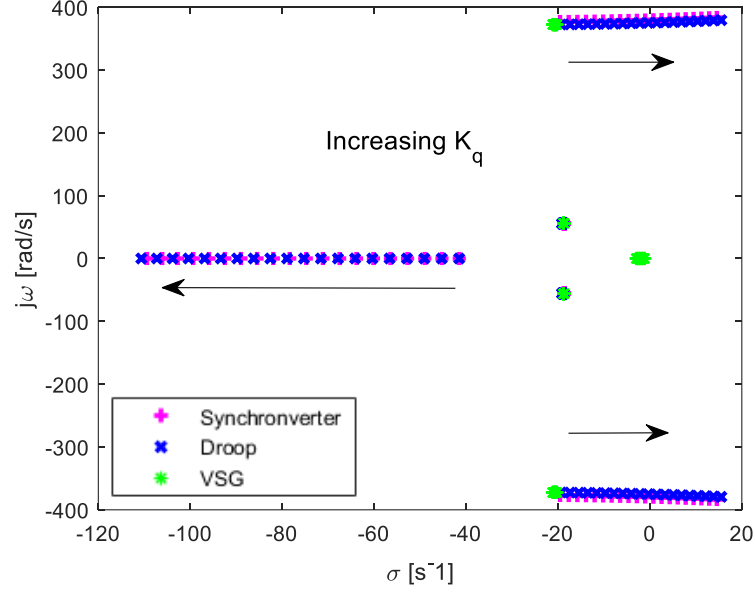


Figure 4.16: Poles of the DQ small-signal model on the complex plane as a function of the gain K_q ($0.01 - 0.2\%$), $K_p = 0.05\%$. Droop-control, synchronverter and VSG.

Regarding the stability of the grid-connected converter under different values of the grid impedance, two conditions are evaluated: one varying the X/R relation and maintaining Z constant and another varying Z and maintaining X/R constant. For the first condition, the considered range of X/R is from 0.5 to 20. Figure 4.17 shows the poles on the complex plane for this scenario. One can see that increasing the X/R ratio can lead to system instability, as the R value decreases. When X/R is kept constant and Z varies from $1m$ to $300m\Omega$, it is observed that the system is more stable for higher impedances. Figure 4.18 illustrates this behavior. Therefore, from the stability analysis of the droop-controlled converter, synchronverter and VSG connected to the grid as function of the grid impedance, some minimum resistance value is needed to guarantee the stability.

Figure 4.19 presents the small-signal admittance of the droop-controlled micro-source, the synchronverter and the VSG. One can see that the admittance of the three control strategies are similar and the differences appears only at the lowest frequencies. The idea of finding the admittance transfer function of the system is based on the Nyquist stability criterion, whereby it is possible to evaluate whether the connection between a converter and the grid will be stable. Thus, having known the admittance of the converter, it is possible to evaluate the stability for different grid impedances.

4.6 Experimental Results

Experimental results are obtained from a Hardware-in-the-Loop model implemented in a Typhoon HIL 600. A TMS320F28335 Texas Instruments micro-controller is used for the control part. Since the results from the droop-controlled converter, the synchronverter and the VSG are similar, only the response of the synchronverter will be presented.

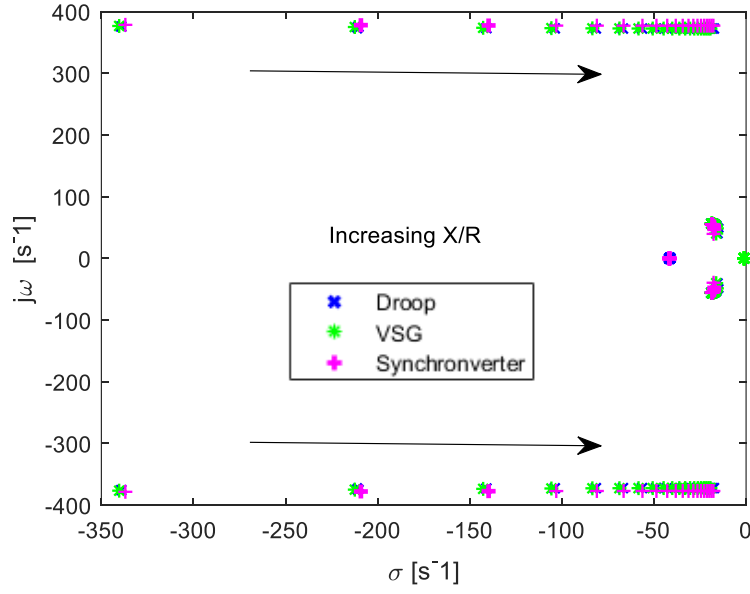


Figure 4.17: Poles of the DQ small-signal model on the complex plane as a function of the X/R ratio (0.5 – 20). Droop-control, synchronverter and VSG.

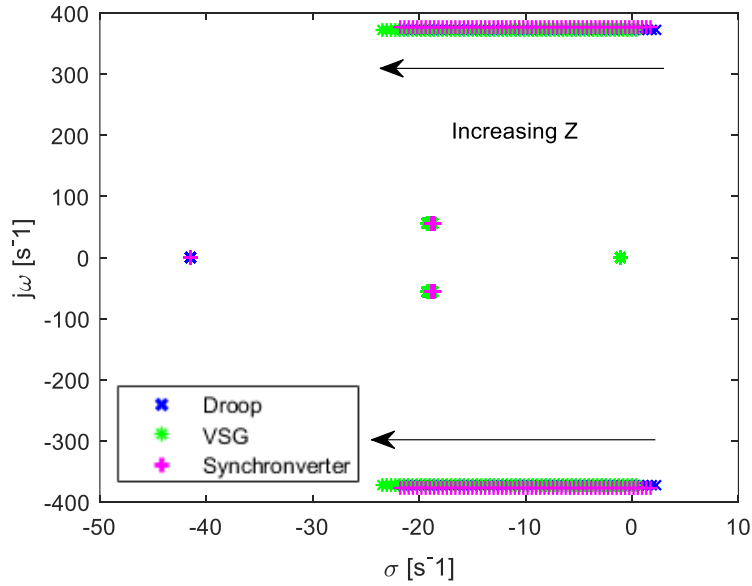


Figure 4.18: Poles of the DQ small-signal model on the complex plane as a function of the Z (1m – 300m Ω). Droop-control, synchronverter and VSG.

The parameters used in the HIL model are the same as the ones used in the simulations. In addition, the same disturbance is also considered. Figure 4.20 shows the frequency and current (i_d) response of the system during the disturbance.

Figures 4.21 and 4.22 illustrate the experimental and the small-signal model responses for the frequency and the d – axis current (i_d) under the disturbance. The curves presented in this figure clearly validate the developed small-signal models.

To evaluate the behavior of the converter against small variations in the active power, an experimental result was obtained, in which ΔP is the input signal and the frequency is the output signal. In this scenario, the active power variation is 2kW during 50ms. Figure 4.23 presents the results. One more

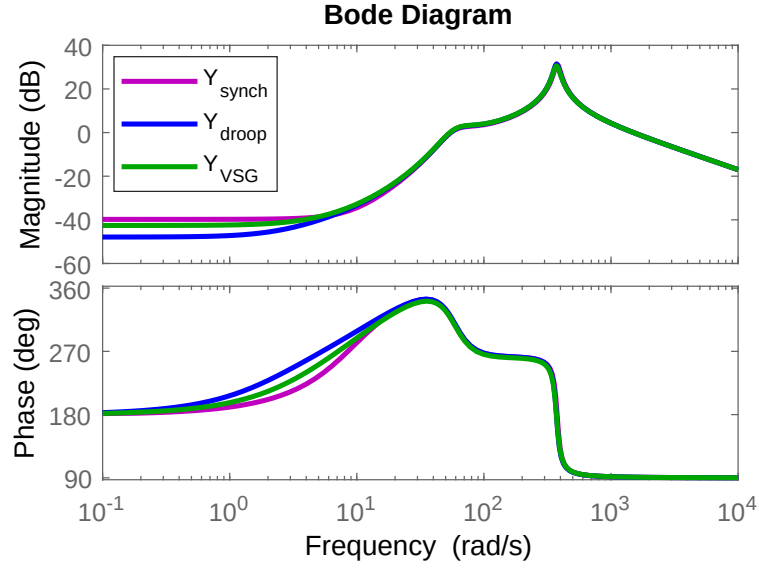


Figure 4.19: Small-signal admittance of the droop-controlled converter and the synchronverter from channel d to channel d (Y_{dd}). $K_p = 0.05\%$. Droop-control and synchronverter.

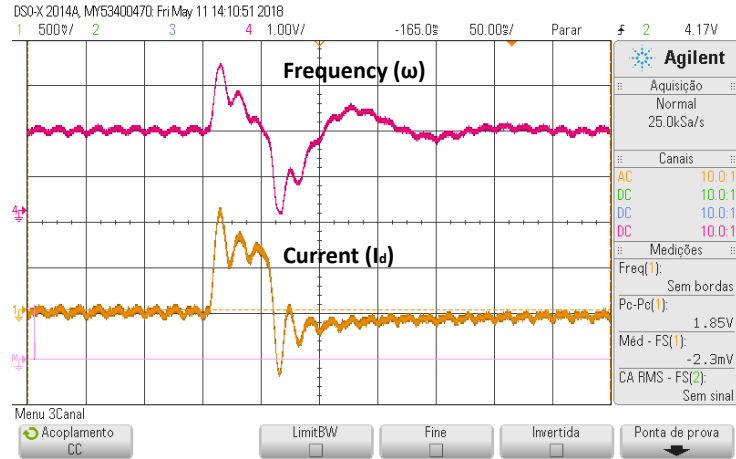


Figure 4.20: Experimental result for the synchronverter: frequency and i_d current under a 5% disturbance in the d – axis component of the bus voltage for 50ms.

time, the dynamic response of the experimental setup and the small-signal model are similar.

4.7 Conclusions

In this chapter, droop control, synchronverter and VSG techniques were evaluated and compared in terms of dynamic operation based on small-signal models.

The small-signal models developed explicitly shows that the difference between the three control structures can be represented by a 2×2 matrix, which can easily be obtained also for other control algorithms for power converters emulating synchronous machines.

The results presented showed that for a considerable range of gains, the control algorithms lead to very similar dynamic responses. However, it was also shown that the synchronverter allows an extended range of gains in the stable region, which gives this control algorithm more flexibility. The

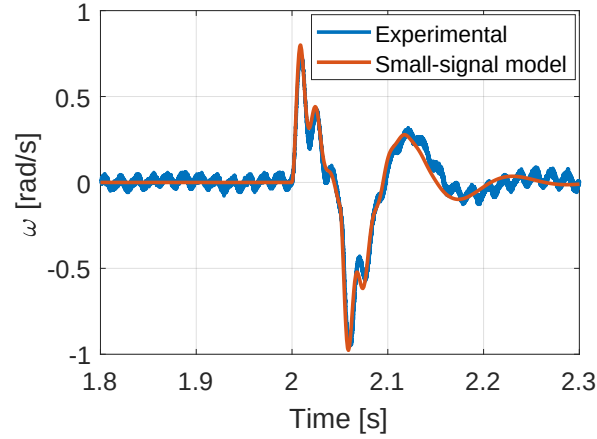


Figure 4.21: Small signal x experimental synchronverter result: frequency under a 5% disturbance in the d – axis component of the bus voltage for 50ms.

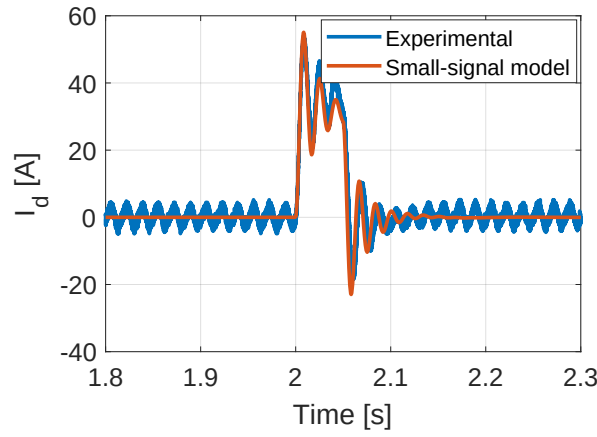


Figure 4.22: Small signal x experimental synchronverter result: current i_d under a 5% disturbance in the d – axis component of the bus voltage for 50ms.

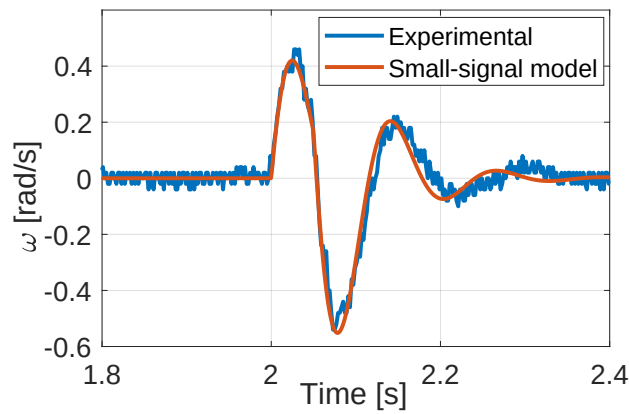


Figure 4.23: Small signal x experimental synchronverter result: frequency under a 2kW disturbance in the active power for 50ms.

simplicity of the droop-control and its more damped dynamic response can be seen as its main advantages over the synchronverter. On the other hand, the VSG carries characteristics of the synchronverter in its reactive power loop, but differs in the reactive power control loop.

5 Positive-negative Sequence Synchronverter with Enhanced Fault Ride-through Capability for AC Grids

THIS chapter proposes a new approach to improve the fault-ride through of three-phase synchronverters connected to grids with unbalanced voltages or subjected to unbalanced voltage sags, based on the Fortescue theory. The Fortescue theory is used to separate positive and negative sequence components from the voltage and the current, allowing to emulate two synchronous machines: one rotating synchronised with the grid and contributing to inject active and reactive power when desired, while the second rotates in the opposite side to compensate unbalanced voltages. Simulation and experimental results validate the proposal and demonstrate the effectiveness: to limit output currents during unbalanced voltage-sags and when the grid is unbalanced, in addition to reducing power oscillations in grid-connected synchronverters, and avoiding unexpected converter shutdown during disturbances.

5.1 Introduction

Grid-connected converters are subject to non-ideal voltage conditions and a number of critical power quality issues must be addressed as part of the overall system reliability.

During the operation of DGs, the electric power system can suffer severe disturbances, such as surges, short circuits, sudden load changes, among others (La Gatta, Passos Filho, and Pereira 2019-Teodorescu, Liserre, and Rodriguez 2007). The most common power quality problem is the voltage sag (McGranaghan, Mueller, and Samotyj 1993), which can lead synchronverters to overload conditions. This condition is due to the fact that synchronverters usually have slow dynamic power responses and tend to keep constant power during sudden voltage variation, which in turn increases their output current values. As a result, the output currents of the VSC-based synchronverters are typically limited to 125% of its nominal value, in order to protect the power electronic switches from overheating (Yazdani and Iravani 2010).

Currently, a photovoltaic (PV) system is required to cease feeding local loads in the occurrence of abnormal grid voltage conditions, e.g., voltage sags and frequency variations (Yang, Wang, and Blaabjerg 2014) as defined by the IEEE Standard 1547 ([“IEEE Standard for Interconnection and Interoperability of Distributed Energy Resources with Associated Electric Power Systems Interfaces”](#) 2018). However, from the next-generation PV systems it is expected that they will be able to provide dynamic grid support in terms of low-voltage ride-through (LVRT) with reactive power injection (RPI), in order to stabilize the

grid in case of failures, avoiding the loss of massive PV generation systems (Hoke et al. 2018). As far as the conventional synchronverter technique is concerned, as presented in 3 a balanced set of voltages are generated at their terminal. During an unbalanced voltage condition, the synchronverter control scheme cannot adjust the output voltages independently, since a balanced synchronous generator is usually being emulated. The possible consequences of this inability are: (i) very unbalanced output currents, which can surpass the maximum current limit and trigger a converter shutdown; (ii) large oscillation in the output power, which causes DC link voltage fluctuation and may degrade the electrolytic capacitors.

In the literature, four control techniques that combine the emulation of the synchronous machine with the use of the theory of symmetrical components are found. In (Zhao et al. 2017), a hierarchical control scheme is proposed to improve fault-ride through (FRT) in a grid-interactive microgrid, in which, a positive-negative sequence droop-control is used in the secondary control level to coordinate the power injection during voltage sags. This control is based on a communication link, which reduces the system reliability.

Another proposal applied to virtual synchronous generators (VSGs) is presented in (Zheng, Wang, and Pang 2019) and aims to keep the output three-phase current balanced and the oscillation in the active power as low as possible. This method contains a positive-sequence current controller that eliminates negative-sequence from the current references, combined with a Decoupled Dual Synchronous Reference Frame Phase-locked Loop (DDSRF-PLL) (Rodriguez et al. 2007), necessary to get the grid angle information. Such approach requires previous knowledge of grid parameters which hampers its applicability.

In Zheng et al. 2018 a control strategy for a virtual synchronous generator under unbalanced voltage conditions is presented. This strategy retains the properties of VSGs and achieves balanced current or reduces the active power oscillation. The control framework is a cascade scheme containing: a traditional VSG algorithm, a novel current reference generator and a proportional–integral (PI)-based double sequence current regulator. Figure 5.1 illustrates the cascade scheme in a block diagram representation.

The traditional VSG algorithm is the so well-known VSG, without any modification. So, all modifications addressed to help the generator behavior under unbalanced voltage conditions are implemented in the current reference generator and in the PI-based double sequence frame current regulator. The first receives the voltage reference from the VSG, uncouples it in the dq reference frame components, and generates positive-sequence current references by controlling the positive-sequence voltage. The *Eqn* block presented in Figure 5.1 refers to the equation:

$$\begin{bmatrix} i_d^{+*} \\ i_q^{+*} \\ i_d^{-*} \\ i_q^{-*} \end{bmatrix} = \frac{1}{(v_d^+)^2 + (v_q^+)^2} \begin{bmatrix} (v_d^+)^2 + (v_q^+)^2 & 0 \\ 0 & (v_d^+)^2 + (v_q^+)^2 \\ N(-v_q^+ v_q^- + v_d^+ v_d^-) & N(v_q^+ v_d^- + v_d^+ v_q^-) \\ N(v_d^+ v_q^- + v_q^+ v_d^-) & N(-v_d^+ v_d^- + v_q^+ v_q^-) \end{bmatrix} \quad (5.1)$$

The PI double sequence frame current regulator is the last structure that control four current terms (i_d^{+*} , i_q^{+*} , i_d^{-*} and i_q^{-*}) to generate two voltage references ($e_d'^*$ and $e_q'^*$) for the SVPWM (Space Vector Pulse Width Modulation).

A VSG control strategy under unbalanced grid voltage conditions is proposed in (Chen et al. 2017) and its blocks diagram representation is illustrated in Figure 5.2. One more time, a voltage and current decomposition to dq frame is applied and the central structure is a miscellaneous composed of active/reactive power control (which carries the characteristics of the traditional VSG) and a negative voltage control signal calculation. With this structure, various voltage control options can be used to achieve different power/current/ quality of a VSG without sacrificing the inertia characteristic.

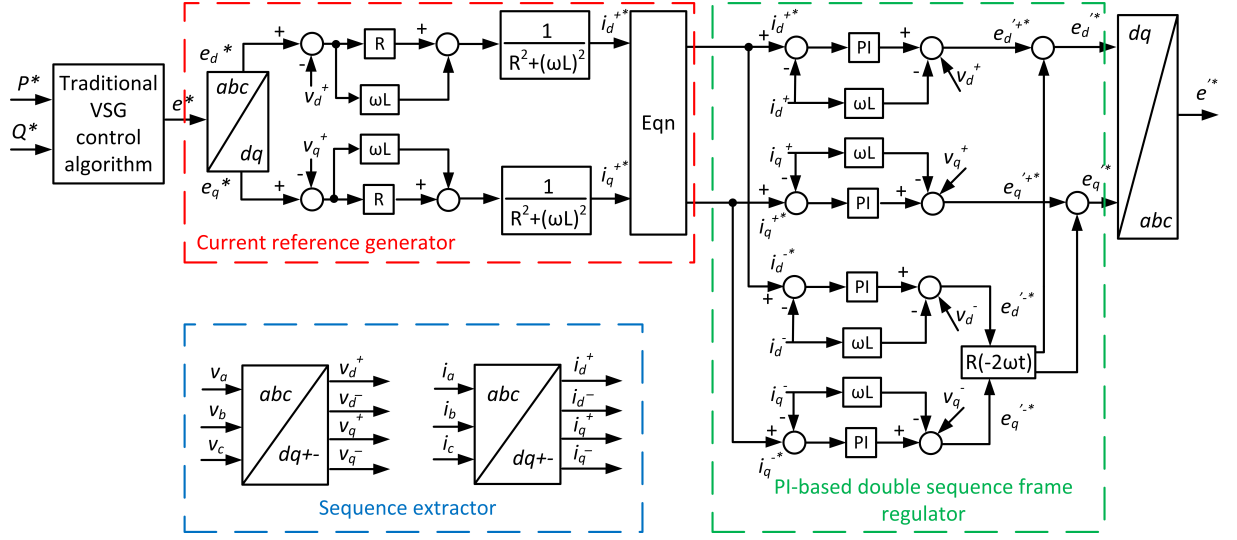


Figure 5.1: Block diagram of the cascaded control framework presented in Zheng et al. 2018

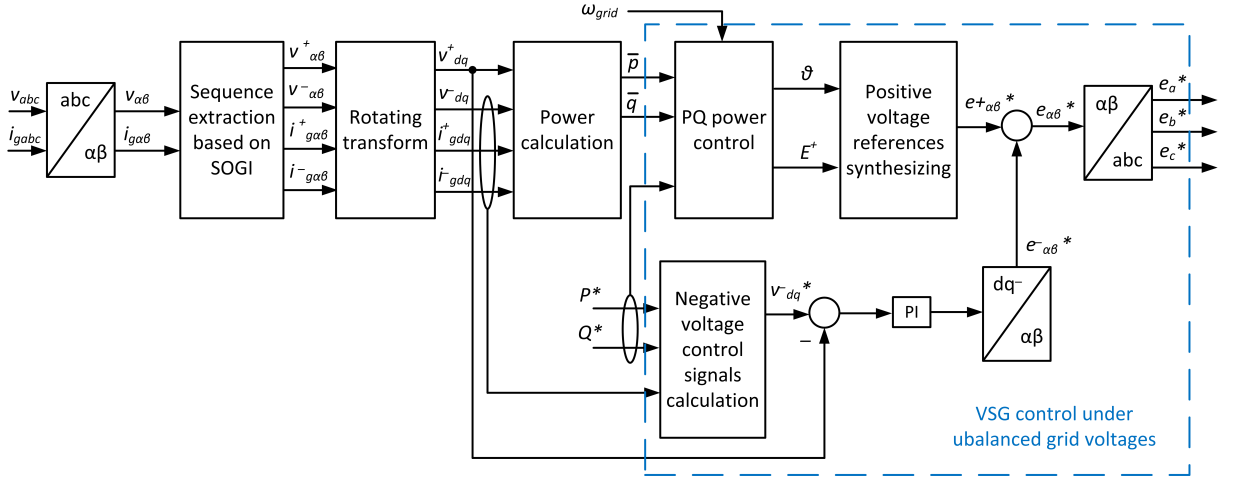


Figure 5.2: VSG control strategy under unbalanced grid voltage conditions.

This chapter proposes a new approach entitled positive-negative sequence synchronverter, to improve the fault-ride through of distributed generators. In this case, a positive-negative sequence method is applied in the primary control level and no communication link is required. The proposal consists in separating the power control into two synchronverter algorithms, working cooperatively: (i) one control-loop for the positive sequence power; and (ii) a second control-loop for the negative sequence power. Thus, if no negative sequence power is desired, the output voltage of the second virtual machine adjusts its reference voltage in order to obtain balanced and limited output currents at the converter terminals. This new method does not require previous knowledge of grid parameters, neither a phase-locked loop (PLL) in its control scheme. The main objectives are to limit the output currents and to reduce power oscillations at the converter terminal during voltage disturbances.

5.2 Positive-negative Sequence Synchronverter

A weak power grid usually has steady-state unbalanced voltages caused by single-phase loads, and suffers from voltage sag that can be symmetrical or asymmetrical. These voltage sags are normally caused by connection or disconnection of heavy loads and start-up of large motors, and faults in the transmission and distribution power lines. In this context, the conventional synchronverter presents some limitations: it is not possible to adjust each phase independently; current can be unbalanced with high magnitude which may cause shutdown of the converter; the active power is oscillatory, which can damage the DC bus capacitors. To overcome these problems, it is proposed a control scheme that separates the PQ control into two virtual machines. Figure 5.3 illustrates the principle of this method, in which "Synch 1" is a virtual machine used to control the positive-sequence variables, while "Synch 2" is used to control the negative-sequence components. As can be seen in the figure, by adding the balanced voltage references from each contribution, it is possible to generate unbalanced voltage references when required.

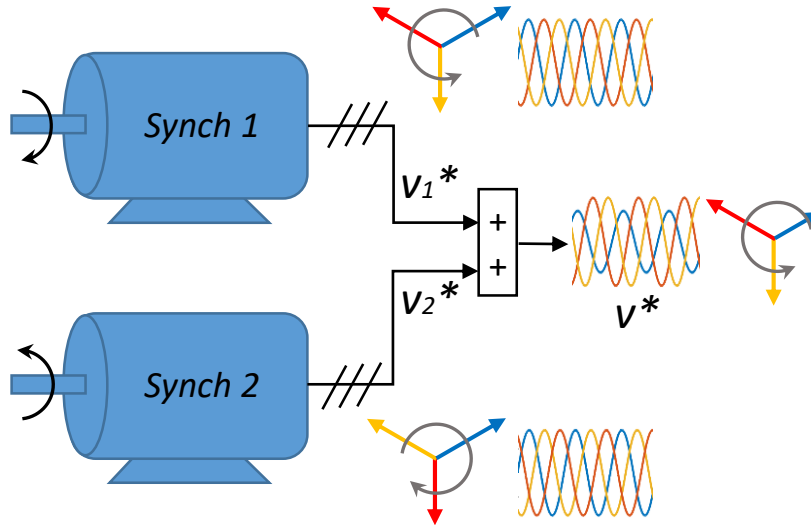


Figure 5.3: Proposed positive-negative sequence synchronverter: two virtual machines rotating in opposite directions to generate unbalanced voltage references.

The positive-negative (PN) sequence synchronverter proposed in this chapter has the same power structured present in the original synchronverter. It basically consists of a primary source, power converter and an inductive-capacitive (LC) output filter. The control part comprises the PN synchronverter algorithm, inner control loops (i.e., voltage and current) and pulse width modulation (PWM). Figure 5.4 shows the PN synchronverter block diagram.

A block diagram of the PN synchronverter algorithm can be seen in the Figure 5.5. First, the measured voltages (v_{abc}) and currents (i_{abc}) at the converter terminals are separated into positive and negative sequence components by sequence detectors, generating (v_{abc1}), (v_{abc2}), (i_{abc1}) and (i_{abc2}). The transformation of Fortescue (Fortescue 1918) is used in the sequence detector block to determine the positive and negative sequences from the three-phase quantities, i.e., voltages and currents:

$$\begin{bmatrix} X_{a1} \\ X_{b1} \\ X_{c1} \end{bmatrix} = \begin{bmatrix} 1 & \alpha & \alpha^2 \\ \alpha^2 & 1 & \alpha \\ \alpha & \alpha^2 & 1 \end{bmatrix} \cdot \begin{bmatrix} X_a \\ X_b \\ X_c \end{bmatrix} \quad (5.2)$$

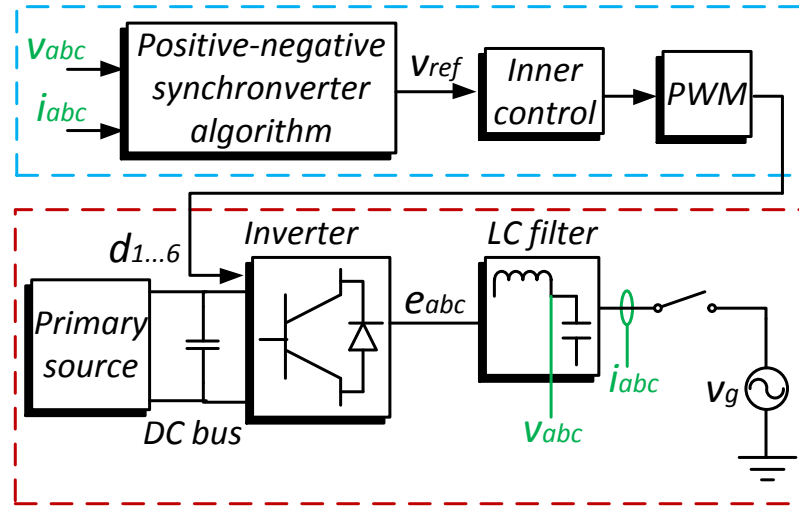


Figure 5.4: Block diagram of the complete positive-negative sequence synchronverter connected to the grid.

$$\begin{bmatrix} X_{a2} \\ X_{b2} \\ X_{c2} \end{bmatrix} = \begin{bmatrix} 1 & \alpha^2 & \alpha \\ \alpha & 1 & \alpha^2 \\ \alpha^2 & \alpha & 1 \end{bmatrix} \cdot \begin{bmatrix} X_a \\ X_b \\ X_c \end{bmatrix} \quad (5.3)$$

in which X is input quantities, index “1” are used for positive sequence components, while index “2” for negative sequence components. The operator α is the rotating phasor equal to $1 \angle 120^\circ$.

The operator α is implemented in the computational model through delays of n samples, which corresponds to 120° of the fundamental component. For example, if the sampling frequency is 10kHz and the grid frequency is 50Hz, we have 200 samples during one cycle of the fundamental component. Therefore, the delay must be 67 samples, which is the closest to 120° .

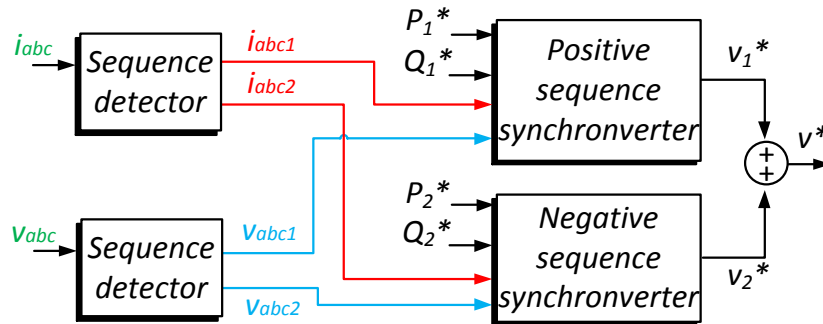


Figure 5.5: Proposed positive-negative synchronverter algorithm: control scheme.

After applying the sequence detectors, the positive and negative sequence synchronverters calculate their voltage reference v_1^* and v_2^* respectively. The sum of both terms corresponds to v^* . One can see in Figure 5.5 that the active and reactive power set-points (P_1^* and Q_1^*) are inputs to the positive sequence synchronverter block, which is the same as the original synchronverter. On the other hand, in the negative sequence synchronverter block, the P_2^* and Q_2^* are the power set-points. In practice, by setting

P_2^* and Q_2^* to zero, the proposed positive-negative sequence synchronverter generates the unbalanced output voltage reference, necessary to account for the unbalance of the grid and generate balanced currents. In comparison with the positive sequence block, the negative block needs to go through some adjustments, as:

- to change the phase sequence;
- to calculate new control gains;
- to replace the voltage control loop with a reactive power control loop.

The first adjustment is due to the negative sequence definition, in which the phasor diagram rotates in opposite direction compared to the positive sequence. The control gains of the positive sequence synchronverter must be tuned equal to the conventional synchronverter. In the negative sequence algorithm, however, it is necessary to recalculate such parameters. The proportional reactive control gain K_{qp2} is calculated from (5.4), while the integral term K_{qi2} can be adjusted to the best dynamic response.

$$K_{qp2} = \frac{0.05\Delta V}{\Delta Q} = \frac{1}{D_{q1}} \quad (5.4)$$

The negative-sequence synchronverter algorithm is shown in Figure 5.6. The active power control is the same as the one used in the positive-sequence algorithm. The voltage control loop must be replaced because it is not possible to predict the voltage magnitude in the negative sequence part. Then, a reactive power control loop based on a proportional-integral controller is applied to. Controlling the reactive power, the voltage magnitude (V_{2m}) is the value necessary to generate the output voltage (v_{abc2}). In the present work, the voltage amplitude in each phase is better adjusted, that is, the PN synchronverter adjusts the affected phases to avoid undesired currents in steady state.

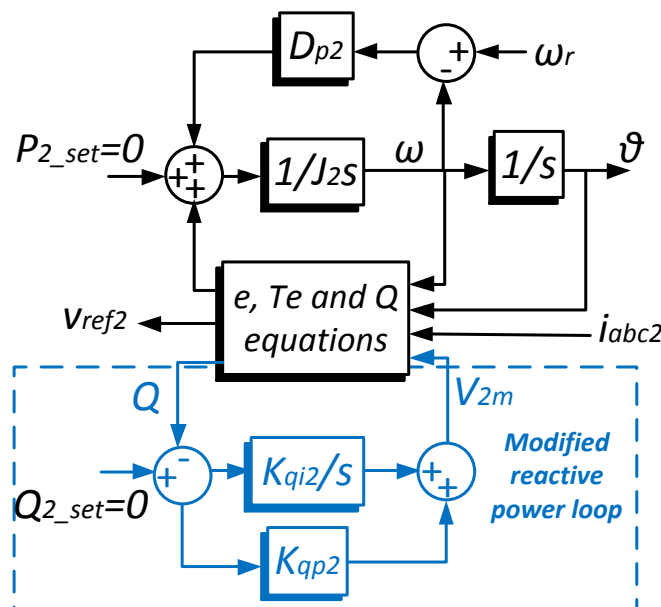


Figure 5.6: Algorithm of the negative sequence synchronverter

5.3 Simulation Results

This section presents simulation results from a Matlab/Simulink implementation. Two scenarios are considered: (i) steady state unbalanced grid voltages; and (ii) unbalanced voltage sag. Table 5.1 shows the simulation parameters used herein.

Table 5.1: Base parameters for simulations of the PN synchronverter.

Grid Parameters	Value	Units
System frequency	60	Hz
Infinite bus voltage magnitude	127	V
Line resistance	15	m Ω
Line inductance	710	μ H
Filter Parameters	Value	Units
Capacitance	411	μ F
Inductance	171	μ H
Resistance	6.4	m Ω
Control Parameters	Value	Units
Friction coefficient (Dp1)	14.07	-
Virtual inertia (J1)	0.3733	-
Voltage droop gain (Dq1)	1114	-
Reactive power gain (K1)	11136	-
Friction coefficient (Dp2)	0.0281	-
Virtual inertia (J2)	7.46 x10-4	-
Proportional reactive power gain (Kqp2)	8.98 x10-4	-
Integral reactive power gain (Kqi2)	5	-
Switching frequency	6k	Hz

5.3.1 Unbalanced Grid Conditions

To evaluate the behaviour of the PN synchronverter under steady state unbalanced voltages, the following values of grid phase voltages are considered: $V_a = 1$ p.u., $V_b = 0.9$ p.u. and $V_c = 1.1$ p.u., corresponding to an unbalance factor (i.e., V_2/V_1) approximately equal to 5.78%. At the beginning of the simulation, the grid voltages are balanced and the active power is set to 2kW. After the instant 2 seconds, the grid becomes unbalanced. The results are shown in Figure 5.7. The grid voltages are shown in Figure 5.7(a). Figure 5.7(b) shows the conventional synchronverter currents, and their significant increase due to the voltage unbalance. On the other hand, the currents of the PN synchronverter method increase during the transient, but decrease in about 100ms. Besides, the current waveforms are practically balanced in steady state, as shown in Figure 5.7(c). Numerically, the unbalance factor for the current (i.e., I_2/I_1), in steady state obtained from the conventional synchronverter is equal to 197.05%, while for the PN synchronverter is equal to 0.01%. The behaviour of the instantaneous active and reactive power is shown in Figure 5.8. The effect of controlling the negative-sequence in the PN synchronverter is highlighted, once the power oscillations are significantly reduced in steady state when compared to the conventional synchronverter under the same conditions.

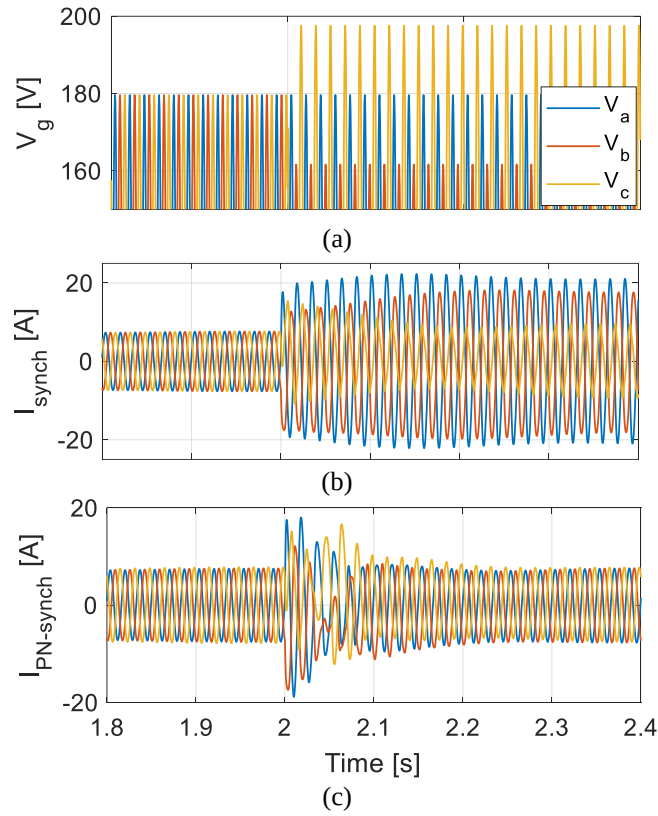


Figure 5.7: Response of conventional synchronverter and positive-negative synchronverter under unbalanced grid voltage: (a) grid voltage, (b) conventional synchronverter currents and (c) PN synchronverter currents.

5.3.2 Unbalanced Voltage-sag

To validate the PN synchronverter under unbalanced voltage sags, a scenario is considered with a single-phase voltage sag where phase “a” drops to 0.8 p.u. from 2 to 3 seconds, what represents an unbalance factor equal to 7.14%. Figure 5.9 shows the three-phase currents for the conventional and PN synchronverters, respectively. One can observe that although the initial transient is similar in both methods, the performance during the steady-state is better controlled by the PN synchronverter, which presents an unbalance factor for the current equal to 0.02%, against 222.94% observed in the conventional synchronverter. This result allows us to conclude that the steady state currents of the PN synchronverter are practically balanced, while in the conventional synchronverter, negative sequence currents are larger than positive sequence currents during the voltage sag. Regarding the power control, Figure 5.10 presents the active and reactive power curves for the conventional and PN synchronverter. The set-points P^* and Q^* are 2kW and 0VAr, respectively. It is possible to observe in Figure 5.10(a) and Figure 5.10(b), as in the previous case, that during the voltage disturbance, the powers of the PN synchronverter are practically constants after the transient time, while the powers of the conventional synchronverter oscillate about twice the power set-point.

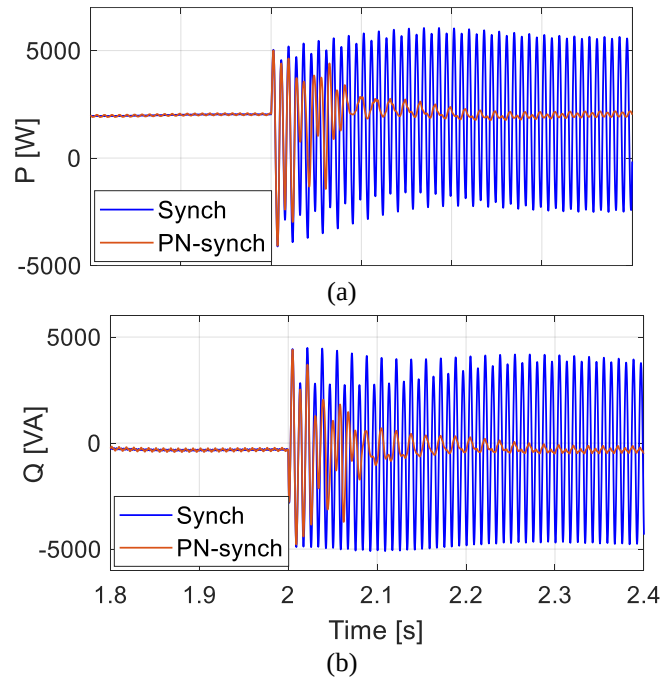


Figure 5.8: (a) Active and (b) reactive power for both conventional synchronverter and PN synchronverter under unbalanced grid voltage.

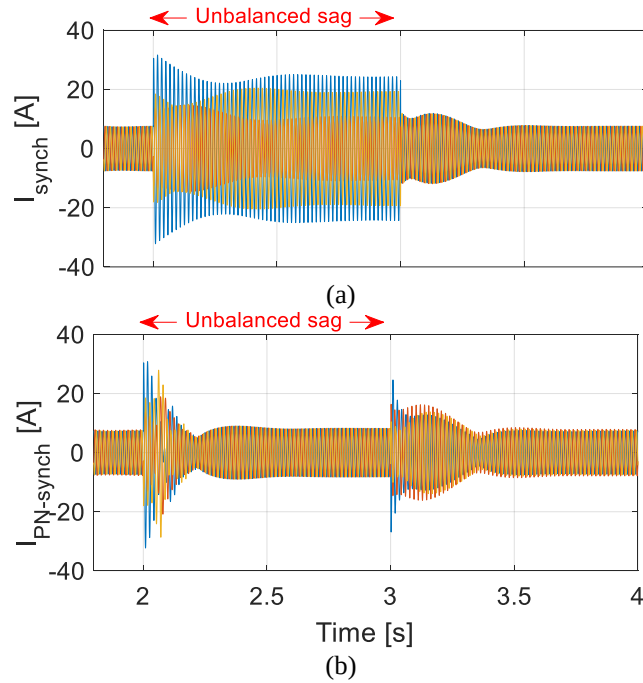


Figure 5.9: Response of conventional synchronverter and positive-negative synchronverter under a single-phase voltage sag: (a) conventional synchronverter currents and (b) PN synchronverter currents.

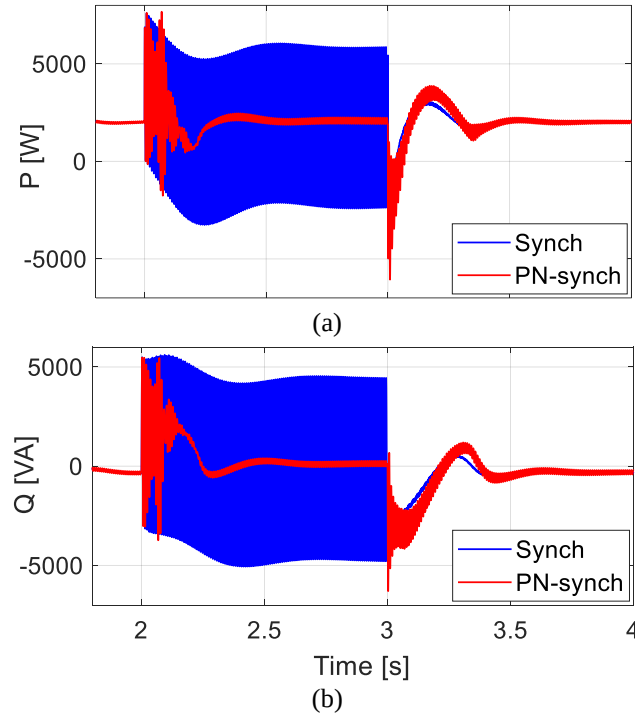


Figure 5.10: (a) Active and (b) reactive power for both conventional and PN Synchronverter under single-phase voltage sag.

5.4 Experimental Results

Experimental results were obtained from a Hardware-in-the-Loop, Typhoon HIL 600 setup. A Texas Instruments microcontroller (TMS320F28335) is used in the control part, as illustrated in Figure 5.11. The power systems and control parameters are the same as those used in the simulation. Therefore, the goal here is to validate the proposed method through experimental results considering the time delay of the A/D converter, and inherit characteristic of digital implementation such as quantization, resolution, memory, etc. Herein, only the unbalanced voltage condition is considered to avoid presenting excessive results.

Figure 5.12 presents the behaviour of conventional and PN synchronverter when the grid voltage is unbalanced with $V_a=1$ p.u., $V_b=0.9$ p.u. and $V_c=1.1$ p.u. In Figure 5.12, it is shown that the synchronverter currents are unbalanced and the current magnitude increases in at least two phases. On the other hand, in the PN-synchronverter shown in Figure 5.13, the currents are balanced and their magnitudes in steady state are balanced. The transient behaviour of the PN-synchronverter is shown in Figure 5.14. As expected, when the grid voltage changes from balanced to unbalanced condition, the currents increase at the beginning of transient, but they are regulated by the PN-synchronverter algorithm in steady state. In this case, the synchronverter is operating with 4kW.

5.5 Conclusions

This chapter proposed the positive-negative synchronverter in order to enhance fault-ride through in synchronverters connected to the grid. Based on the results, the proposal presents good performance when submitted to unbalanced conditions. The PN-synchronverter can be considered an effective solu-

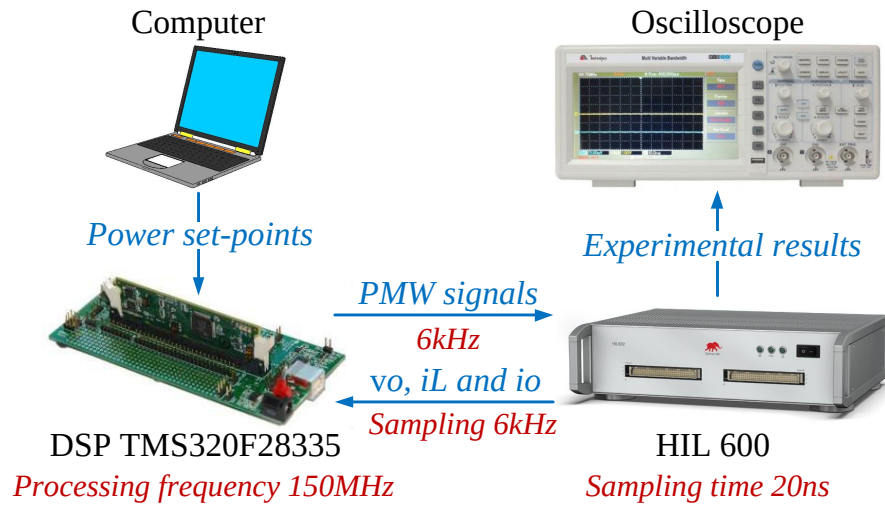


Figure 5.11: Microcontroller and Hardware-in-the-Loop used to obtain the experimental results.

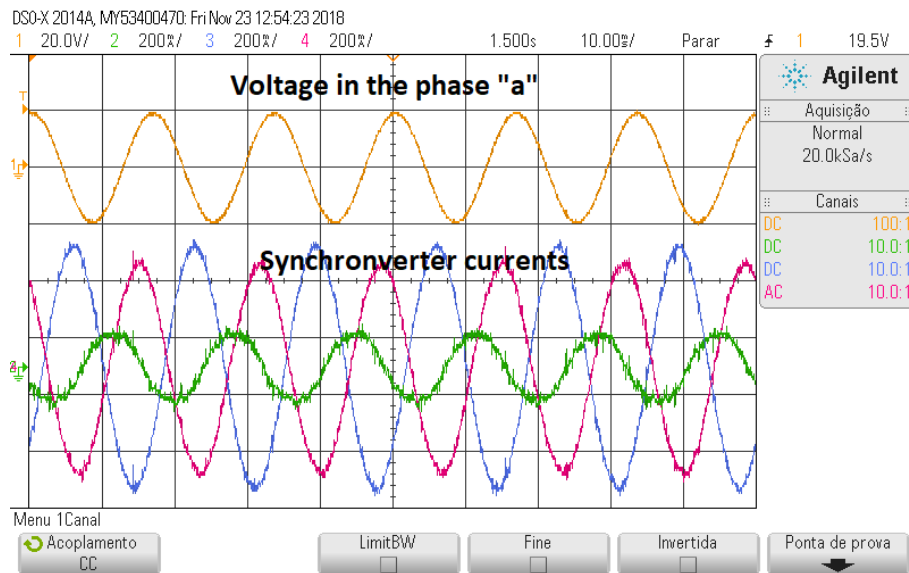


Figure 5.12: Experimental results: conventional synchronverter under unbalanced grid voltage.

tion to enhance FRT in converters emulating synchronous machines in distributed generation applications. The main conclusions on this new method are:

- the PN synchronverter is able to limit output currents during unbalanced voltage-sags and when the grid is unbalanced;
- the proposal allows to reduce power oscillations in grid-connected synchronverters when submitted to grid disturbances;
- the use of PN synchronverter helps to avoid unexpected converter shutdown due to disturbances.

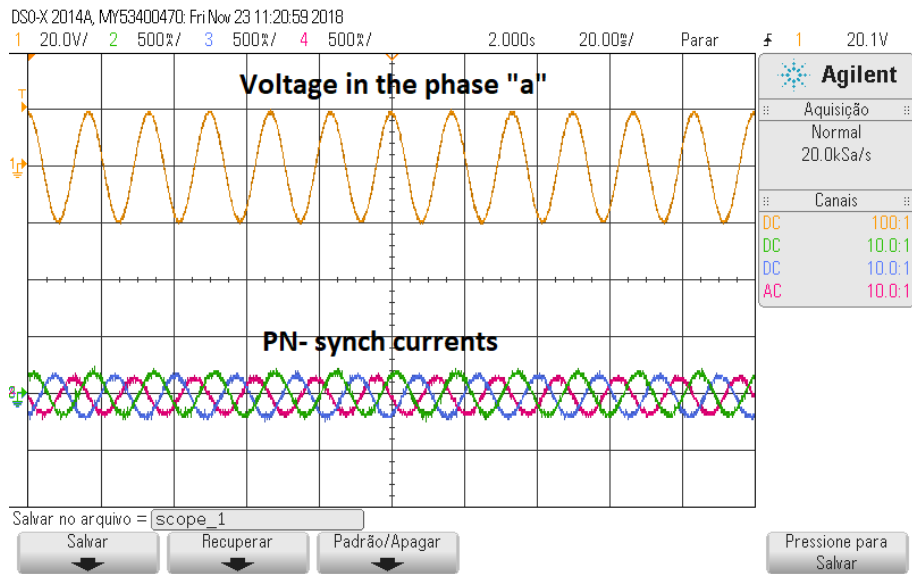


Figure 5.13: Experimental results: PN-synchronverter under unbalanced grid voltage.

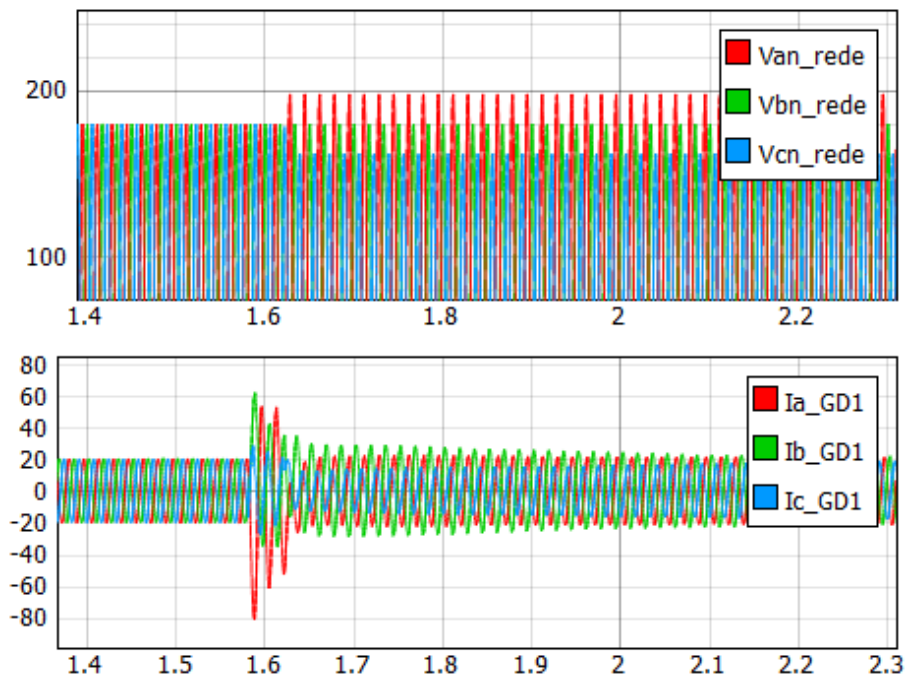


Figure 5.14: Grid voltage and output current of the PN-synchronverter in the transition from balanced to unbalanced grid voltage condition.

6 Single-phase Synchronverter for Residential PV Power Systems

IN this chapter a single-phase synchronverter model is proposed for application in distributed generation and micro-grids. The power and control parts are explained, then simulation results are presented. Results illustrate the behavior of the proposal in normal voltage, voltage sag and harmonic conditions.

6.1 Introduction

Single-phase generation systems have received increasing attention due to the growing market of residential photovoltaic generators, small wind turbines and electric vehicles (Zeng et al. 2013), for rated power less than 15kW, for example.

The first use of the term "single-phase synchronverter" was proposed by (Ming and Zhong 2014), with the application of the synchronverter control part for inverters in half bridge, with accessible neutral, applicable to photovoltaic generators. The model, which does not have a DC/DC converter, presents as a limitation the need for the terminal voltage of the photovoltaic panels to be greater than the peak value of the grid voltage.

In September 2016, Misra et. al (Mishra et al. 2016b) proposed a control scheme for PV converters emulating the synchronous machine, for which they use the term single-phase synchronverter. The authors claim that the main contributions of the proposed control method are: the control of the single-phase converter without the use of a PLL, the converter responds to deviations in the grid frequency through the emulation of synchronous machine and the correction in the variations of the grid voltage by means of the reactive power injection. In other words, a single-phase synchronverter in this case is also used as a grid-support converter.

The paper entitled "Single-phase synchronverter for residential PV power systems" (Ferreira et al. 2016a) was presented in a IEEE conference in October 2016 and it proposes a new way of implementing the single-phase synchronverter, which has as objective to establish the least number of modifications in the original three-phase model to allow its application in residential PV micro-generators. This single-phase proposal emulates three-phase currents to avoid changing the method of calculating T_e , e and Q . The generated voltage, e , of the single-phase synchronverter is not a vector as in (3.13), but a scalar quantity. The work also evaluates the performance of the single-phase synchronverter under voltage sag conditions and harmonic distortions in the distribution grid. It is important to note that until the paper was accepted, the authors were only aware of the reference (Ming and Zhong 2014), since the other works had not been published yet.

In December 2016, the paper (Younis et al. 2016) appears as the last reference, which proposes to present a way of implementing single-phase synchronverter. In this case, the method does not require the use of a PLL and acts by minimizing a limitation of the voltage controlled converters, which is the output over-current susceptibility to disturbances in the grid voltage. Thus, the proposed method establishes a maximum limit for the current injected into the grid.

This contextualization basically clarifies that the synchronverter implementation technique presented in this chapter is not unique and was proposed at a time when several research groups developed correlated and non-interfering works on one another.

6.2 Single-phase Synchronverter Description

The objective of this section is to present a single-phase synchronverter in comparison with the three-phase synchronverter, applicable to residential PV power systems. The present approach is not the same mentioned in (Ferreira et al. 2016a), but a new proposal that eliminates the use of the PLL. The section is organized in power part and control part. Simulation results allow the evaluation of the operation of the algorithm in normal grid conditions, voltage sags and harmonic distortions in the grid to which the inverter is connected.

The power part is basically a single-phase inverter with a LC output filter to interface the converter to the grid as illustrated in Figure 6.1. Voltage and current are measured in the capacitor (C_s) and inductor (L_s) respectively. The quantity Z_g represents the grid impedance.

The control part is an adapted version of the three-phase synchronverter to single-phase conditions and it is based on transform single-phase measurements of voltage and currents in $\alpha\beta$ components. The idea is to remove the PLL, because of its dynamics and computational cost and only use a transport delay corresponding to 90 degrees. Figure 6.2 presents this new control scheme in which the modifications in comparison with the three-phase synchronverter can be seen in red. Basically, the transport delay already mentioned is used to emulate the component β from voltage and current measured and the quantities T_e , Q and e are calculated from (6.1), (6.2) and (6.3). This new approach has a different behavior when compared with the original concept of the synchronverter: T_e is not dependent of the $M_f i_f$, so active and reactive power control loops are now decoupled.

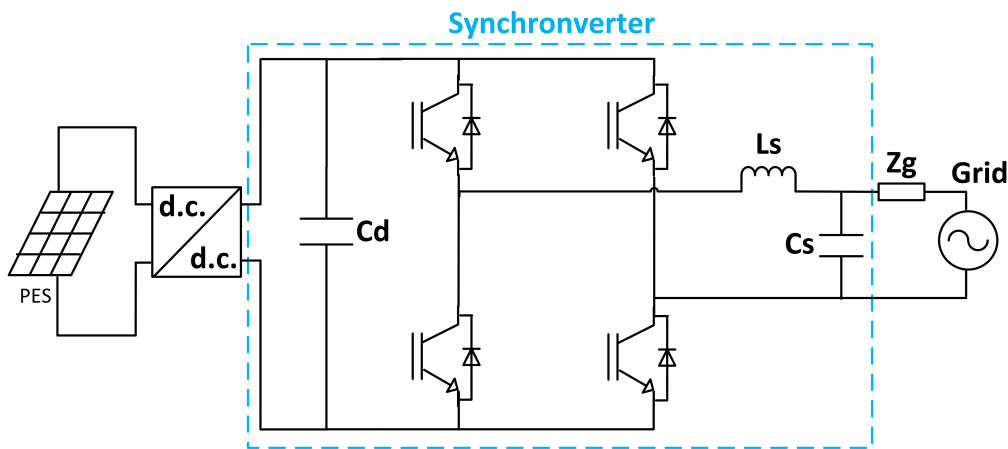


Figure 6.1: Power diagram of the single-phase synchronverter.

$$T_e = \frac{1}{\omega_n} (v_\alpha i_\alpha + v_\beta i_\beta) \quad (6.1)$$

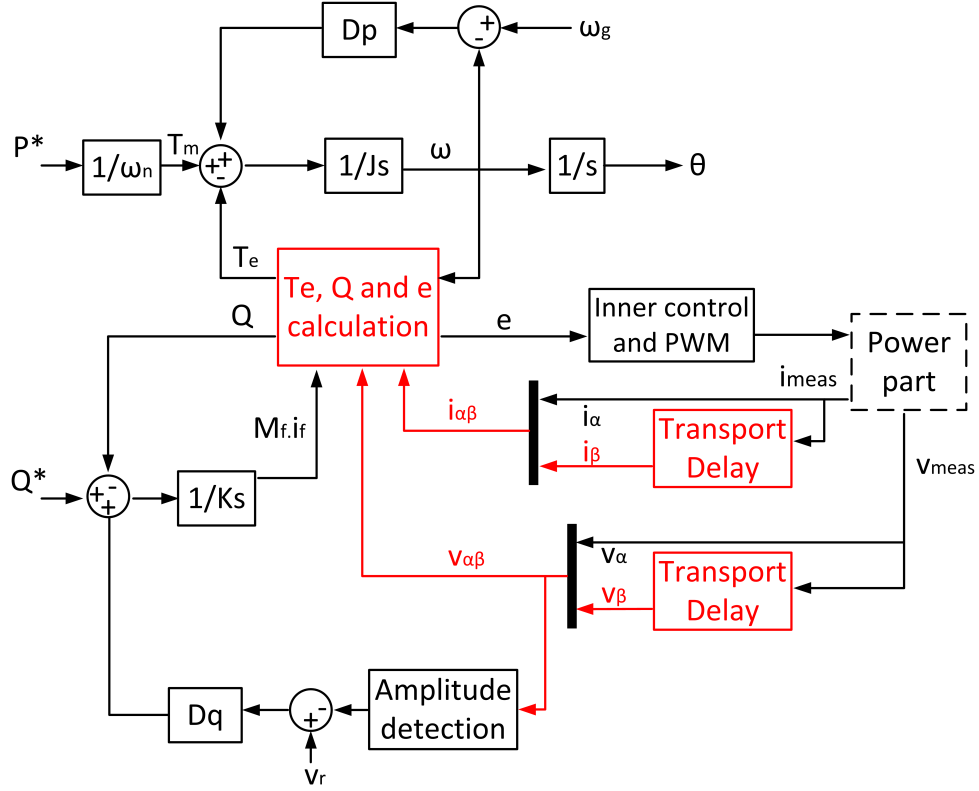


Figure 6.2: Control part of single-phase synchronverter.

$$Q = v_{\alpha}i_{\beta} - v_{\beta}i_{\alpha} \quad (6.2)$$

$$e = M_f i_f \omega \sin(\theta) \quad (6.3)$$

6.3 Simulation Results

This section presents simulation results that allow a performance evaluation of the proposed single-phase synchronverter. All the simulations were developed in the Matlab/Simulink, R2015b. For the sake of simplicity and without loss of generality, the converters were modeled as controlled voltage sources. Simulation parameters were defined in Table 6.1. It is observed that a three-phase grid is simulated with R/X ratio equal to 3, or in other words, the system impedance has a resistive characteristic, typical of low voltage grids.

Figure 6.3 shows the dynamic response of the single-phase synchronverter under grid voltage without any disturbance (ideal conditions). At $t=1.5s$ the active power reference is set to 10kW and at $t=2.0s$ the reactive power reference is set to 5kVAr. Again, the two power variables are coupled during the transient state, but not affected in steady state. Figure 6.4 presents the voltage and frequency, showing that the voltage increases about 5% to provide the desired power and the frequency oscillate about 0.5Hz during the transient time.

With respect to the performance of the single-phase synchronverter against power quality problems, Figure 6.5 shows the simulation results for the active and reactive powers for a voltage sag with 0.75 p.u. residual voltage, starting in $t=3s$ and finishing in $t=4s$. Note that during the transitions the system

Table 6.1: Simulation characteristics for the single-phase synchronverter.

Grid parameters		Synchronverter parameters	
RMS voltage	220V	Rated active power	10kW
Frequency	60Hz	Rated reactive power	5kVAr
Phase number	3	Phase number	1
Lg	5.34 μ H	Ls	0.225 mH
Rg	6 m Ω	Rs	13.5 m Ω

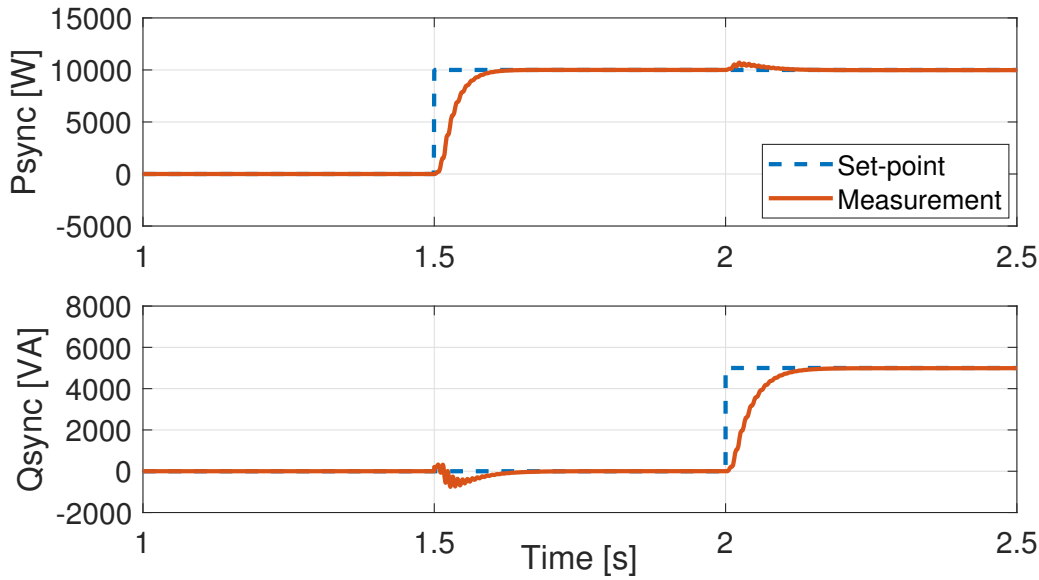


Figure 6.3: Response of the active and reactive power control of single-phase synchronverter in the connected mode.

presents a transient behavior and in steady state, when the voltage grid is 0.75 p.u., the reactive power presents an error of about 5kVAr because the reactive control gains are not adjusted to provide large variations in the voltage magnitude. From Figure 6.6, it is possible to observe that the output voltage (V_{sync}) during the voltage sag is near to 160V, that corresponds to 0.89p.u., therefore, it is clear that the converter cannot maintain an output voltage such that it gets zero error in reactive power. Regarding the current we observe that during the two first cycles after the beginning of the voltage sag, the current increases from 90A to 300A, because no current limit has been imposed. After this transient time, the current is slightly increased to provide the power reference.

Figure 6.7 shows the voltage grid (V_{grid}), the synchronverter voltage (V_{sync}) and its output current (I_{sync}), during the transition from an ideal conditions to a voltage sag. It is possible observe that when the voltage sag starts, a transient behavior in the synchronverter current and voltage occurs, lasting for approximately five cycles. In steady state, the current is increased to provide the power desired.

In the simulation of the operation of the proposed single-phase synchronverter during harmonic conditions, the voltage grid is initially composed of 1.0p.u. of magnitude and fundamental frequency. From $t=3s$ to $t=4s$, 0.2p.u. of the fifth harmonic and 0.15p.u. of the seventh harmonic are included. Figure 6.8 shows the power performance of the single-phase synchronverter under these harmonic conditions. Note that the algorithm is able to work in this harmonic simulation, providing the power set

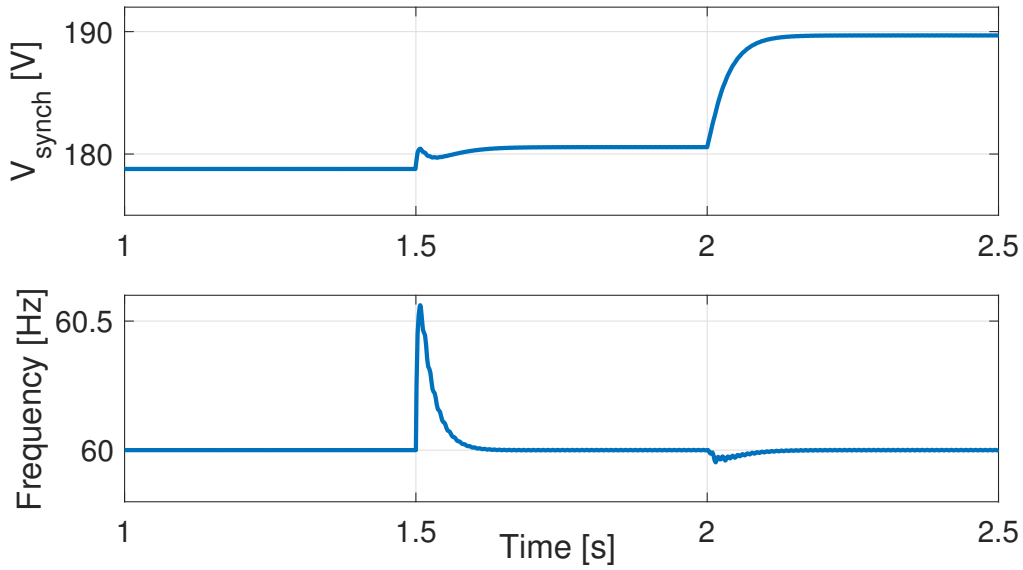


Figure 6.4: Voltage and frequency of single-phase synchronverter in the connected mode.

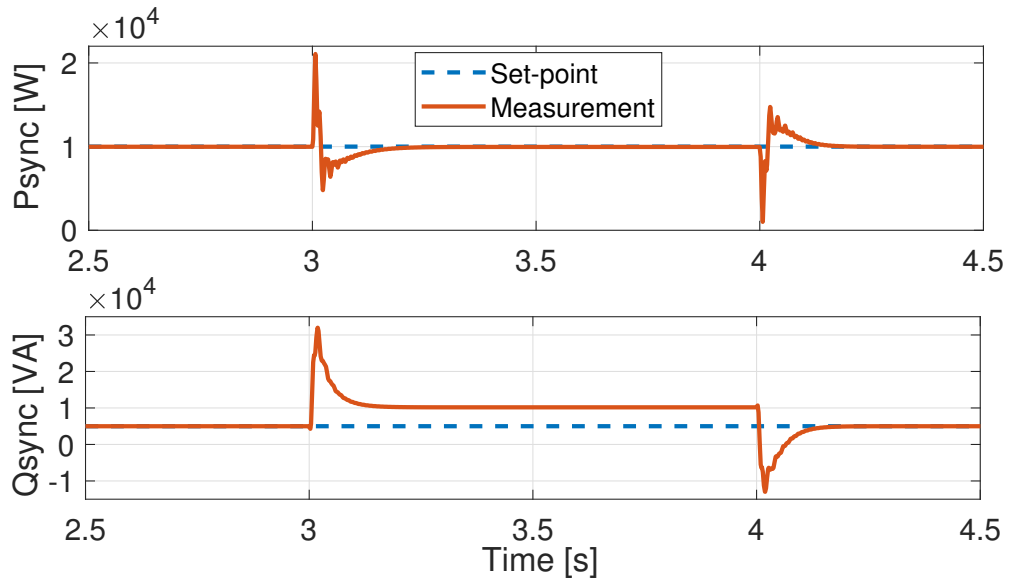


Figure 6.5: Response of the active and reactive power control of single-phase synchronverter front to 0.75 p.u. voltage sag.

and showing transient behavior only in about 12 cycles (0.2 seconds).

Figure 6.9 presents the voltage grid (V_{grid}), the synchronverter voltage (V_{sync}) and the output current (I_{sync}), during the transition from ideal conditions to harmonics. Note that after $t=3s$, although V_{grid} is composed by fundamental frequency, fifth and seventh harmonics, V_{sync} is sinusoidal, causing distortion in the I_{sync} through the grid power line impedance.

The islanded mode is simulated and the results include a transition from connected mode to islanded mode at $t=2.5s$. Figure 6.10 shows the active and reactive synchronverter powers. The load has the same nominal power of the DG. Note that the synchronverter is able to change the operation mode and provide active and reactive powers as required by the load. Active and reactive powers present

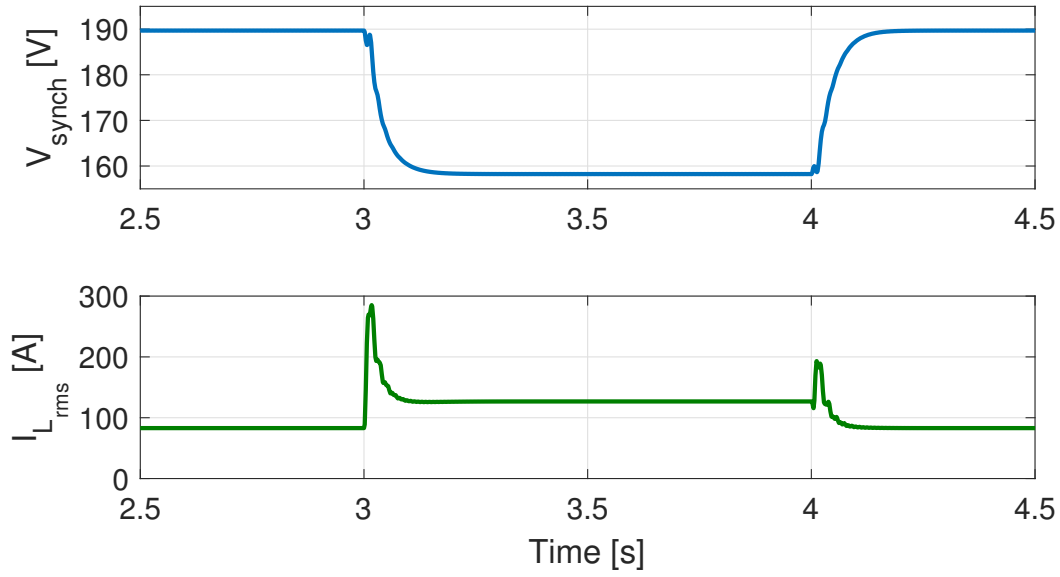


Figure 6.6: Magnitude of voltage and current rms of single-phase synchronverter under 0,75 p.u. of voltage sag.

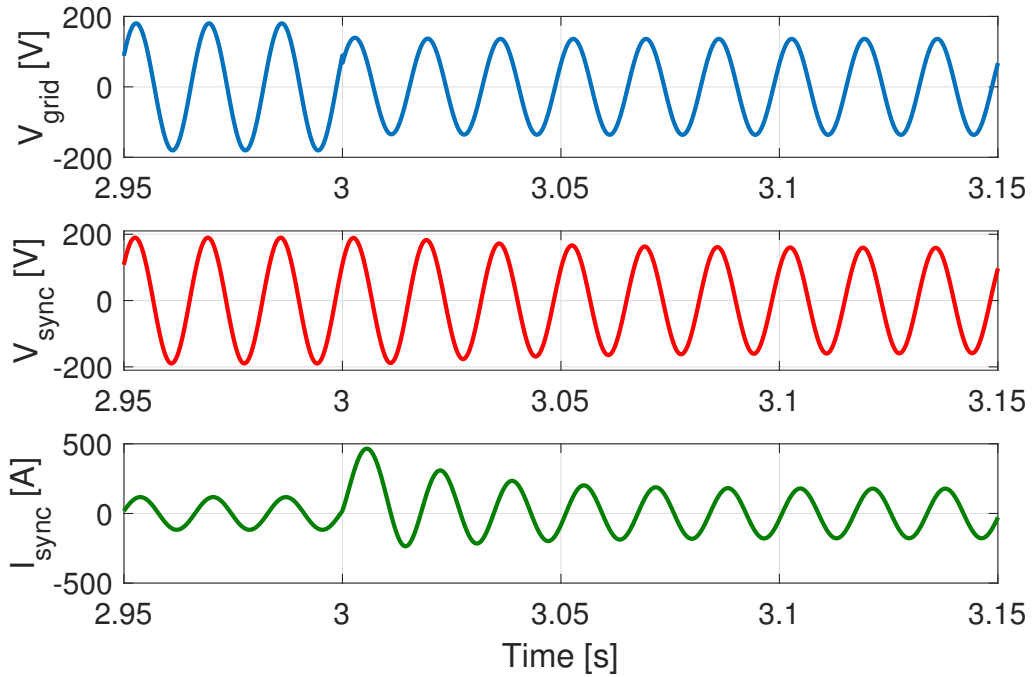


Figure 6.7: Voltages and current behavior of single-phase synchronverter under voltage sag.

errors of about 7% and 14% respectively, because voltage and frequency have new operation points in the islanded mode.

Figure 6.11 shows the frequency and voltage of the synchronverter during the transition from connected mode to islanded. After the breaker opens, the system finds a new operation point and is stable in steady-state. As an external loop was not included to regulate precisely the voltage magnitude and frequency, both quantities show a deviation causing the observed error in P_{sync} and Q_{sync} .

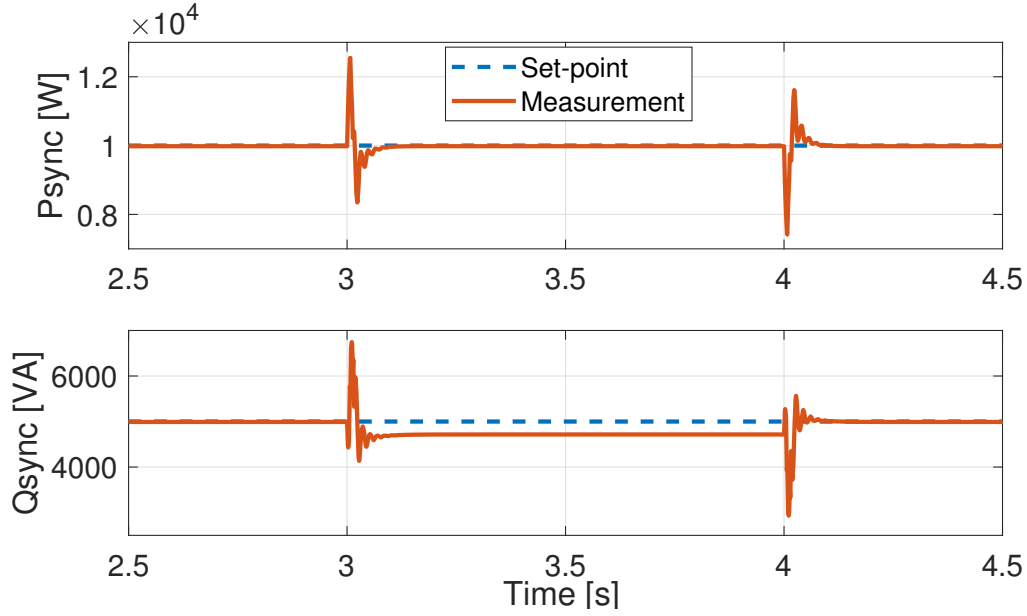


Figure 6.8: Response of the active and reactive power control of single-phase synchronverter under harmonics.

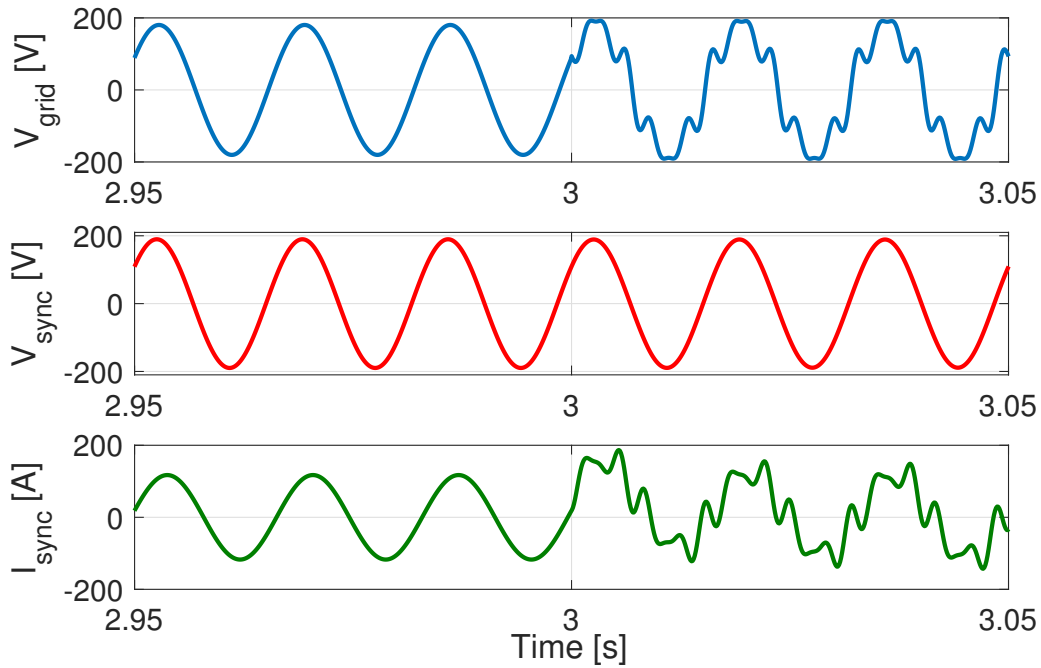


Figure 6.9: Voltages and current behavior of single-phase synchronverter under harmonics.

6.4 Conclusions

A new single-phase model is proposed for the synchronverter. The proposed changes in the control part eliminate the PLL and the coupling between active and reactive power. In steady state, the method has shown to be effective in the nominal conditions.

The simulation results have shown that the method properly works under voltage sags conditions and in the presence of harmonics. The proposed single-phase synchronverter is able to operate in the

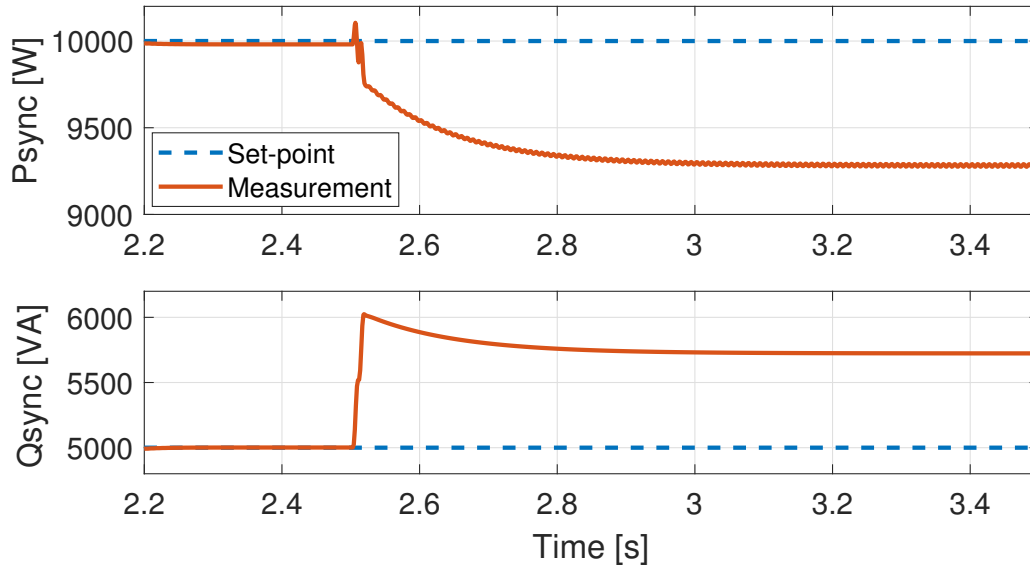


Figure 6.10: Response of the active and reactive power control of single-phase synchronverter in the island mode.

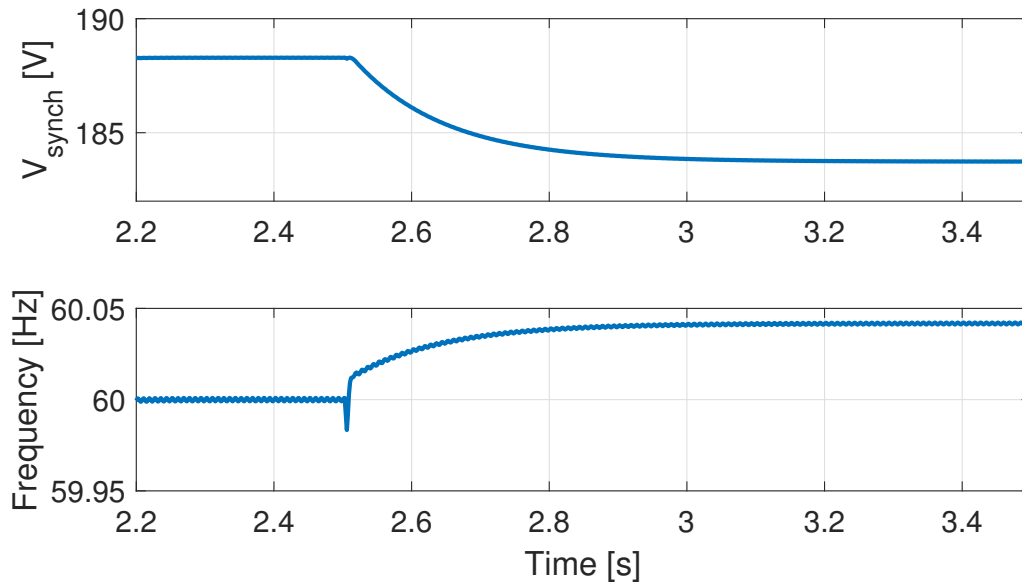


Figure 6.11: Voltage and frequency of single-phase synchronverter in the island mode.

islanded mode, including in the grid transition time. Therefore, the authors believe that the proposed method is a good alternative for PV residential generators, because it has the inherent advantages of the three-phase synchronverter applicable to single-phase systems.

7 Final Remarks and Future Works

7.1 Final Remarks

First, this dissertation revised the converters in distributed generation and micro-grid applications. The operation modes, the classification of the converters according to the function and the explanation of the hierarchical control levels in micro-grids are the main topics covered. The power control techniques based on synchronous machines emulation were studied in the sense of understand the state of the arts. The VISMA, VSG and synchronverter are control techniques that have as principle to emulate the synchronous machine in the electronic power converter. The VISMA is current-controlled, while the other two are voltage-controlled. A preliminary comparative analysis of droop control, by means of dynamic computational models allowed to conclude that droop control can be considered a particular case of the techniques that emulate the synchronous machine.

A dynamic analysis of droop controlled converters and synchronverters was developed from a small-signal model, which demonstrated good accuracy when compared with the dynamic model. The small-signal models developed explicitly shows that the difference between the two control structures can be represented by a 2×2 matrix. The results showed that for a considerable range of gains, both control algorithms lead to very similar dynamic responses, however, the synchronverter allows an extended range of gains in the stable region, which gives this control algorithm more flexibility. On the other hand, the droop control is simpler and presents more damped dynamic response when compared to the synchronverter.

A positive-negative synchronverter in order to enhance fault-ride through in synchronverters connected to the grid was developed. This approach solved characteristic problems of three-phase converters emulating synchronous machines when submitted to unbalanced voltage conditions, that are: (i) very unbalanced output currents, which may cause shutdown of the converter and (ii) oscillations in the active power, which can damage the DC bus capacitors. The PN synchronverter is based on the Fortescue transformation and it was implemented by simulations and experimental setup, which demonstrated good performance in unbalanced grid and voltage sags.

From the previous study, the single-phase synchronverter applied to PV power systems was proposed aiming to adapt the original control part of the synchronverter for a single-phase power part. The result was an algorithm with minimal modifications and able to operate under voltage sags and harmonic distortion. In addition to the connected mode, the single-phase synchronverter is capable of operating in the island mode.

In Appendix A the main publications that resulted from this work are listed.

7.2 Future Works

The study of converters emulating the synchronous machine still has a number of opportunities to be investigated. The following are some proposals for continuity of the dissertation:

- to develop analysis of micro-grids with multiple synchronverter units, with converters operating as grid-forming, grid-feeding and grid-support;
- to develop analysis of micro-grids using different primary control techniques for each converter (droop control, synchronverter, VSG, etc);
- to compare the strategies applied to grid connected converters in unbalanced voltage conditions using simulations and small-signal models;
- to study the adaptations for the small-signal model evaluate the dynamic behavior of converters emulating synchronous machines when a main grid is not available;
- to develop a comparative study to clarify the dynamic differences between the PN synchronverter and the other techniques applied to converters emulating synchronous machines in unbalanced conditions presented in Zheng et al. [2018](#) and Chen et al. [2017](#);
- to evaluate the use of synchronverters connected in series with the grid, operating as active filter;
- to finish the commissioning of the experimental prototype to validate the control methods in grid-connected mode.

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A Written Papers

1. R. V. Ferreira, S. M. Silva, H. M. A. Antunes, G. Venkataramanan. "Dynamic Analysis of Grid-Connected Droop-Controlled Converters and Synchronverters" in Journal of Control, Automation and Electrical Systems, 2019.
2. D. I. Brandao, F. Mendes, R. V. Ferreira, S. M. Silva and I. A. Pires, "ACTIVE AND REACTIVE POWER INJECTION STRATEGIES FOR THREE-PHASE FOUR-WIRE INVERTERS DURING SYMMETRICAL/ASYMMETRICAL VOLTAGE SAGS," in IEEE Transactions on Industry Applications, 2019.
3. FERREIRA, R. V. et al. Assessment of energy storage viability for a PV power plant injecting during the peak load time. 2017, Florianópolis: 8th International Symposium on Power Electronics for Distributed Generation Systems (PEDG), 2017. p.1–8.
4. H. Antunes, S. Silva, B. Filho, R. Ferreira and D. I. Brandao, "A New Configuration for a Grid Forming Converter in AC Islanded Microgrid," PCIM Europe 2017; International Exhibition and Conference for Power Electronics, Intelligent Motion, Renewable Energy and Energy Management, Nuremberg, Germany, 2017, pp. 1-8.
5. H. M. A. Antunes, S. M. Silva, D. I. Brandao, R. V. Ferreira and B. d. J. C. Filho, "Analysis of a grid-forming converter based on repetitive control in centralized AC microgrid," 2017 IEEE 8th International Symposium on Power Electronics for Distributed Generation Systems (PEDG), Florianopolis, 2017, pp. 1-8.
6. FERREIRA, R. V. et al. Single-phase Synchronverter for Residential PV Power Systems. 2016, Belo Horizonte: ICHPQ, 2016. p.1–6.
7. H. M. A. Antunes, S. M. Silva, B. d. J. C. Filho, R. V. Ferreira and T. M. G. Costa, "Connection of a series hybrid filter in isolated microgrid for harmonic compensation," 2016 17th International Conference on Harmonics and Quality of Power (ICHQP), Belo Horizonte, 2016, pp. 24-29.
8. FERREIRA, R. V. et al. Estudo e Simulação de uma Microrrede de Energia Elétrica nos Modos Conectado e Isolado. 2015, São Paulo: AGRENER GD 2015 10o Congresso sobre Geração Distribuída e Energia no Meio Rural, 2015. p.1–10.
9. FERREIRA, R. V. e SILVA S. M. Técnicas de controle utilizadas em microrredes de energia elétrica: uma revisão. 2015, Campina Grande: XI CBQEE 2015 10ª Conferência Brasileira sobre Qualidade de Energia Elétrica.

B Experimental Grid-connected Converter

IN this chapter, it is shown the converter started-up during the final step of this doctoral work. The objective of the setup is to enable testing different algorithms applied to voltage-source converters, especially the converters emulating synchronous machines as discussed in previous chapter.

B.1 Experimental Setup Overview

Figure B.1 illustrates the experimental setup in block diagram. The system is constituted by an uncontrolled three-phase rectifier, an three-phase low-voltage inverter and a LCL filter.

The rectifier is responsible for providing the DC bus voltage. The three-phase inverter is equipped with a control system developed by the Tesla Power Engineering lab, containing a TMS320F28335 DSP from Texas Instruments, which can be programmed using Matlab / Simulink software, two voltage signal conditioning boards (one for voltage grid and another for converter output voltage) and one current signal conditioning board for the inductor current measurement. For the current measurement, three hall effect sensors are used. Figure B.2 illustrates the experimental setup.

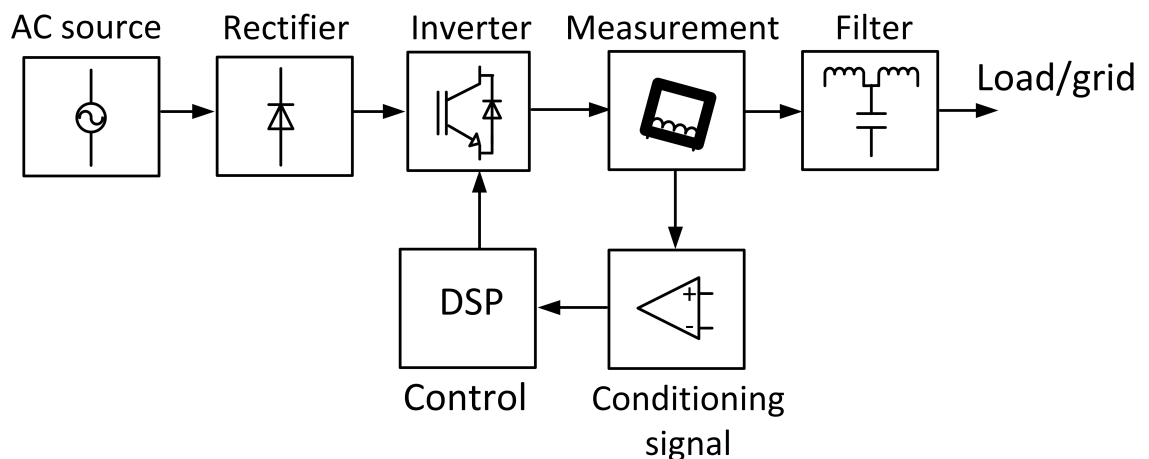


Figure B.1: Experimental setup in block diagram.

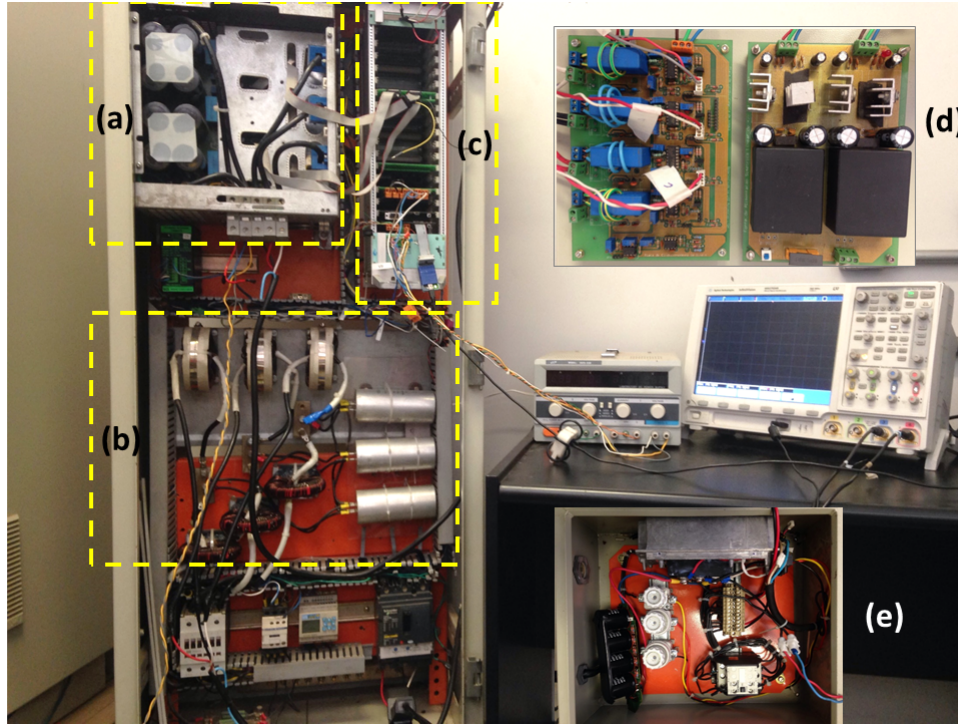


Figure B.2: Photo of the experimental setup: (a) power inverter, (b) LCL filter, (c) control system, (d) current signal conditioning board and (e) rectifier.

B.2 Converter characteristics

Figure B.3 presents the topology of the power inverter used in the experimental grid-connected converter. It is a three-phase full bridge IGBT inverter with anti-parallel diodes. The DC bus consists of two sets of capacitors with the central point accessible. A LCL filter is connected to the converter output to reduce current ripple and to attenuate switching components.

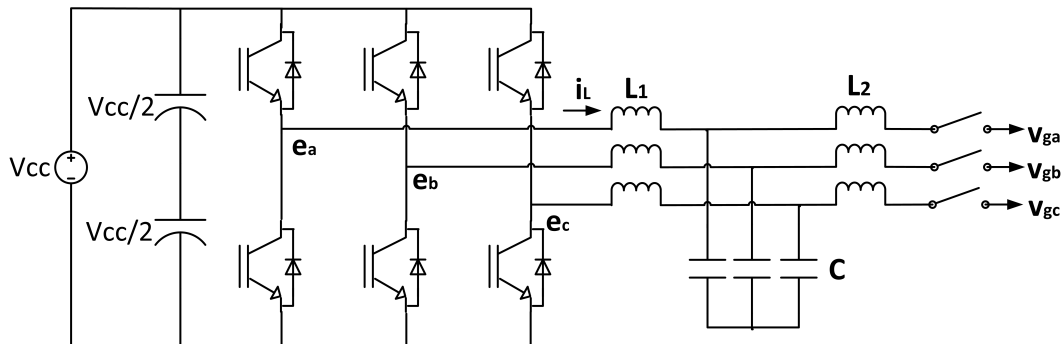


Figure B.3: Converter topology used in the experimental setup.

Next section presents the design of the LCL filter.

B.3 LCL Filter

The LCL filter, as its name suggests, is a third order filter constituted by two inductors and one capacitor. This configuration chosen for eliminating harmonic components generated by PWM inverters is capable of achieving higher harmonic attenuation (-60db/dec) when compared to the purely inductive first order filter (-20db/dec), as well as allowing for weight reduction and filter size (Silva et al. 2015).

In this section, it is briefly explained the sizing of each component based on the methodology proposed by Reznik (Reznik et al. 2012).

The first inductor is responsible for making the VSI inverter behave as a current source. For L_1 , the current ripple criterion is adopted, according equation B.1.

$$L_1 = \frac{V_{DC}}{(2 \cdot n \cdot f_{sw} \cdot \Delta i_{Lmax})} \quad (B.1)$$

where: V_{DC} is the DC bus voltage (V), f_{sw} is the switching frequency (Hz), i_{Lmax} is the maximum current ripple of the inductor (A) and n is the number of converter levels. Considering 10% current ripple and switching frequency equal to 10kHz:

$$L_1 = \frac{400}{(6.10k \cdot 40 \cdot \sqrt{2.0, 1})} \rightarrow L_1 = 294,63\mu H \quad (B.2)$$

The inductor L_1 used in the experimental setup has inductance equal to $320\mu H$, which meets the design criteria.

The capacitor is responsible for absorbing the high frequency components resulting from the converter switching. the capacitance per phase is calculated from the equation B.3.

The factor 0.05 is due to the consideration that with the capacitor connection there is a variation of the power factor of approximately 5%.

$$C = 0,05 \cdot \frac{S_n}{2 \cdot \pi \cdot f_g \cdot E_n^2} \quad (B.3)$$

where: S_n is the three-phase apparent power of the converter (kVA), f_g is the grid frequency (Hz), and E_n is the line voltage (V). The term 0.05 is due to the consideration that, with the capacitor connection, there is a variation of the power factor seen by the grid of approximately 5% (Liserre, Blaabjerg, and Hansen 2005).

Knowing that the apparent power of the converter is 15kVA, the line voltage is 220V and that, the grid frequency is 60Hz, the calculated capacitance is:

$$C = 0,05 \cdot \frac{.15k}{2 \cdot \pi \cdot 60 \cdot 220^2} \rightarrow C = 41,11\mu F \quad (B.4)$$

The capacitance of the filter installed in the experimental setup is $40\mu F$.

Equation B.5 is useful to calculate the inductance L_2 , due to the attenuation factor K_a , which is commonly set to 20%.

$$L_2 = \frac{\sqrt{\frac{1}{k_a^2} + 1}}{(2\pi f_g E_n)^2 \cdot C_f} \quad (B.5)$$

Applying the parameters of the experiment, the L_2 inductance is:

$$L_2 = \frac{\sqrt{\frac{1}{0,2^2} + 1}}{(2\pi 60 \cdot 220)^2 \cdot 40\mu} \rightarrow L_2 = 21,80\mu H \quad (B.6)$$

The inductor L_2 of the experimental setup has $56\mu H$, greater than the minimum designed, or in other words, the attenuation factor of L_2 used is smaller than the calculated.

B.3.1 Current Measurement

The sensor used for current measurement is a Hall effect sensor, capable of measuring direct and alternating currents, with galvanic oscillation between primary and secondary (LEM 2014). The primary nominal current is 200Arms, however, since the purpose of the experiment is to work with currents below 40A, the measurement range was reduced, applying 5 cable turns around the sensor core.

The signal conditioning is obtained by an electronic circuit that converts the Hall sensor output current in a -10 to -10V, compatible with the DSP analog inputs.

B.4 Experimental Results

The first test to validate the experimental setup consisted of controlling the current in the inductor L_1 . To this end, the three L_1 inductors are connected in short-circuit. It was used a proportional (P) controller designed as shown in the appendix B. The first result is illustrated in the Figure B.4 and as can be seen, the presented currents are distorted because the dead time of the converter. Including the dead time compensation, a new result is presented in Figure B.5 and reveals a behavior very close to sinusoidal.

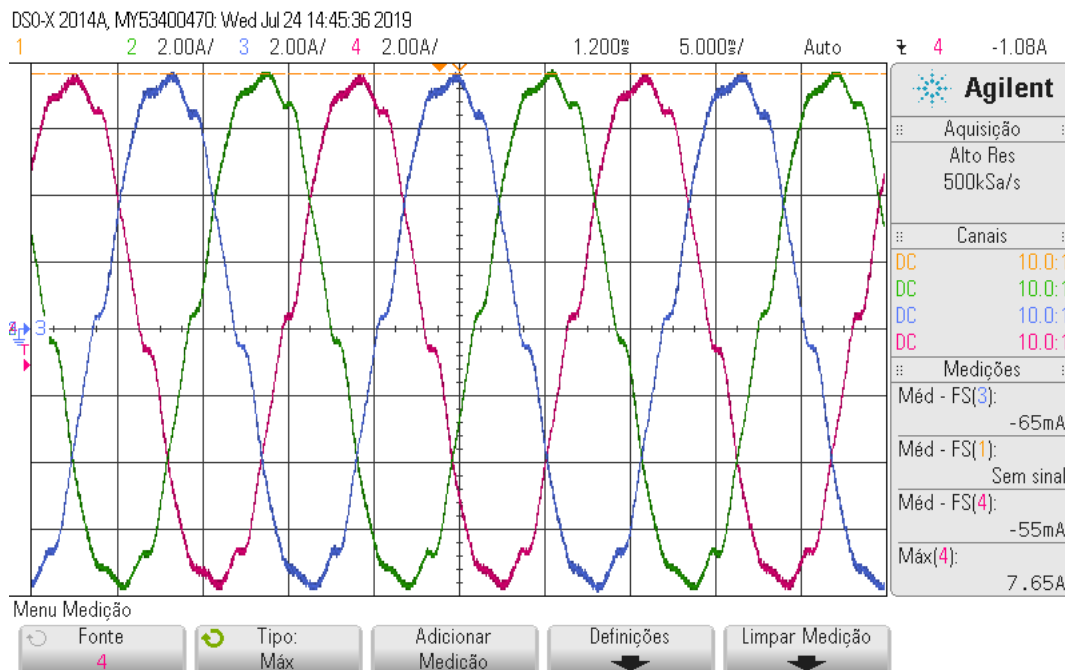


Figure B.4: Current control implementation using P controller.

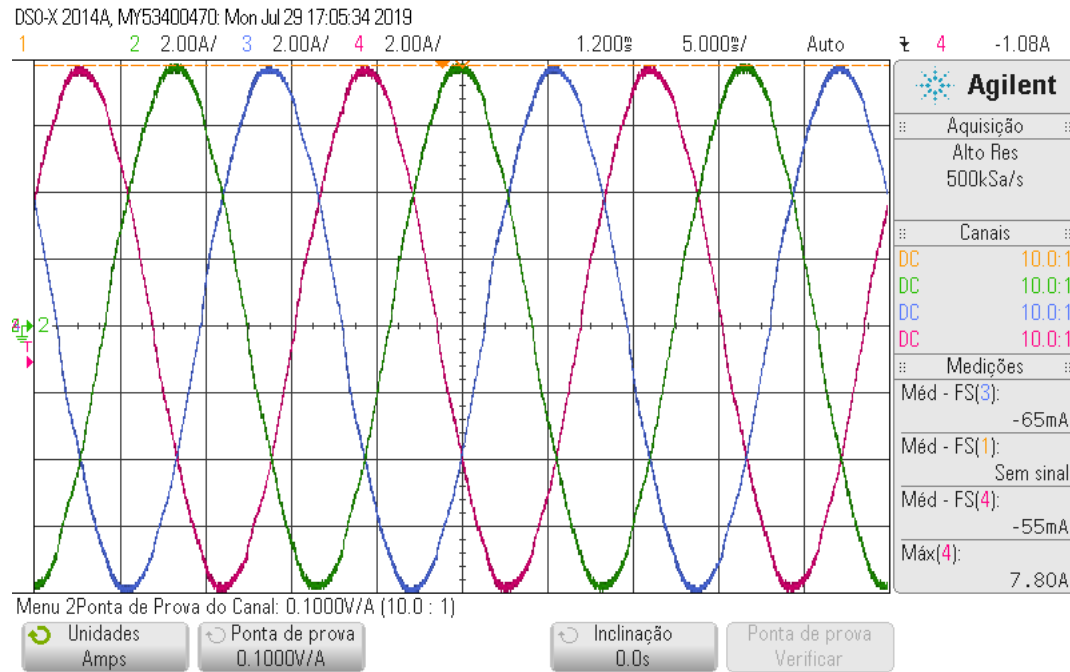


Figure B.5: Current control implementation adding dead-time correction.

From this experiment step and from the results presented for current control, the following analyzes can be performed:

- the dead time of the converter affects the current wave form and this effect is worse for lower controlled currents;
- the dead time compensation is effective to correct the current wave form;
- as expected, the currents present higher ripple for higher DC bus voltage (many DC bus voltages have been tried and not presented here to avoid repetitive illustrations);
- the designed gains generated satisfactory results in stead state;
- off-set problems in the measurement signals had to be computationally corrected through a moving average algorithm, avoiding associated errors and control instability.

After adjusting the current control, it is time to evaluate the operation of the voltage control. In this control level it was used a proportional-integral (PI) controller. Figure B.6 presents a voltage control response when the set-point of the voltage is 180Vp and the load is a resistive bank of 13Ω . The design of the voltage and current controllers are explained in the appendix B.

A nonlinear load test is also realized, having as load a rectifier with a 1.88mF output capacitor, feeding a resistance. Figure B.7 presents the voltage set-point, voltage output and the load current. One can see that after the introduction of the nonlinear load, a distortion in the voltage wave form appears during the charge time of the rectifier capacitor.

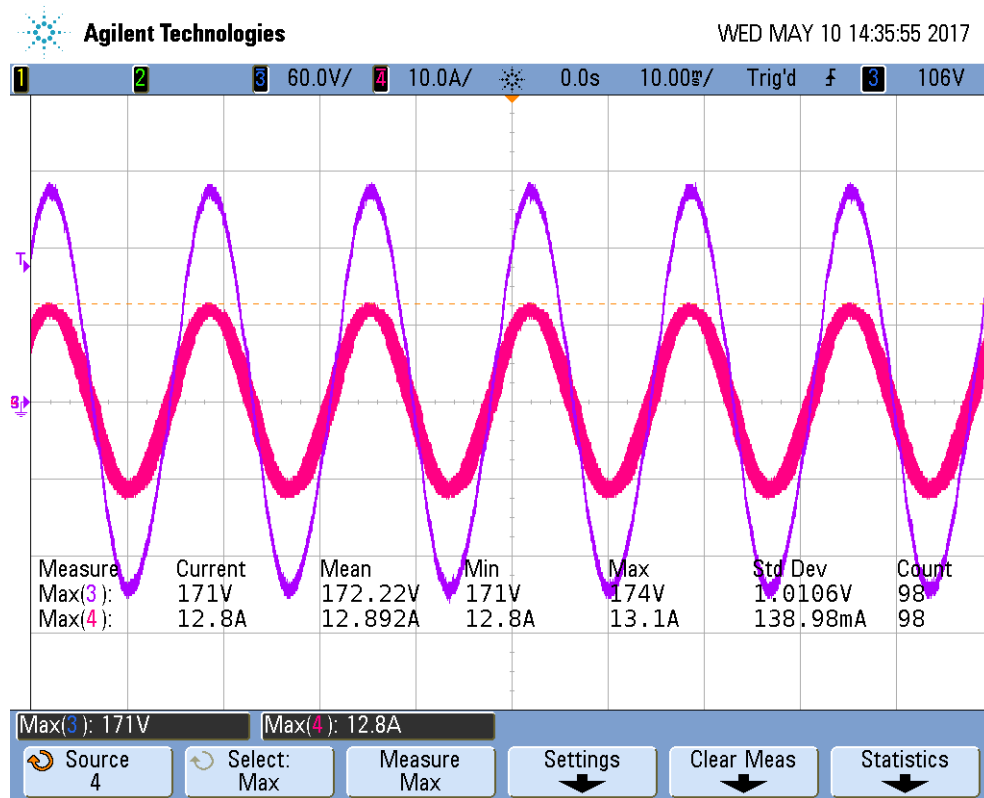


Figure B.6: Voltage control implementation with linear load: voltage x output current for a 180Vp set-point and DC bus voltage equal to 200V.

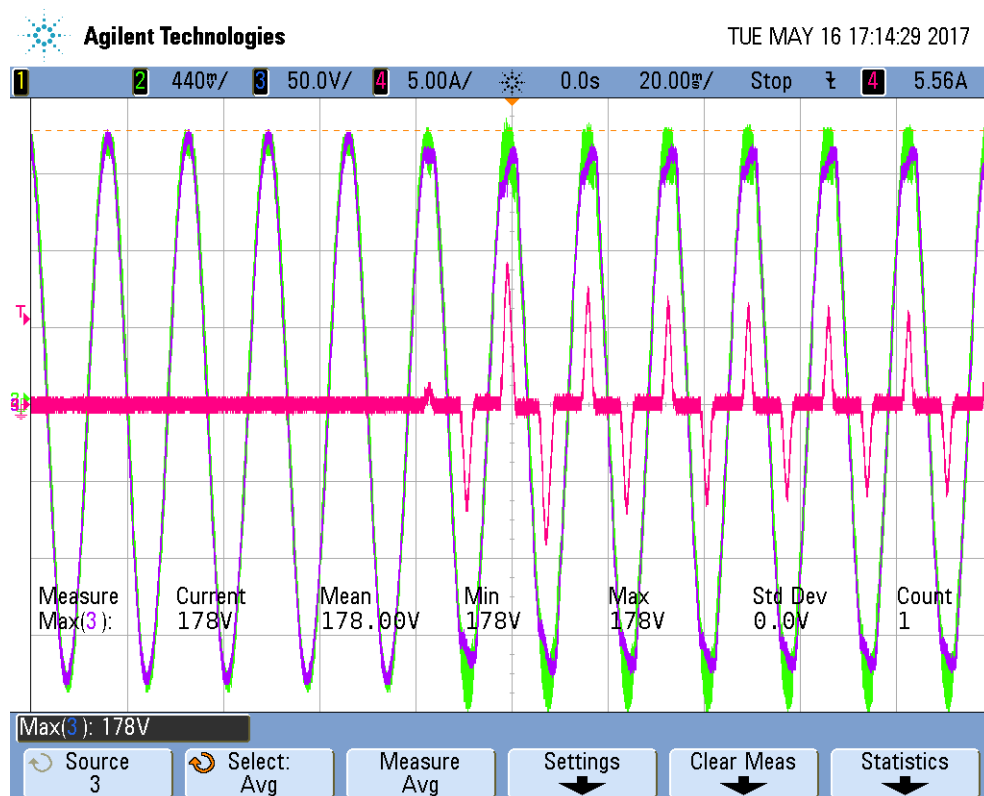


Figure B.7: Voltage control implementation with nonlinear load.

The results shown for the voltage control allow us to conclude that the converter operates as expected, showing deviations only under purely nonlinear load condition. These deviations do not mean problems to the connected operation of the inverter, since the grid has high inertia and contributes to the power quality.

After completing the voltage control testing step, the primary control level was implemented. Firstly, the droop control algorithm was evaluated in island mode. In this case, the active and reactive power references are set to zero and consequently the converter provides the voltage and frequency regulation. The control gains were set to $K_p=0.05\%$ and $K_q=0.01\%$, exactly the same of previous chapters. Figure B.8 shows the voltage and output current when the converter is feeding a 2kW resistive load, and, Figure B.9 illustrates the active power and the output current. The objective of this test is not evaluate transient behaviors, but to validate the converter operation with the droop control method in the primary level.



Figure B.8: Droop control implementation in island mode: voltage and output current.

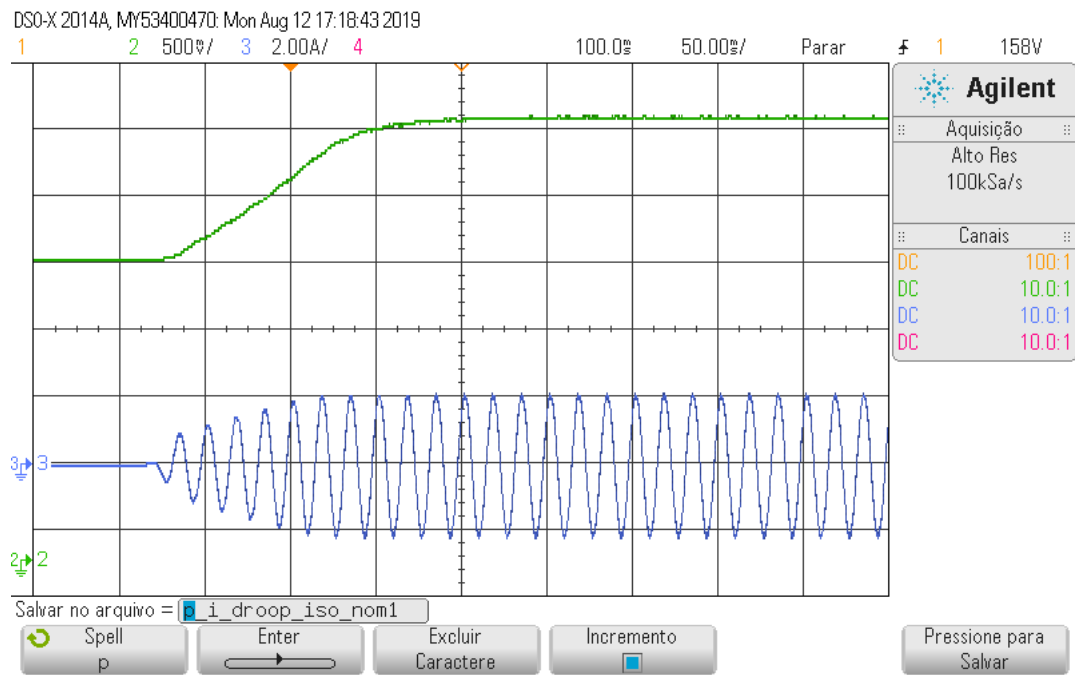


Figure B.9: Droop control implementation in island mode: active power and output current.

C Controllers Design

This appendix aims to present the design of the current and voltage controllers, which constitute the PWM control level, necessary for the operation of the primary control. The experimental results presented in chapter 4 were obtained from the controllers described herein.

C.1 Current Control Loop

The current controller design is a fundamental design step as it is the innermost loop in converter operation as voltage control. The proportional control gain is defined using the gain margin and phase margin criteria. The cutoff frequency of the closed loop transfer function should be approximately 10% of the converter switching frequency to eliminate the effect of switching harmonics (Yazdani and Iravani 2010).

To model the current control loop it is considered that the LCL filter capacitor impedance is very high at the fundamental frequency and therefore can be disregarded (Wang et al. 2012). The block diagram illustrated in Figure C.1 shows the current controlled on the L_1 inductor of the output filter. K_p is the proportional control gain, K_{PWM} is a gain attributed to the drive modulation, R and L are intrinsic parameters of the L_1 inductor. From this model the transfer function necessary to determine the controller gains is extracted.

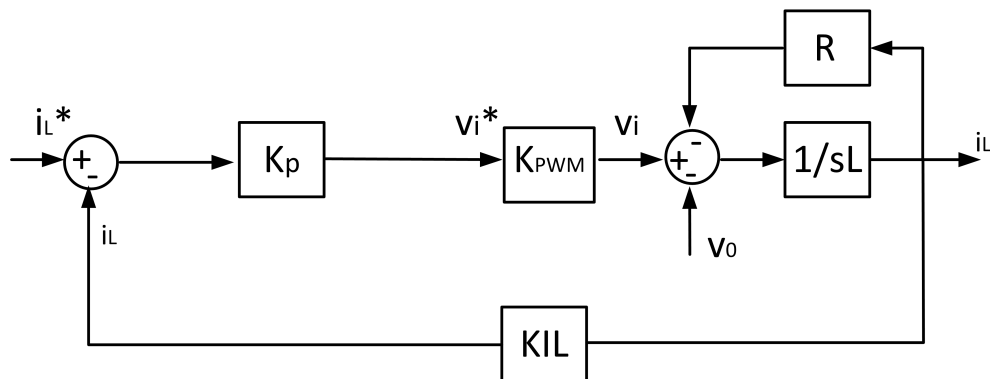


Figure C.1: Block diagram of the current control loop.

The electric parameters of the experimental converter are: DC bus voltage ($V_{cc}=200V$), filter parameters ($L=640\mu H$, $C=20\mu F$) and sensor gains ($KVCA=1/285$ for the voltage sensor and $KIL=1/25$ for the current sensor).

Figure C.2 presents the frequency response of the current control closed loop for phase margin equal to 85° and cut frequency equal to 1kHz. The calculation of the proportional gain resulted in a $K_p=0,5$. There is an error close to zero in magnitude and a significant phase error of approximately 3.4° , which will be corrected in the voltage loop.

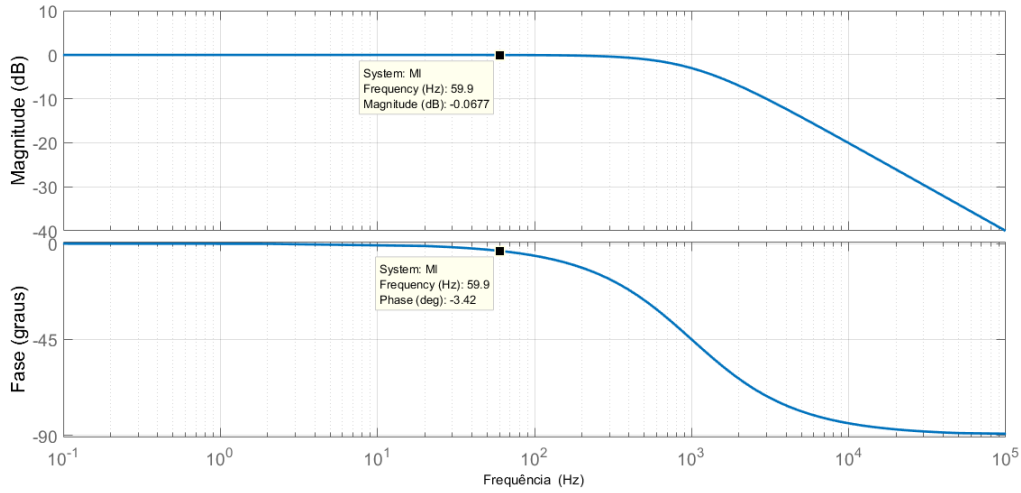


Figure C.2: Frequency response of closed loop current control.

C.2 Voltage Control Loop

Figure C.3 illustrates the voltage control loop in block diagram, from which the open loop transfer function can be obtained to allow projecting the K_{pv} e K_{iv} gains.

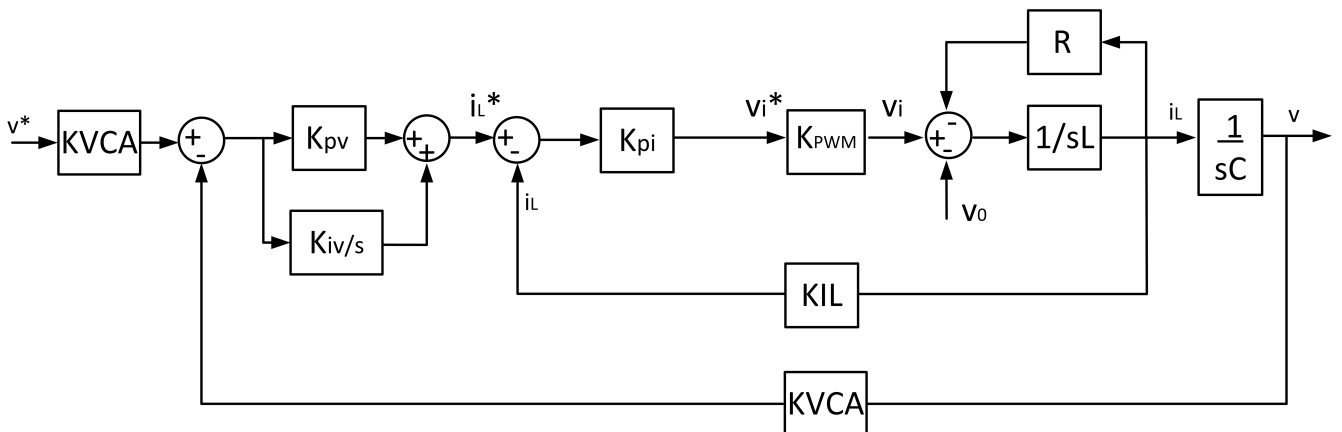


Figure C.3: Voltage control loop in block diagram.

For the design of the voltage controller, the cutoff frequency was set to 400Hz and the phase margin was set equal to 60° , resulting in $K_{pv}=0,6$ and $K_{iv}=195$. The frequency response of the PI control in the closed loop voltage can be seen in the figure C.4.

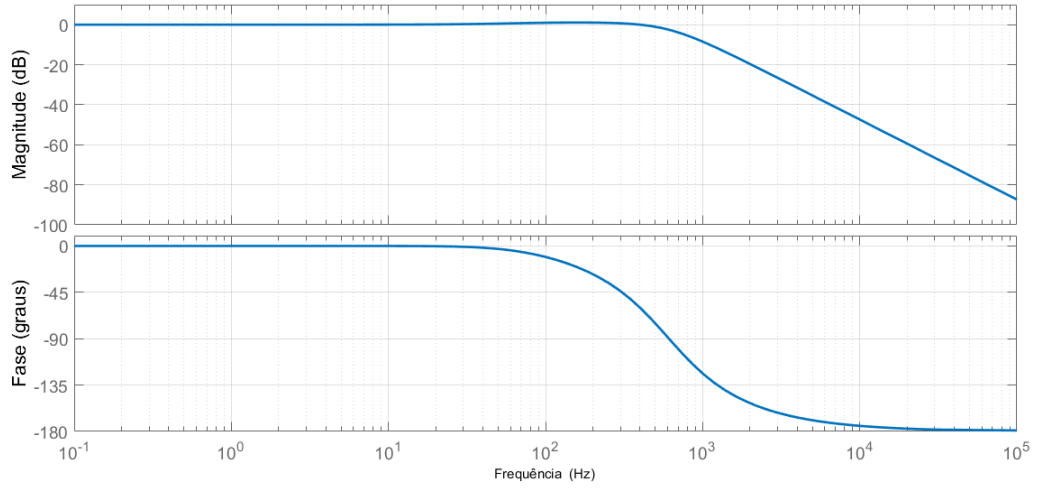


Figure C.4: Frequency response of the PI controller designed for the voltage loop.

C.2.1 Resonant Controller Design

As seen earlier, the pure PI controller is unable to produce zero error for sinusoidal voltage references, especially at its phase angle. This section presents the design of the resonant controller, which operates in parallel with the PI controller, tuned to a frequency of 60Hz. This technique allows to have resonant actions in as many frequencies as you want, but for each of them a new polynomial function is required and therefore, increasing the number of resonant terms, the computational cost is also increased.

The transfer function of the real resonant controller is expressed by the equation (C.1).

$$F.T._{res} = \frac{K_i \cdot 2\omega_b s}{s^2 + 2\omega_b s + \omega_0^2} \quad (C.1)$$

in which ω_0 is the desired resonant frequency and ω_b represents the filter bandwidth in rad/s.

For the design of the resonant controller, the considered frequencies were $\omega_0 = 2\pi \cdot 60$ rad/s and $\omega_b = 2\pi \cdot 2$ rad/s. For obtaining the polynomial coefficients in the "z" domain, the Matlab "c2d" function was used and the numerator and denominator coefficients were extracted. Figure C.5 illustrates the behavior of the PI controller combined with the resonant controller, showing the correction of the phase error.

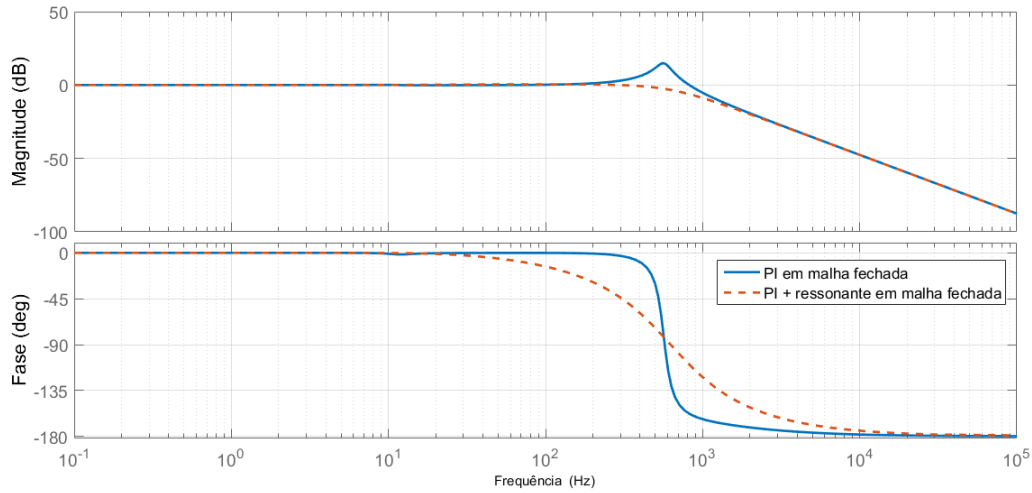


Figure C.5: Frequency response of the PI controller in parallel with the resonant controller.

C.2.2 Repetitive Controller Design

Another control possibility to correct the phase error is the repetitive control. Besides that, the repetitive control improves the performance of the voltage controller under load disturbances that repeat periodically. The repetitive control consists of use the error information on past samples to generate the control action. Figure C.6 shows one of its ways of implementation, in which T is the period of the fundamental frequency of the input signal and the minus sign indicates that the frequency response of the transfer function presents gain peaks for odd harmonics.

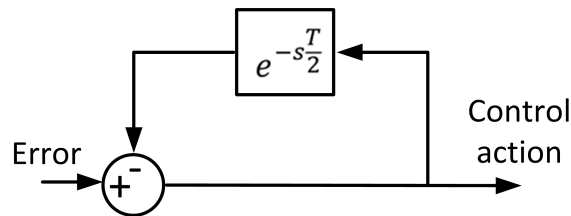


Figure C.6: Repetitive controller for odd harmonics in block diagram.

This type of controller has infinite poles at multiple frequencies of the fundamental component and is applicable to voltage control for loads with significant levels of harmonic distortion. As a limitation, the repetitive control has slow transient response, due to the delay implemented in the feedback (Silva 2012) and for this reason is commonly used in association with some type of fast dynamic response control. In the present work, the repetitive control is being used in parallel with the PI controller in the voltage loop, aiming to correct the PI limitations.

The repetitive controller design basically consists of calculating a K_r gain and designing the low-pass filter, which limits the voltage loop passage range. Higher K_r implies faster transient response, but too high values may lead to system instability. In the experimental tests of chapter 4 the K_r used was equal to 0.5.

The frequency response for pure repetitive action is shown in Figure C.7. As expected, the controller has resonances at all odd harmonic orders.

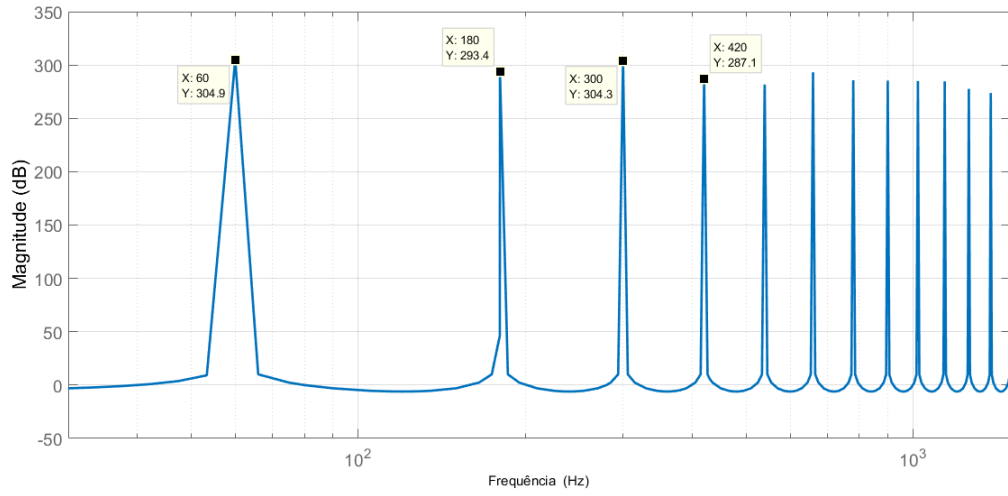


Figure C.7: Frequency response for the repetitive controller in open loop.

The low-pass filter is implemented in discrete time, but for ease of design, we start from the transfer function in the "s" domain, as represented in (C.2).

$$F(s) = \frac{e^{\tau s}}{\frac{s^2}{\omega_c^2} + \frac{2\zeta s}{\omega_c} + 1} \quad (C.2)$$

where, ω_c is the cutoff frequency of the filter, which must be less than the current loop cutoff frequency, and ζ is the damping factor.

For the present project were considered $\omega_c = 2\pi \cdot 1000 \text{ rad/s}$ and $\zeta = 0,7$, obtaining the frequency response of the association of repetitive controller and low-pass filter, as illustrated in Figure C.8. The full implementation of the voltage control is shown in Figure C.9 and its frequency response can be seen in Figure C.10.

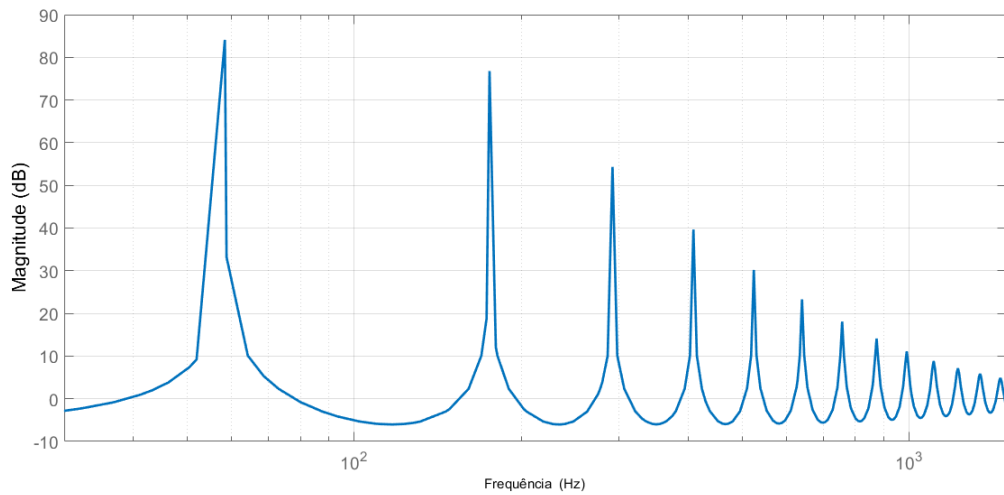
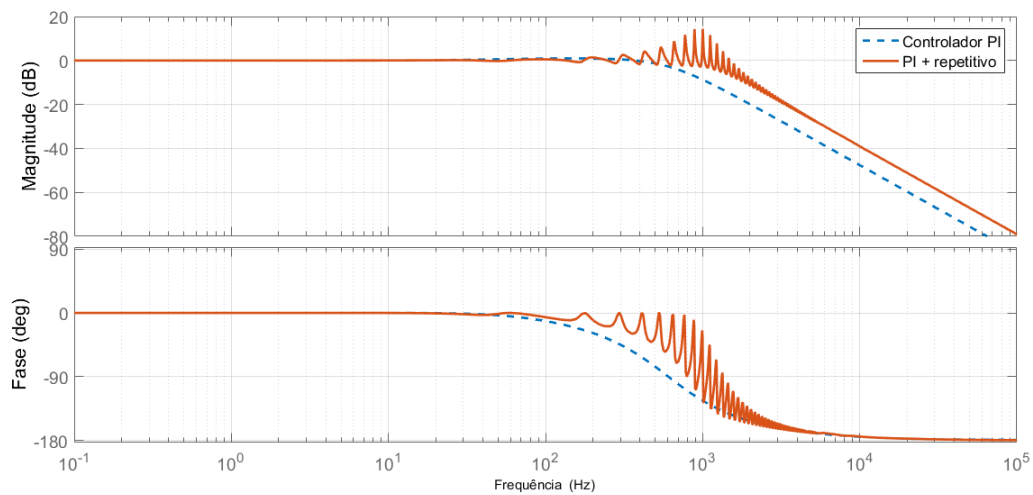
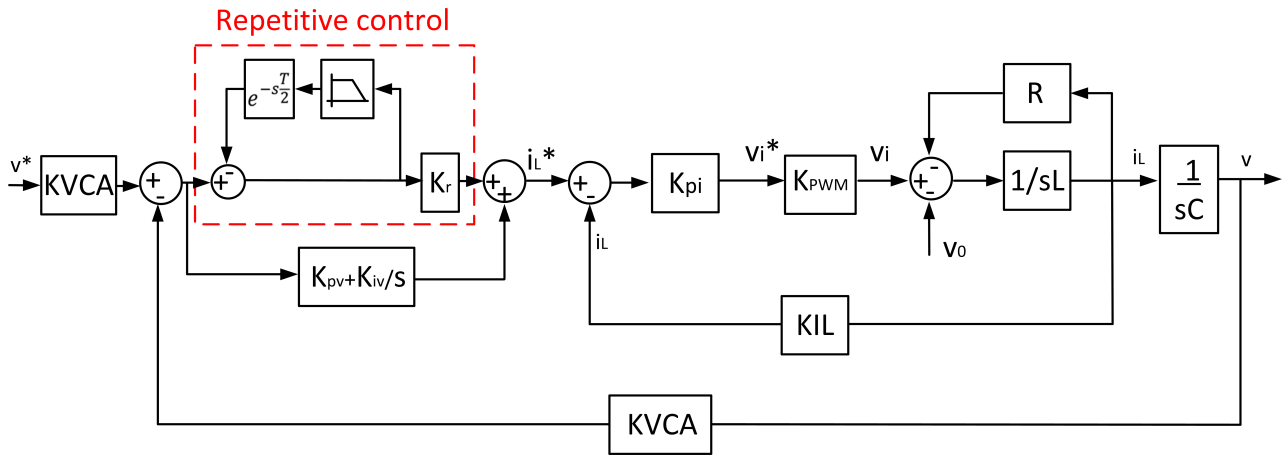


Figure C.8: Frequency response for repetitive controller with low pass filter.



D Hardware in the Loop - Typhoon Hil 600

This appendix presents a brief description of the hardware in the loop technology and the characteristics of the Typhoon HIL 600, hardware used to validate the algorithms of droop control, synchronverter and the proposed positive-negative synchronverter.

D.1 The Hardware in the Loop Technology

Hardware in the loop technology is believed to have emerged in the aerospace industry in the design and testing of complex flight systems. Since then, its use has been growing and expanding to other areas such as automotive and industrial. The main reasons for the growing interest of the hardware in the loop are related to the increasing complexity of systems and shorter project deadlines. The controller is "deceived" because its input signals come from a real-time simulation of a mathematical model instead of a physical system. This technology, therefore, makes it possible to eliminate the difficulties imposed by the need for a physical control plant.

Basically, three issues justify the acceptance of the hardware in the loop to validate control algorithms: cost, time and safety. An experimental hardware in the loop test can be performed as shown in Figure D.1. As can be seen, the setup is composed of computer, DPS, HIL and oscilloscope. The computer allows download the control algorithm to the DSP, download the power system configuration to the HIL and monitoring quantities measured from the power system.

In general, HIL technology manufacturers provide libraries containing power sources, measurements instruments, many converter topologies, passive elements among other library parts useful for emulating power electronics setups.

The interface of signals between HIL and DSP is performed by an interface board containing analog and digital inputs and outputs. If the experimental test includes a power converter for example, the PWM signals generated from the DSP are connected to the HIL digital inputs to switch the emulated converter. On the other hand, measurement signals, such as current and voltage, are sent to the DSP from the HIL analog outputs.

It is possible to monitor quantities using a SCADA software available from the HIL manufacturers or using a oscilloscope.

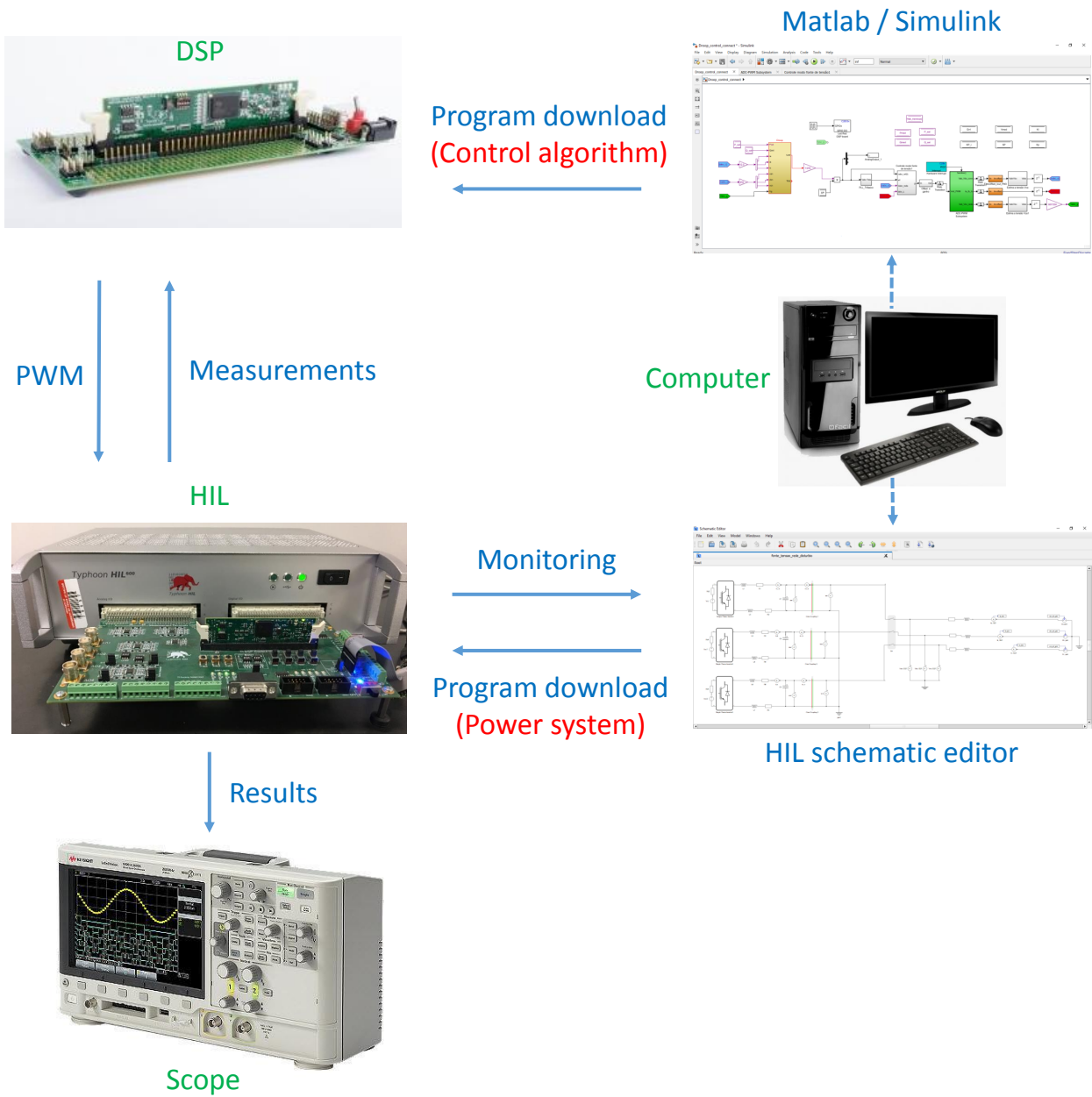


Figure D.1: Experimental setup for hardware in the loop technology.

Next section explain the hardware in the loop available by the Tesla Power Engineering laboratory and used in this dissertation.

D.2 Typhoon HIL 600

The Typhoon HIL 600 is a hardware in the loop technology made in Switzerland. The hardware is constituted by a set of processors, communication interfaces and high speed input-output (I/O) board. Figure D.2 illustrates the Typhoon HIL 600.



Figure D.2: Typhoon HIL 600 illustrative photo.

Some of the most important Typhoon Hil 600 features are listed below:

- Processor type: FPGA based;
- HIL processors: up to 8 cores;
- Communication protocols: Ethernet IP, Modbus, IEC 61850 and SunSpec;
- Simulation time step: Down to 500ns;
- Digital sampling resolution: 20ns;
- High speed I/O: 16 analog inputs, 32 analog outputs, 32 digital inputs and 32 digital outputs.

The Typhoon HIL technology contains its own software named Typhoon HIL Control Center. The basic configuration of the power system is developed into the Schematic Editor as can be seen in the Figure D.3 while the monitoring system is elaborated in the HIL Scada. At least three files need to be generated in the HIL 600 configuration:

- the ".tse" - where the power system is configured;
- the ".cus" - which corresponds to the HIL Scada Screens;
- the ".runx" - which contains the I/O configurations.

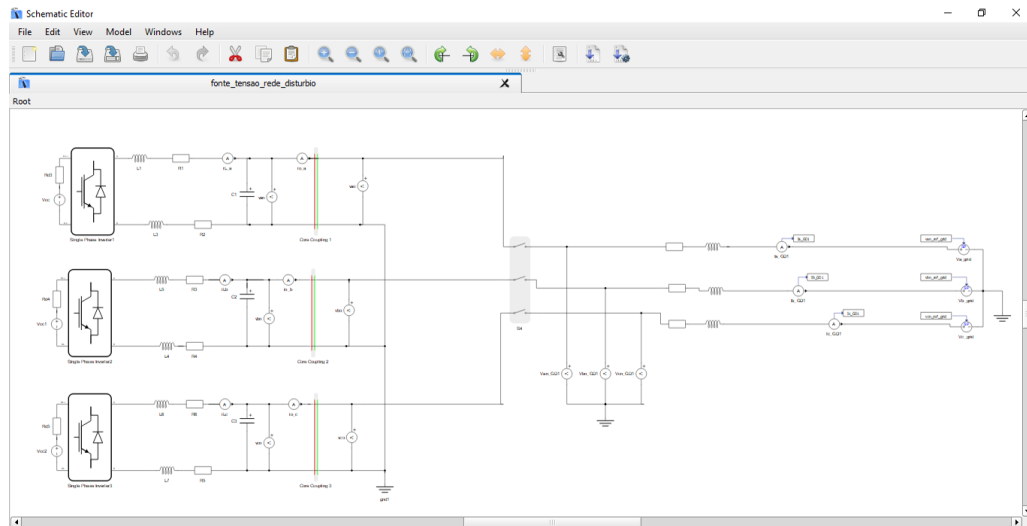


Figure D.3: Typhoon HIL schematic editor.