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Stability Analysis and Enhancement for the Power Conversion System

LIAO HUANYUE

School of Electrical and Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2021

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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Supervisor Declaration Statement

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

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Cai Wenjian

Authorship Attribution Statement

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Chapter 3 is published as **Huanyue. Liao** and Xin. Zhang, “Stabilization of Inverter–Based Distributed Generation System via Virtual Impedance Regulator,” in 2019 IEEE Energy Conversion Congress and Exposition (ECCE), USA, 2019, pp. 609–615.

The contributions of the co-authors are as follows:

- Prof Zhang provided the initial project direction and edited the manuscript drafts.
- I prepared the manuscript drafts. The manuscript was revised by Prof Zhang.
- I co-designed the study with Prof Zhang and performed all the laboratory work at the School of Electrical and Electronic Engineering. I also analyzed the data.
- When the paper was in revision, Prof. Zhang and Dr. Sun Changjiang gave me suggestions and helped to revise the response letter.

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Summary

The penetration level of renewable energy sources is increasing with the trend of power system innovation. As novel ac/dc loads and resources are integrated, modern distribution systems are becoming more complex, requiring more power conversion stages. Therefore, to reduce the number of power conversion stages and increase energy efficiency; hybrid ac/dc grids are emerging and attracting more attention. The interlinking ac/dc converter is one of the basic components of the conversion system between the dc bus and the ac bus, which is also indispensable in the hybrid system. The bus conversion system in the hybrid ac/dc grid usually comprises a three-phase interlinking converter and a dc side EMI filter. The bus conversion system can be treated as the interlinking ac/dc converter integrated with the EMI filter. The tightly regulated converters are commonly referred to as constant power loads (CPLs) on the dc bus side. The CPL exhibits negative resistive impedance characteristics, which is unfavorable for the stability of the conversion system.

According to the Middlebrook criterion, the ratio between the output impedance of the source subsystem (Z_{out}) and the input impedance of the load subsystem (Z_{in}) can be treated as the minor loop gain of the system, which can determine the system's stability. Therefore, the instability problems of the system can be solved by appropriately modifying the output impedance of the source subsystem or the input impedance of the load subsystem. However, the active damping methods of the output impedance are not applicable when the control of the source subsystem cannot be modified, such as when the upstream supply for the converter is an LC filter. For the active damping of the input impedance methods, it is necessary to a trade-off between the dc voltage stability and dynamic performance in the interface of the conversion system. Therefore, it is highly preferable to find a general method to moderate the degradation problems for the power conversion system.

First, the stability analysis for the power conversion system in the design and operation phase is described and summarized. A detailed review of techniques used to compensate for the destabilizing effect of CPLs in power systems is

presented along with their merits and associated challenges.

Then, the virtual impedance regulator (VIR) is one of the active damping methods to regulate the input impedance of the inverter without affecting the dynamic performance. In this paper, the VIR method is proposed to improve the stability of the power conversion system. The impedance-based method is established based on the equivalent input impedance of the three-phase inverter. Only a small range of frequency is modified to limit their influence on the system dynamic performance. The power conversion systems in grid-connected applications and stand-alone operation are both detailed discussed with the proposed VIR.

The dichotomy-based model predictive control (DS-MPC) method can directly compensate for the input impedance of the ac cascaded system. Then, a DS-MPC method with a fixed switching frequency is also proposed for the grid-connected inverter. The proposed fast dichotomy algorithm can select and deduce the optimal voltage vector dynamically through the space vector plane. Therefore, the proposed DS-MPC strategy could ensure dynamic performance and steady-state performance as well.

Simulation results are provided based on MATLAB/Simulink environment to validate all the performance and robustness of the VIR and FCS-MPC method. The experiments are also applied to the proposed methods' performance and robustness, which could emulate delays and errors.

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Abstract

This thesis investigates the instability problems of the ac/dc power conversion system due to the interactions between the EMI filter and the three-phase interlinking converter. A virtual impedance regulator (VIR) is proposed by this thesis to improve stability based on the detailed model of the power conversion system. This proposed virtual impedance can shape the input impedance of the source subsystem only in the specific frequency range. Therefore, the system can be stabilized while ensuring dynamic performance. The power conversion systems in grid-connected and stand-alone operation are both detailed discussed with the proposed VIR. The robust design of the virtual impedance regulator considering the deviation of the system model parameters is also proposed and verified. Then, the dichotomy-based model predictive control (DS-MPC) method with a fixed switching frequency is also proposed for the converter in the weak grid. The proposed fast dichotomy algorithm can select and deduce the optimal voltage vector dynamically through the space vector plane. Therefore, the proposed DS-MPC strategy could ensure dynamic performance and steady-state performance as well. Finally, experimental results are provided to validate the effectiveness of the proposed virtual impedance regulator.

Keywords: stability, three-phase converter, impedance modeling, model predictive control (MPC), voltage vector dichotomy.

List of Acronyms

REG	Renewable Power Generation
DC	Direct Current
AC	Alternating Current
DGS	Distributed Generation System
RE-DPGS	Renewable Energy Based Distributed Power Generation System
EV	Electric Vehicles
LED	Light-Emitting Diode
DG	Distributed Generation
PV	Photovoltaic
EMI	Electromagnetic Interference
CPL	Constant Power Load
DS-MPC	Dichotomy Solution-Based Model Predictive Control
VIR	Virtual Impedance Regulator
EKF	Extended Kalman Filter
POL	Point-Of-Lod
VSC	Voltage Source Converter
MPC	Model Predictive Control
GCI	Grid Connected Inverter
THD	Total Harmonic Distortion
FS-MPC	Finite Control Set Model Predictive Control
CCS-MPC	Continuous Control Set Model Predictive Control
GPC	Generalized Predictive Control
PCC	Predictive Current Control
PI	Proportional–Integral
d-axis	Direct-Axis
q-axis	Quadrature-Axis
RMS	Root Mean Square
PWM	Pulse Width Modulation
DO	Disturbance Observer

PCC	The Point of Common Coupling
KF	Kalman Filter
CORDIC	Coordinate Rotation Digital Compute
MSB	Most Significant Bit
MIMO	Multiple Inputs and Multiple-Outputs
PLL	Phase-Locked-Loop
FPGA	Field Programmable Gate Array

Chapter 1 Introduction

The general background of the power conversion system in the hybrid grids is described in this chapter. Following the trend of energy technology development, the current power systems are transforming into more electronics systems. However, the stability issues are brought with the new topologies and technologies, and thus solving them becomes the research objective of this thesis.

1.1 General Background

The ever-growing consumption of fossil energy sources, such as coal, oil, and natural gas, has brought SEVERE problems such as energy crisis and environmental pollution. The global electricity generation field is undergoing an unprecedented revolution process following this trend.

Table 1-1 Global energy share of the primary energy consumption

Modern Renewables 11.0%				Fossil Fuels	Nuclear Energy	Traditional Biomass
Biomass/ Solar/ Geothermal heat	Hydro -power	Wind/ Solar/ Biomass/ Geothermal\ Ocean Power	Biofuel	79.9%	2.2%	6.9%
4.2%	3.6%	2.0%	1.0%			

The proportion of various types of primary energy consumption in 2021 is shown in Table 1-1[1]. The fossil energy sources, which include the total primary energy consumption of oil, coal, and natural gas globally, are still the dominant energy sources. However, the reserves-to-production ratio of oil, coal, and natural gas is only 49.9, 132, and 49.8 years separately in 2019 [2]. The reserve-to-production ratio is the ratio of yearly proven energy reserves to the annual output, reflecting the sustainability of the energy supply. Fossil fuel is still the primary energy source, accounting for 79.9% of global primary energy consumption. In response to energy shortages and environmental problems, renewable energy is considered an alternative source. The global renewable share of total final energy consumption is shown in Table 1-1, which is the highest level in history. Renewable electricity generation accounts for 27.3% of

the global electricity production in 2019, and is set to increase more than 8% to 8300 TWh in 2021, despite the global electricity consumption being declined due to the influence of COVID-19 [3]. Among them, renewable power generation (REG), such as photovoltaic, wind power, and biomass energy, are considered as new energy resources and will play an essential and indispensable role in the power system.

The main approach to the large-scale development of renewable energy is the distributed generation system (DGS). Since the limitation of climate and weather, most renewable energy sources have randomness and intermittent characteristics. Therefore, in the renewable energy-based distributed power generation system (RE-DPGS), the storage units such as batteries, supercapacitors, and flywheels are required to smooth out the power pulsation and provide a reliable power supply to the loads.

Most renewable sources generate dc power or connect to the grid through the dc link, and the dc storage units and dc loads are also emerging, such as electric vehicles (EVs) and light-emitting diode (LED) lights. The dc microgrid is attracting growing attention because of the increasing demand and its strength of flexibility, efficiency, and less cost for fewer conversion stages [4]. The dc microgrids have been presented as a solution to integrate various distributed generators. However, the traditional power systems are mainly ac type. The conversion between the ac grid and dc microgrid is indispensable for integrating the system. The hybrid ac/dc grid attracts considerable interest for providing high-quality power to customers with optimal operational efficiency and reducing conversions. The stability of the hybrid grid is essential for stable operation for both ac and dc grids.

1.1.1 The Power Conversion System in Hybrid AC/DC Grids

As novel ac/dc loads and resources are integrated, modern distribution systems become more complex and require more power conversion stages [5]. Thus, to reduce the number of power conversion stages and increase energy efficiency, hybrid ac/dc grids are attracting more attention [6]. The hybrid ac/dc grids are recently emerged with the increasing applications of the smart grids due to their high reliability, design flexibility, and easy maintenance [4], [7]. The

hybrid ac/dc grids effectively optimize the connection of the large-scale distributed generation (DG) and grids. The hybrid ac/dc grid effectively combines the dc and ac grids' advantages and is widely used in more extensive applications [8].

Figure 1.1 The schematic diagram of the hybrid ac/dc power system

In Figure 1.1, the schematic diagram of the ac/dc hybrid grid with various applications is shown below. The typical hybrid ac/dc grid topology includes dc and ac sub-grids and the bus power conversion system. In the dc sub-grid, various REG units and storage units are connected to the dc bus through the converters, such as photovoltaic (PV) panels with dc/dc boosters and wind turbines with converters. REG units such as PV panels, wind turbines provide electricity to the loads. Energy storage units such as batteries and super capacitors are smoothing the intermittent characteristics of REG. The ac grid also relies on power electronics to realize its functions and achieve the required performance similar to dc grid. The new architectures of the DGS employ multi-

voltage level hybrid ac and dc systems. The optimum choice of the system voltage levels and forms includes complex trade-offs between cost, performance, safety, and feasibility [6]. These trade-offs and different architectures have been investigated in different applications. The reliable and stable operation of the conversion system is always a guarantee for the stable operation of the hybrid ac/dc grids.

In the hybrid ac/dc grids, the interlinking ac/dc converter is the key component to interface the dc bus with the ac bus. Operating in the control mode, the interlinking converters realize the active power transmission and adjust between the ac and dc grids. However, these high switching frequency converters are also known to bring electromagnetic interference (EMI) noise to the dc bus [9]. Hence, to address the EMI issues, an EMI filter [10] (usually an LC filter) can be installed at the dc link to suppress the noise. The interlinking ac/dc converter integrated with the EMI filter can be treated as the bus conversion system, as shown in Figure 1.2.

Figure 1.2 Schematic of the studied power conversion system

The EMI filter and the interlinking converter can be deemed as a typical cascaded system where the power-flow front-end converter (i.e., the EMI filter) is defined as the source subsystem and the power-flow back-end converter (i.e., interlinking converter) is defined as the load subsystem [11], as shown in Figure 1.2. When the converter is tightly regulated, it behaves as a constant power load (CPL) and exhibits a negative incremental input impedance characteristic [12]. The converter's negative input impedance characteristic degrades the cascaded system's stability. It may even cause issues like oscillation [13]. The exact reasons for the degradation problems are discussed below.

1.1.2 The Degradation Reason of the Conversion System

The switching frequencies of modern power electronics have increased from several hundred kHz to several hundred MHz with the technology development. The size, delay time, harmonic components of the electromagnetic devices have significantly reduced due to the increased frequency. However, each switching transient of the electronic devices will interact with the inductors and capacitors in the circuit and then generate the pulse current. Due to the high-speed switching process of power electronic equipment, severe EMI noise will be introduced to the overall system and interrupts or even degrades the performance of the electrical devices [14], [15].

EMI noises in power electronics are mainly conducted interference which propagated through cables in the system. The practical applications of the electronic devices must meet the EMI standards to ensure the acceptance level of the interference. Generally, even a well-designed device with proper shielding and grounding measures does not provide complete EMI protection. There are still conducted disturbances due to the cable on the equipment and the system for receiving and transmitting the disturbance. A suitable EMI filter design is an important means to suppress conducted interference to meet the relevant standards. The EMI filter could attenuate high-frequency noises and interference signals. A general-purpose EMI power filter can be defined as a low-pass network, consisting of passive components such as inductors, capacitors, or resistors. Compared to the electrolytic capacitor, the LC filter, consisting of an inductor and a small capacitance film capacitor, is more widely used because it requires much less space and has a longer lifetime [16].

Figure 1.3 The CPL cascaded with LC input filter.

In this section, the degradation reason for CPL cascaded with LC EMI filter is discussed. The closed-loop controlled converter behaves as a CPL and tends to degrade the stability of the upstream dc bus. The equivalent circuit with LC EMI filter is shown in Figure 1.4. The LC input filter consists of L and C .

Then, according to Figure 1.3, the transfer function (1-1) can be deduced based on the equivalent circuit according to the well-known second-order function. The input is the dc source voltage v_s and the output is the voltage of the converter v_{dc} .

$$T(s) = \frac{1}{s^2 LC + s \frac{L}{R_{inv}} + 1} \quad (1-1)$$

Here the resonant frequency ω_0 and damping factor ζ is introduced to be a more familiar format

$$T(s) = \frac{1}{\frac{s^2}{\omega_0^2} + 2\zeta \frac{s}{\omega_0} + 1} \quad (1-2)$$

where

$$\omega = \frac{1}{\sqrt{LC}} \quad (1-3)$$

$$\zeta = \frac{L\omega_0}{2R_{inv}} \quad (1-4)$$

Considering there is an excitation of unit-step function at the input side, the Laplace equation can be deduced as

$$Y(s) = \frac{1}{\frac{s^2}{\omega_0^2} + 2\zeta \frac{s}{\omega_0} + 1} \frac{1}{s} \quad (1-5)$$

To demonstrate the time domain response, the inverse Laplace transform can be given as (1-5)

$$Y(t) = 1 - \frac{e^{-\zeta\omega_0 t}}{\sqrt{1-\zeta^2}} \sin(\omega_d t + \theta) \quad (1-6)$$

for $\zeta < 1$, where

$$\omega_d = \omega_0 \sqrt{1-\zeta^2} \quad (1-7)$$

$$\theta = \cos^{-1}(\zeta) \quad (1-8)$$

The denominator roots of (1-2) noted as ρ_1 and ρ_2 , are the poles affecting the system stability. The value of the damping factor ζ can deduce the two poles and the stability of the system, according to (1-5).

i) $\zeta > 0$: There will be a real portion in the two poles ρ_1 and ρ_2 , which also means ohmic losses in the system. The system output will be a decaying oscillating response depending on the different values of the damping factor ζ . There are three different cases of ζ : $\zeta > 1$ (overdamping), $\zeta = 1$ (critical damping) or $0 < \zeta < 1$ (under damping).

ii) $\zeta = 0$: Both of the two poles ρ_1 and ρ_2 are pure imaginary poles, which $\rho_{1,2} = \pm j\omega_0$. Therefore, the system output is permanently oscillating without decay.

iii) $\zeta < 0$: The two poles ρ_1 and ρ_2 are including the positive real portion. Then, the transient response of the system diverges at any level of excitation.

As well known, the closed-loop controlled converter behaves as negative impedance, which means $R_{inv} < 0$. According to (5), $\zeta < 0$. So, it is case iii discussed above, i.e., the transient response of the whole system will diverge. Therefore, the CPL cascaded with an LC input filter will be unstable, where the essential reason is the CPL's input impedance behaves as a negative resistor.

As a result, this thesis targets the system's instability problem due to the interaction of the EMI LC input filter and the three-phase converter. It is highly desirable to find more general techniques for the converters to ensure system stability and converter's good performance simultaneously.

1.2 Motivation

The bus conversion system in the hybrid ac/dc grid usually comprises a three-phase interlinking converter and a dc side EMI filter. Based on the small-signal model and the equivalent circuits of the converter, this thesis real that the instability reason of the dc voltage is the interaction between the converter and the EMI filter. The main reason is that the converter's negative impedance characteristic degrades the cascaded system's stability. It is still a big challenge to propose a practical active method that ensures the stability of the power conversion system:

- 1) Firstly, the detailed model of the ac/dc power conversion system is not considered in many proposed stabilization methods. The impedance

model is simplified as an ideal negative impedance; the high-order terms in the model are neglected in the design and analysis. As a result, the dynamics of the system will be further influenced.

- 2) All the existing stabilization methods are only focused on enhancing the stability but ignore their impact on the original performance of the cascaded system. The existing stabilizing methods can degrade the original performance of the cascaded system. The coefficient of the compensator requires a trade-off between the dc voltage stability and the converter's dynamic performance. The optimal value of the parameters in the compensator is not proposed with the theoretical analysis.
- 3) The most active damping methods at the dc side are mainly applied in motor control, which cannot be used in the bus conversion system and cannot stabilize the system. The stability criteria and the impedance requirement will be entirely different. Therefore, it is better to develop stabilization methods for the ac/dc power conversion system.
- 4) A large impedance will occur on the grid side for the weak grid scenario, which will be overlapped in the output filter parameters. However, the detailed information of the plant model influences the performance of model predictive control. Meanwhile, considering the grid voltage could be unbalanced and distorted, the amplitude and frequency of the grid voltage are also varied. Therefore, the control performance could be influenced by the disturbance of equivalent grid impedance. Furthermore, the model used for control could be mismatched significantly in the weak grid due to MPC being deduced through the detailed control plant model.

To summarize, despite the abovementioned studies, few studies have been reported in the literature to develop a practical active stabilization method for the interlinking three-phase converters that ensures the stability of the system and minimizes the influence on the dynamic performance of the converter. Therefore, the abovementioned issues need further insights to enhance the performance of the power conversion system.

1.3 Objectives

In this thesis, the research objectives are providing methods for enhancing

the stability of the ac/dc power conversion system and minimizing the influence of the original performance of the system. The abovementioned issues need further insights to enhance the stability and performance of the power conversion system.

The specific objectives of this thesis are summarized as follows:

- 1) The bus conversion system in the hybrid ac/dc grid usually comprises a three-phase interlinking converter and a dc side EMI filter. The detailed model and the stability analysis of the ac/dc power conversion system with the grid-following control of the interlinking converter need further discussed as the fundamental of the stabilization methods.
- 2) The three-phase interlinking converters integrated with the input LC filter are essential between the ac and the dc buses. The LC filter and the interlinking converter will bring instability problems to the dc interface voltage. The detailed model and the stability analysis need further discussion as the fundamental of the stabilization methods.
- 3) The stabilization methods for the bus conversion system are proposed which can improve the stability while minimizing its influence on the converter's dynamic performance. All parameters in the proposed stabilization methods are selected by theoretical analysis. Moreover, the robust design considering the deviation of the system parameters is analytically implemented and the influence of the digital delay is also considered.
- 4) For the weak grid scenarios, a robust model predictive control (DS-MPC) stabilization method is proposed to reduce the total harmonic distortion and ensure dynamic and steady-state performance, even if the grid voltages are greatly distorted.

The proposed control techniques will improve the ac/dc power conversion system's stability, which may help extend their application fields further.

1.4 Major Contribution of the Thesis

This thesis aims to develop efficient stabilization methods for the ac/dc power conversion system. The main contributions can be summarized as follows:

- 1) The virtual impedance regulator (VIR) is proposed based on the detailed model of the bus conversion system with the grid-following control of the interlinking converter. First, this thesis presents and analyzes the comprehensive small-signal impedance model and the equivalent circuits of the interlinking converter. Then, the design of the proposed VIR is based on the detailed model. The proposed method can improve the stability while minimizing its influence on the converter's dynamic performance because only the phase of the input impedance is modified in the specific interaction frequency range. All parameters for the VIR are selected by theoretical analysis. Moreover, the robust design considering the deviation of the system parameters is analytically implemented and the influence of the digital delay is also considered. Finally, the influence of the virtual impedance regulator applied on the d or q axis is discussed and analyzed.
- 2) The VIR is deduced according to the detailed impedance model of the point-of-load (POL) converter with the stand-alone operation. The design of the VIR is based on modifying the phase of the input impedance only in the oscillation frequency range. Hence, the proposed VIR can minimize the impact from the regulator on the closed-loop performances of the POL converter while stabilizing the dc voltage. The derivation and verification of the VIR are all carefully analyzed and discussed. The proposed VIR can be implemented either on the d-axis or q-axis, and its influence is discussed and analyzed.
- 3) According to the model predictive control (MPC), a robust dichotomy method based on the disturbance observer of the grid voltage through the Extended Kalman filter (EKF) for the three-phase inverters. The fast dichotomy solution approach is employed in the proposed DS-MPC to dynamically choose and compute the finest voltage vector for every sampling period. Using the FPGA controller, the high-frequency control can be executed in a short time. Besides, the proposed DS-MPC strategy contains the basic finite control set model predictive control (FS-MPC) and the PWM modulator together with constant frequency essentially. Therefore, compared with the grid-connected inverter (GCI) based on the conventional FS-MPC strategy, it can be simplified for the design of the output filter. With the proposed DS-MPC, the total harmonic distortion

(THD) can also be reduced. Moreover, based on the disturbance observer, the model accuracy used in MPC is compensated. Hence, the proposed method for control of the inverter keeps the dynamic characteristics of FS-MPC and presents excellent steady-state performance and high robustness to disturbances of grid voltages at the same time.

1.5 Organization of the Thesis

Chapter 1 has given a background on the stability issues of the power conversion system in the hybrid ac/dc grids. The degradation problem of the power conversion system is introduced, which is also the motivation of this paper. Then a detailed survey of different compensation strategies to improve the stability of the power conversion system is presented. The chapter has motivated this work and the research contributions of this study.

Chapter 2 has given a literature review on the existing stability methods of the power conversion system in the hybrid ac/dc grids and their applications.

Chapter 3 presents the parallel-connected VIR control strategies for the power conversion system in grid-connected operation. Firstly, the simulation of the cascaded system in grid-connected operation cascaded with EMI LC filter is given to demonstrate the degradation problem. Then, the detailed modeling of the three-phase interlinking converter is presented. The VIR strategy to enhance stability without influencing the dynamic system is proposed. Both modulation strategies are tested and validated in experiments.

Chapter 4 presents a series-connected VIR based on the detailed small-signal model of the POL converter that feeds a linear three-phase load. The proposed method can improve the stability by modifying only the phase of the input impedance of the POL converter within the interaction frequency range. Therefore, the impact of the regulator on the converter's dynamic performance is minimized. The experiments show that the proposed stabilization methods exhibit good performance under different linear load scenarios.

Chapter 5 presents the robust dichotomy method according to the DS-MPC based on the disturbance observer of the grid voltage through the EKF for the

three-phase inverters. The fast dichotomy solution approach is employed in the proposed DS-MPC to dynamically choose and compute the most exemplary voltage vector for every sampling period. Using the FPGA controller, the high-frequency control can be executed in a short time. Besides, the proposed DS-MPC strategy essentially contains the basic FS-MPC and the PWM modulator with constant frequency.

Chapter 6 presents an overview of the research work up to now and the obtained results. The future work to develop in the Ph.D. studies is discussed as well.

Chapter 2 Literature Survey

This chapter summarized the existing stabilization methods for the dc voltage in the interface of the ac/dc power conversion system. A detailed review of techniques applied to enhance the stability of the interlinking converter is presented.

2.1 Literature Review

As discussed in Chapter 1, a stabilization method is necessary for the ac/dc power conversion system to improve the reliability and power quality. The existing methods based on hardware modifications and specialized control approaches are summarized there.

2.1.1 Passive Damping Methods

The negative impedance characteristics of the interlinking converter need to be compensated for increasing the stability of the system.

Figure 2.1 Schematic diagram for the passive damping methods

In [17], a passive damping method for the source subsystem is proposed as shown in Figure 2.1. The passive components are used to compensate for the negative impedance characteristic of the converter in the entire frequency range. However, the additional passive components lead to larger size requirements, extra cost, and power consumption. The typical applications for the passive methods are shown in Figure 2.2. In [18], the optimal damping filter design is proposed for the high switching frequency converters (1MHz or more) considering the parasitic parameters for the converters and the filters.

Figure 2.2 Typical passive dampers: (a) RC parallel damper; (b) RL parallel damper; (c) RL series damper.

2.1.2 Active Damping Methods

The active damping methods for the subsystem are proposed, which utilizes advanced control methods to modify the converter instead of the passive components.

The stability of the cascaded system can be analyzed using the Middlebrook criterion [19]. According to the Middlebrook criterion, the ratio between the output impedance of the source subsystem (Z_{out}) and the input impedance of the load subsystem (Z_{in}) can be treated as the minor loop gain of the system, which can determine the system's stability. Therefore, the instability problems of the system can be solved by appropriately modifying the output impedance of the source subsystem or the input impedance of the load subsystem. The corresponding virtual impedances are summarized and shown in Figure 2.3.

Figure 2.3 Virtual impedances inserted from the source subsystem and load subsystem.

1) Source-Side Active Damping Methods

The active damping methods for the source subsystem are proposed, which utilizes advanced control methods to modify the source converter instead of the passive components. An active damping method is proposed in [20] through the droop control of the load converter to emulate a virtual resistor in series with the

inductor. A nonlinear control loop stabilization method based on the output voltage feedback signal is presented in [21], which can be presented by an equivalent positive resistor. In [22], three active compensators are proposed to modify the output impedance of the source converter by injecting the dc voltage feedback signal to different nodes in the control loop of the converter. In [23], a virtual impedance connected in parallel with the source side rectifier is proposed for the source converter. In [24]–[28], the active damping methods are designed using feedback controllers to emulate the virtual resistance, which is connected in series with the inductor of the input filter. However, these methods are only applicable when the control of the source subsystem can be modified. In this thesis, the converter is fed by the input filter to the regulated dc bus. Therefore, the active damping methods for the source subsystem are not applicable to this scenario.

2) Load-Side Active Damping Methods

Some researchers have turned their attention to modifying the load subsystem via active stabilization methods. In [29], the stabilization methods are realized by decomposing and reconstructing the voltage reference of the SVPWM to modify the negative impedance characteristics of the converter with the motor into the positive impedance. A virtual resistor is proposed in [30], [31] to be connected in parallel with the line inductor by directly changing the motor voltage reference. In [32], a direct damping current approach is presented to stabilize the dc voltage of the drive system. In [33], an inductor current feedback-based active damping method is proposed to suppress the instability issue in the drive system. In [34], a feedforward compensation scheme is proposed based on decomposing the estimated rectified component of the distorted voltage; the stability analysis is not proposed in this paper. The above-mentioned methods directly modify the voltage reference of the SVPWM; thus the converter might enter over-modulation, and the dynamics are inevitably influenced. Besides, the impedance model is simplified as an ideal negative impedance; the high-order terms in the model are neglected in the design and analysis. As a result, the dynamics of the system will be further influenced.

There are also many active damping methods by introducing stabilization

feedbacks to the converter's linear control structure. The negative input-resistance compensator is proposed in [35] based on the dc voltage feedback into the current control loop realized by the high-pass filter. The negative impedance characteristics of the converter have been changed in an extensive frequency range, and the dynamic performance is thereby degraded. In [36], the feedback of the dc-voltage is injected into the q-axis current reference to increase the stability. In [37], the damping control for the converter with the PMSM drive is designed by injecting the inductor current feedback to the q-axis current reference. The stabilizers in [36] and [37] are both realized by the band-pass filter, which has a relatively slighter impact on the control loops. However, these methods can improve the stability, but the coefficient of the compensator requires a trade-off between the dc voltage stability and the converter's dynamic performance. The optimal value of the parameters in the compensator is not proposed with the theoretical analysis. These methods proposed in [31]-[37] focus on the drive system.

3) Active Damping Methods by Auxiliary Circuits

The active damping methods can also be achieved using the additional auxiliary circuits to stabilize the system. In [38], a desired compensating power is injected with the bidirectional converter with the additional super-capacitor. Similarly, the dc electric springs with the bidirectional converter are proposed in [39] as the energy storage system for voltage stabilization and power quality improvement. In [40], the shunt capacitor with the converter is presented as a virtual capacitor to increase the system stability. However, additional circuits are necessary for the methods mentioned above, which increase the investment and the complexity of the overall system.

To summarize, despite the abovementioned studies, few studies have been reported in the literature to develop a practical active damping method for the interlinking three-phase converters that ensures the stability of the system and minimizes the influence on the dynamic performance of the converter. Therefore, a detailed comparison is proposed in Table 2-1.

The virtual impedance-based stabilization methods are proposed in [41]–[43] for improving the stability of the dc/dc converters. These methods could increase

the stability by modifying the phase of the input impedance in the minimum frequency range. However, [41]–[43] are proposed only for the dc/dc converter. Compared to dc/dc converters, the design and analysis of the virtual impedance for the three-phase ac/dc interlinking converters are more complicated.

Table 2-1 Comparison between existing virtual impedance and proposed methods

	[31]–[34]	[35]–[37]	Proposed VIR
Applications	AC motor		AC grid / AC loads
Impedance model	Simplified as ideal constant power loads	Detailed model with high-order terms	
Control of the converter	SVM	PWM	
Design of the virtual impedance	The parameters of the regulator need to be roughly trade-off with the performance		Evaporated design based on the detailed model
Robustness design with model parameter deviation	No		Yes
Digital delay analysis	No		Yes
Influence on the converters' performance	The optimal compensator is not proposed with the theoretical analysis.		Stabilize the system with maximum retention of the dynamic performance

Inspired by the virtual impedance method in [41]–[43], a virtual impedance regulator (VIR) is proposed based on the detailed model of the bus conversion system. Both the grid-following control and the stand-alone control of the interlinking converter are studied in this paper. The idea was first introduced in [44], and the analysis pertaining to it is further extended in this paper. First, is paper presents and analyzes the comprehensive small-signal impedance model and the equivalent circuits of the converter. Then, the design of the proposed VIR is based on the detailed model. The proposed method can improve the stability while minimizing its influence on the converter's dynamic performance because only the phase of the input impedance is modified in the specific interaction frequency range. All parameters for the VIR are selected by theoretical analysis. Moreover, the robust design considering the deviation of the system parameters are analytically implemented and the influence of the digital delay is also considered. Finally, the influence of the virtual impedance regulator applied on the d or q axis is discussed and analyzed.

2.1.3 Model Predictive Control

Continued development of computing has made the high-frequency control methods, such as predictive control, has widely studied in the control of GCIs, with the advantages of satisfying dynamic performance, the non-linearity and

constraints are easily contained in the model and control [45]–[47]. The predictive control algorithm in which the suitable control action is chosen is based on forecasting the system's future performance by using a typical optimization standard [48]–[50].

The application of model predictive control (MPC) in industrial control continues to develop rapidly, mainly due to its simple control logic and responding to problems such as nonlinearity, multi variables, and multiple constraints. MPC applied in the field of power electronics can be divided into two categories: finite control set model predictive control (FCS-MPC) and continuous control set model predictive control (CCS-MPC) [51], [52]. The characteristic continuous MPC is the generalized predictive control (GPC), demonstrated on linear models and simply refers to include constraints [53]. However, the precision of the control in nonlinear systems might be decreased based on GPC. FS-MPC involves only a limited set of viable switching states of the GCIs, then for every state to resolve its corresponding cost function. According to the optimized control parameters, FS-MPC could directly generate the switching signal with a sound dynamic response of power converters. Predictive current control (PCC) is one kind of MPC used for the current control of power converters [54]–[56].

In particular, a weak grid can lead to a large impedance in the grid side, which will be overlapped in the output filter parameters. However, the detailed information of the plant model influences the performance of MPC. Meanwhile, considering the grid voltage could be unbalanced and distorted. The amplitude and frequency of the grid voltage are also varied [57]. Therefore, the control performance could be influenced by the disturbance of equivalent grid impedance.

However, the model used for control could be mismatched significantly in the weak grid due to MPC being deduced through the detailed control plant model. Therefore, grid voltage would be distorted and there is a relatively large grid impedance in this scenario. Both distorted grid voltage and extra grid impedance cannot be considered in the control itself. It is necessary to improve the performance and robustness through the disturbance observer [58].

It could be summarized as two categories for the disturbance observers for the grid voltages: one is estimating the parameters based on the current response and the active excitation; the other is based on the voltage and current measurements to set the state-space observers. The active excitation methods include high-frequency injection [59], excitation of *LCL* filter resonance [60], impedance estimation based on the PWM output switching current ripple, and grid impedance estimation using active and reactive power variation. Noticed that signal injection is required in all these methods to obtain the impedance information. However, this signal injection will introduce additional disturbance. There are some other approaches, such as extended Kalman filter (EKF) observer for estimating the impedance, recursive least square method, and other disturbance observers [61]–[64]. These solutions are received more attention due to not disturbing the grid current. To compare the advantages and disadvantages of closed-loop control and existing FS-MPC control methods, as well as their relation to the proposed control method in terms of cost-effectiveness, and features, as shown in Table 2-2.

Table 2-2 Comparisons between closed-loop PI control, FS-MPC, and DS-MPC

	Closed-loop PI control [47]	FS-MPC [54], [56]	Proposed DS-MPC with EKF observer
Design domain	Frequency	Discrete-time	Discrete-time
Large-signal stability	No	Yes	Yes
Active damping requirement	Multi-loop feedback control	Intuitive turning through cost function	Intuitive turning through cost function
Transient performance	Dependent on PI gain turning	Fast with a high control bandwidth	Fast with high-frequency control
Sensitive to the grid voltage disturbance	Yes	Yes	No
Switching frequency	Fixed	Varying	Fixed
Harmonics	Discrete and quantized	Board and continuous	Regularly spread around the switching frequency

This paper proposes a robust dichotomy method according to the model predictive control (DS-MPC) based on the disturbance observer of the grid voltage through the EKF for the three-phase inverters. The fast dichotomy solution approach is employed in the proposed DS-MPC to dynamically choose and compute the finest voltage vector for every sampling period. Using the FPGA controller, the high-frequency control can be executed in a short time. Besides, the proposed DS-MPC strategy contains the basic FS-MPC and the PWM modulator together with constant frequency essentially. Therefore,

compared with the GCI based on the conventional FS-MPC strategy, it can be simplified for the design of the output filter. With the proposed DS-MPC, the total harmonic distortion (THD) can also be reduced. Moreover, based on the disturbance observer, the model accuracy used in MPC is compensated. Hence, the proposed method for control of the inverter keeps the dynamic characteristics of FS-MPC and presents excellent steady-state performance and high robustness to disturbances of grid voltages at the same time.

2.2 Conclusions

This paper aims to provide solutions to the instability issue of the dc voltage in the interface of the bus conversion system in the hybrid ac/dc grid. In this bus conversion system, due to the interaction of the LC filter with the ac/dc interlinking converter, the stability of the dc voltage might be degraded. Despite the above-mentioned studies, few studies have been reported in the literature to develop a practical active damping method for the interlinking three-phase converters that ensures the stability of the system and minimizes the influence on the dynamic performance of the converter. The abovementioned issues need further insights to enhance the performance of the power conversion system.

Chapter 3 Stability Enhancement for the Bus Conversion System in Hybrid AC/DC Grid

This chapter reveals that the interaction between the interlinking converter and the EMI filter causes instability of the dc voltage in the interface of the bus conversion system in the hybrid ac/dc grid. A virtual impedance regulator is introduced to compensate for the input impedance of the interlinking converter. The robust design of the regulator is ensured by considering the deviation of the system model parameters. The effectiveness of the proposed method to enhance the stability of the dc voltage is validated both theoretically and experimentally.

3.1 Introduction

As novel ac/dc loads and resources are integrated, modern distribution systems become more complex and require more power conversion stages [5]. Thus, to reduce the number of power conversion stages and increase energy efficiency; hybrid ac/dc grids are emerging and attracting more attention [6]. In the hybrid ac/dc grids, the interlinking ac/dc converter is the key component to interface the dc bus with the ac bus. Operating in the control mode, the interlinking converters realize the active power transmission and adjust between the ac and dc grids. However, these high switching frequency converters are also known to bring electromagnetic interference (EMI) noise to the dc bus [9]. To address the EMI issues, an EMI filter [10] (usually an LC filter) can be installed at the dc link to suppress the noise. The interlinking ac/dc converter integrated with the EMI filter can be treated as the bus conversion system, as shown in Figure 3.1.

The EMI filter and the interlinking converter can be deemed as a typical cascaded system where the power-flow front-end converter (i.e., the EMI filter) is defined as the source subsystem and the power-flow back-end converter (i.e., interlinking converter) is defined as the load subsystem [11], as shown in Figure 3.1. When the converter is tightly regulated, it behaves as a constant power load

(CPL) and therefore exhibits a negative incremental input impedance characteristic [12]. The converter's negative input impedance characteristic degrades the cascaded system's stability and may even cause issues like oscillation [13].

Figure 3.1 The diagram for a typical hybrid ac/dc grid.

The virtual impedance regulator (VIR) is proposed based on the detailed model of the bus conversion system. The grid-following control of the interlinking converter is studied in this paper. First, this paper presents and analyzes the comprehensive small-signal impedance model and the equivalent circuits of the converter. Then, the design of the proposed VIR is based on the detailed model. The proposed method can improve the stability while minimizing its influence on the converter's dynamic performance because only the phase of the input impedance is modified in the specific interaction frequency range. All parameters for the VIR are selected by theoretical analysis. Moreover, the robust design considering the deviation of the system parameters is analytically implemented and the influence of the digital delay is also considered. Finally, the influence of the virtual impedance regulator applied on the d or q axis is discussed and analyzed.

3.2 The Degradation Problem of the Studied Systems

Figure 3.2 depicts the studied system, which is consisted of an LC filter and a three-phase converter. The grid-connected converters are typically controlled as current sources injecting current to the grid [65]. The input of the converter is assumed as a stiff dc source supplying the power and neglected the dynamic

from the dc input [66].

To demonstrate the instability problem of the studied system, this instability case is simulated by MATLAB/Simulink, whose results are presented. This section will present the design example with the corresponding system parameters from Table 3-1.

Figure 3.2 The studied bus conversion system in this paper.

(a)

(b)

Figure 3.3 Instability problems of the studied system: a) the simulation results when the converter works stably individually; b) the simulation results when the converter cascaded with the EMI LC filter.

The time-domain simulation results of the system are shown in Figure 3.3. As seen in Figure 3.3 (a), the three-phase converter is working very well

individually. In Figure 3.3 (b), the simulation results of the system are presented, where the three-phase converter is cascaded with an LC filter. Then, both the DC bus voltage and the three-phase converter output current will incur a large oscillation. Therefore, the overall system suffers from instability problems.

Table 3-1 Parameters of the bus conversion system

Parameter	Value
Sampling frequency	10 kHz
Switching frequency	10 kHz
DC bus voltage V_s	360 V
AC bus voltage	110 V (RMS)
Filter inductance L_f	6 mH
AC bus fundamental frequency	50 Hz
DC filter inductance L	0.6 mH
DC filter capacitance C	100 μF
Line resistance R	0.01 Ω

3.3 Modeling of the Interlinking Converter

3.3.1 Modeling of the Interlinking Converter and Its Equivalent Circuit

1) System Modelling

As shown in Figure 3.3, the converter is typically controlled as the current source [65]. The dc-link can be fed by a dc voltage source with v_{dc} or a well-regulated capacitor; for simplification, it is assumed as a dc voltage source. The semiconductor switches are noted as $S_1 \sim S_6$, operate at switching frequency f_0 . The EMI filter on the dc side is L and C , and on the ac side are L_f which are used for shunting the high-frequency harmonics generated by the converter side.

The mathematical model of the grid-connected converter is derived based on the steady-state and dynamic characteristics. The three-phase voltage source with voltage noted as v_{ga} , v_{gb} , v_{gc} . The output currents and voltages of the converter are i_a , i_b , i_c and v_a , v_b , v_c .

$$L_f \frac{d}{dt} \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} = \begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} - \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \cdot r_f - \begin{bmatrix} v_{ga}(t) \\ v_{gb}(t) \\ v_{gc}(t) \end{bmatrix} \quad (3-1)$$

$$i_{dc}(t) = [s_a(t) \ s_b(t) \ s_c(t)] \cdot \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (3-2)$$

$$\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \begin{bmatrix} s_a(t) \\ s_b(t) \\ s_c(t) \end{bmatrix} v_{dc} \quad (3-3)$$

Then the mathematical model can be derived into the synchronous $dq0$ -frame, in which the current can be traced to active and reactive powers, and sinusoidal signals can be traced to dc signals [67]. Therefore, the proportional-integral (PI) controller can realize zero steady-state errors based on the $dq0$ frame.

Figure 3.4 Schematic diagram of the grid-connected converters

Transformation of the model into the synchronous $dq0$ -frame required two steps: Clark transformation from the abc stationary frame into the stationary $\alpha\beta$ -frame, and then rotates signals into the synchronous $dq0$ -frame based on Park transformation. The transfer matrix can be deduced as:

$$T_{abc/dq0} = T_{\alpha\beta0/dq0} \cdot T_{abc/\alpha\beta0} = \frac{2}{3} \begin{bmatrix} \cos \theta & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\ -\sin \theta & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta + \frac{2\pi}{3}) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \quad (3-4)$$

Noted that $T_{abc/dq0}^{-1} = T_{abc/dq0}^T$.

Therefore, the previous mathematical model of the three-phase converter can be deduced as:

$$L_f \frac{d}{dt} \begin{bmatrix} i_d(t) \\ i_q(t) \\ i_0(t) \end{bmatrix} = \begin{bmatrix} s_d(t) \\ s_q(t) \\ s_0(t) \end{bmatrix} v_{dc} - \begin{bmatrix} r_f & -L_f \omega & 0 \\ L_f \omega & r_f & 0 \\ 0 & 0 & r_f \end{bmatrix} \begin{bmatrix} i_d(t) \\ i_q(t) \\ i_0(t) \end{bmatrix} + r_f \begin{bmatrix} v_{gd}(t) \\ v_{gq}(t) \\ v_{g0}(t) \end{bmatrix} \quad (3-5)$$

$$i_{dc}(t) = [s_d(t) \ s_q(t) \ s_0(t)] \cdot \begin{bmatrix} i_d(t) \\ i_q(t) \\ i_0(t) \end{bmatrix} \quad (3-6)$$

The subscript d, q, 0 indicates the value of variables in the d-q reference frame, i_d , i_q , i_0 are the output currents of the converter, v_d , v_q , v_0 are the output voltages of the converter, v_{gd} , v_{gq} , v_{g0} are the three-phase grid voltages, respectively.

2) Average Modelling

Considering a balanced three-phase system, then through the transformation of symmetrical components into the d - q coordinate system, the model can be simplified and converted into dc quantities under the d - q reference frame in steady-state.

The switching function can describe the state of the switches to model the power converters. The switching function can be defined by its algebraic average over a switching cycle, d_d , d_q , d_0 are the duty ratios of the switch in the $dq0$ -frame. Therefore, in each switching cycle T_s , the three-phase converter model in the $dq0$ -frame system ground on the generalized state-space averaging method can be described as

$$L_f \frac{d}{dt} \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} = \begin{bmatrix} d_d \\ d_q \\ d_0 \end{bmatrix} v_{dc} - \begin{bmatrix} r_f & -\omega L_f & 0 \\ \omega L_f & r_f & 0 \\ 0 & 0 & r_f \end{bmatrix} \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} + \begin{bmatrix} \langle v_{gd} \rangle_{T_s} \\ \langle v_{gq} \rangle_{T_s} \\ \langle v_{g0} \rangle_{T_s} \end{bmatrix} \quad (3-7)$$

$$\langle i_{dc} \rangle_{T_s} = [d_d \ d_q \ d_0] \cdot \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} \quad (3-8)$$

where $\langle \cdot \rangle_{T_s}$ denotes the algebraic mean over a switching cycle T_s , ω is the rotor frequency of the grid. It is worth nothing that the 0 component does not exist in a balanced three-phase system. Therefore, the expressions of the 0-axis are neglected.

3) The Steady-State Operating Point

The nonlinear averaged model can be linearized by given by the steady-state analysis. The three-phase converter model, together with the quiescent operating point, can reach the quiescent values.

$$D_d = \frac{V_d}{V_{dc}} \quad (3-9)$$

$$D_q = \frac{V_q}{V_{dc}} \quad (3-10)$$

where D_d, D_q are the switch duty ratio at the quiescent operating point, V_{dc} is the total voltage fed into the converter.

$$V_d = \sqrt{\frac{3}{2}} V_{gm} \quad (3-11)$$

$$V_q = 0 \quad (3-12)$$

where V_d and V_q are the imaginary output voltages of the converter at the quiescent operating point.

$$V_{gd} = \sqrt{\frac{3}{2}} V_m \cos(\theta) \quad (3-13)$$

$$V_{gq} = \sqrt{\frac{3}{2}} V_m \sin(\theta) \quad (3-14)$$

where V_{gd} and V_{gq} are the imaginary grid voltage at the quiescent operating point, V_m is the maximum voltage fed into the converter, θ is the angle between the voltage fed into the converter with the grid.

$$I_d = \frac{r_f (V_d - V_{gd}) - \omega \cdot L_f \cdot V_{gq}}{r_f^2 + (\omega \cdot L_f)^2} \quad (3-15)$$

$$I_q = \frac{(V_{gd} - V_d) \cdot \omega \cdot L_f - r_f \cdot V_{gq}}{r_f^2 + (\omega \cdot L_f)^2} \quad (3-16)$$

where I_d and I_q are the imaginary line current at the quiescent operating point.

4) The Small-Signal Model

can deduce the small-signal model. Also, the equivalent model can be transferred into the Laplace domain to facilitate the controller design.

Therefore, the perturbed state can be described as

$$(sL_f + r_f) \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} = V_{dc} \begin{bmatrix} \hat{d}_d \\ \hat{d}_q \end{bmatrix} + \hat{v}_d \begin{bmatrix} D_d \\ D_q \end{bmatrix} - \begin{bmatrix} 0 & -\omega L_f \\ \omega L_f & 0 \end{bmatrix} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} - \begin{bmatrix} \hat{v}_{gd} \\ \hat{v}_{gq} \end{bmatrix} \quad (3-17)$$

$$\hat{i}_{dc} = [D_d \ D_q] \cdot \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} + [I_d \ I_q] \cdot \begin{bmatrix} \hat{d}_d \\ \hat{d}_q \end{bmatrix} \quad (3-18)$$

where $\hat{\cdot}$ denotes the disturbance of each quantity.

Figure 3.5 Block diagram of the grid-connected converter system plant

Figure 3.6 The equivalent circuit of the small-signal model

The block diagram of plant models is shown in Figure 3.5, where the cross-coupling terms $(-\omega L_f i_d)$, $(\omega L_f i_q)$ are introduced by the abc to $dq0$ transformations can be observed. The equivalent circuit of the small-signal models is displayed in Figure 3.6. The modeling and analysis of the converter system are based on the traditional dq transformation in the synchronous reference frame. The average model is deduced from the mathematical model with its algebraic average over a switching cycle. Then the nonlinear averaged model can be linearized by applying the Taylor Series around the operating point and discarding the high-order terms.

The small-signal model can be rewritten as

$$\begin{bmatrix} \hat{i}_{dc} \\ \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} 1/Z_{in_o} & G_{12} & G_{13} \\ G_{21} & G_{22} & G_{23} \\ G_{31} & G_{32} & G_{33} \end{bmatrix} \begin{bmatrix} \hat{v}_{dc} \\ \hat{d}_d \\ \hat{d}_q \end{bmatrix} \quad (3-19)$$

where

$$\frac{1}{Z_{in_o}(s)} = \frac{\hat{i}_{dc}}{\hat{v}_{dc}} \Big|_{\hat{d}_d=\hat{d}_q=0} = \frac{(D_d^2 + D_q^2)(sL_f + r_f)}{\text{den}(s)} \quad (3-20)$$

$$G_{12}(s) = \frac{\hat{i}_{dc}}{\hat{d}_d} \Big|_{\hat{v}_{dc}=\hat{d}_q=0} = I_d + \frac{(sL_f + r_f)V_d + \omega L_f V_q}{\text{den}(s)} \quad (3-21)$$

$$G_{13}(s) = \frac{\hat{i}_{dc}}{\hat{d}_q} \Big|_{\hat{d}_d=\hat{v}_{dc}=0} = I_q + \frac{(sL_f + r_f)V_q + \omega L_f V_d}{\text{den}(s)} \quad (3-22)$$

$$G_{21}(s) = \frac{\hat{i}_d}{\hat{v}_{dc}} \Big|_{\hat{d}_d=\hat{d}_q=0} = \frac{(sL_f + r_f)D_d + \omega L_f D_q}{\text{den}(s)} \quad (3-23)$$

$$G_{22}(s) = \frac{\hat{i}_d}{\hat{d}_d} \Big|_{\hat{v}_{dc}=\hat{d}_q=0} = \frac{(sL_f + r_f)V_{dc}}{\text{den}(s)} \quad (3-24)$$

$$G_{23}(s) = \frac{\hat{i}_d}{\hat{d}_q} \Big|_{\hat{d}_d=\hat{v}_{dc}=0} = \frac{\omega L_f V_{dc}}{\text{den}(s)} \quad (3-25)$$

$$G_{31}(s) = \frac{\hat{i}_q}{\hat{v}_{dc}} \Big|_{\hat{d}_d=\hat{d}_q=0} = \frac{(sL_f + r_f)D_q - \omega L_f D_d}{\text{den}(s)} \quad (3-26)$$

$$G_{32}(s) = \frac{\hat{i}_q}{\hat{d}_d} \Big|_{\hat{v}_{dc}=\hat{d}_q=0} = \frac{-\omega L_f V_{dc}}{\text{den}(s)} \quad (3-27)$$

$$G_{33}(s) = \frac{\hat{i}_q}{\hat{d}_q} \Big|_{\hat{d}_d=\hat{v}_{dc}=0} = \frac{(sL_f + r_f)V_{dc}}{\text{den}(s)} \quad (3-28)$$

Noticed that $\text{den}(s) = (sL_f + r_f)^2 + (\omega L_f)^2$.

Noted that the Z_{in_o} represents the original open-loop input impedance. Through (3-19), the transfer function of different states can be acquired.

3.3.2 The Open-Loop Input Impedance of the Converter

According to the transfer function in (3-19), the open-loop input impedance $Z_{in_o}(s)$ can be deduced as

$$Z_{in_o}(s) = \frac{(s \cdot L_f + r_f)^2 + (\omega \cdot L_f)^2}{(D_d^2 + D_q^2)(s \cdot L_f + r_f)} \quad (3-29)$$

where $Z_{in_o}(s)$ is the open-loop input impedance.

The bode plot of the open-loop input impedance with its measurement is shown in Figure 3.7. The red dots show the test points under certain frequencies for the input impedance of the converter. The measurement values match very well with the prediction of the developed model.

The equivalent Thévenin circuit of the input impedance of the converter is presented in Figure 3.8.

Figure 3.7 Bode diagram for the open-loop impedance

Figure 3.8 The equivalent Thévenin circuit of the open-loop impedance

The open-loop input impedance of the converter can be considered as an equivalent LC filter. Neglected the resistor r_f , the expressions of the equivalent capacitor, and equivalent

$$C_e = \frac{k \cdot (D_d^2 + D_q^2)}{\omega^2 \cdot L_f} \quad (3-30)$$

$$L_e = \frac{L_f}{D_d^2 + D_q^2} \quad (3-31)$$

where C_e is the equivalent capacitor, L_e is the equivalent inductor.

3.3.3 Close-Loop Input Impedance of the Converter

Figure 3.9 The closed-loop control of the decoupled model

In Figure 3.9, the closed-loop control block of the converter is presented. The input impedance of the converter is derived as follows.

Considering the time delay brought by the digital control [68], the time delay transfer function $G_{de}(s)$ can be presented as

$$G_{de}(s) = \frac{1 - 0.5sT_{de} + (sT_{de})^2}{1 - 0.5sT_{de} + (sT_{de})^2} \quad (3-32)$$

where $T_{de} = 1.5T_{sw}$. T_{sw} presents the sampling period. Current feedforward compensators can mitigate the cross-coupling terms. The input impedance is deduced by the single closed-loop control of the current in the d-q frame. According to the decoupled model of the duty ratio which can be described as

$$\hat{d}_d = -\hat{i}_d \cdot G_{cd} - \frac{\omega \cdot L_f}{V_{dc}} \cdot \hat{i}_q \quad (3-33)$$

$$\hat{d}_q = -\hat{i}_q \cdot G_{cq} + \frac{\omega \cdot L_f}{V_{dc}} \cdot \hat{i}_d \quad (3-34)$$

where G_{cd} , G_{cq} are the inner loop controller to the current in the d-q frame of the closed-loop control. The small-signal model can be incorporated as

$$\begin{bmatrix} \hat{i}_{dc} \\ \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} 1/Z_{in_FB} \\ G_{2_FB} \\ G_{3_FB} \end{bmatrix} \hat{v}_{dc} \quad (3-35)$$

where Z_{in_FB} represents the open-loop output admittance G_{2_FB} and G_{3_FB} represent the transfer function through the d and q-axis components, respectively.

The detailed input impedance transfer function can be described as

$$\begin{aligned}
\frac{1}{Z_{in_FB}(s)} = & \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{13}(s)G_{22}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_1(s)} \\
& + \left(\frac{G_{11}(s)}{2} - \frac{G_{12}(s) \cdot D_d}{V_{dc}} \right) \cdot \frac{T_1(s)}{1+T_1(s)} \\
& + \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{12}(s)G_{33}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_2(s)} \\
& + \left(\frac{G_{11}(s)}{2} - \frac{G_{13}(s) \cdot D_q}{V_{dc}} \right) \cdot \frac{T_2(s)}{1+T_2(s)}
\end{aligned} \tag{3-36}$$

where

$$T_1(s) = G_{cd}(s) \cdot G_{22}(s) \tag{3-37}$$

$$T_2(s) = G_{cq}(s) \cdot G_{33}(s) \tag{3-38}$$

$$H(s) = \frac{\omega \cdot L_f}{V_{dc}} \tag{3-39}$$

The bode plot of the closed-loop input impedance with its measurement of the converter is presented in Figure 3.10. As G_{22} is the same as G_{33} in this model, setting $G_{cd} = G_{cq}$ for simplification can be deduced that $T_1 = T_2$. The change of the current controller parameters will not influence the magnitude of the input impedance in the low-frequency range. Then it can be calculated that the cut-off frequency of T_1 is equivalent to the cut-off frequency of the input impedance of the converter.

Figure 3.10 Bode diagram for the closed-loop input impedance of the converter.

Therefore, the input impedance can be further simplified into two parts. The low-frequency range mainly presents as

$$\frac{1}{Z_{in_FB_L}(s)} = G_{11}(s) - \frac{G_{12}(s)D_d + G_{13}(s)D_q}{V_{dc}} \tag{3-40}$$

and the high-frequency band as

$$\frac{1}{Z_{in_FB_H}(s)} = G_{11}(s) - \frac{H(s) \left(\begin{matrix} G_{13}(s)G_{22}(s)D_d \\ + G_{12}(s)G_{33}(s)D_q \end{matrix} \right)}{V_{dc}} \quad (3-41)$$

The bode plots for the $Z_{in_FB_L}$ and $Z_{in_FB_H}$ are presented in Figure 3.11. The bode plot for the low-frequency range and the high-frequency range of the input impedance can be given.

Figure 3.11 Bode diagram for the closed-loop input impedance of the converter: (a) $Z_{in_FB_L}$, (b) $Z_{in_FB_H}$.

The high-order model can be further simplified as its corresponding mainly presents frequency range. Therefore, the characteristics of the close-loop input impedance are summarized as follows:

a) The closed-loop input impedance is a negative resistor in the low-frequency range (lower than the cut-off frequency).

b) In the high-frequency band (higher than the cut-off frequency), the closed-loop input impedance behaves like an equivalent inductor just like the open-loop input impedance.

Note that the high-frequency band is inductive, which is the same as the open-loop. The input impedance of the load converters is necessary to perform as a negative resistor to meet the dynamic requirement. (3-42)

$$R_e = - \frac{V_{dc}^2}{I_d \cdot V_d + I_q \cdot V_q} \quad (3-43)$$

$$L_e = \frac{L}{D_d^2 + D_q^2} \quad (3-44)$$

where R_e is the equivalent resistor, L_e is the equivalent inductor.

3.3.4 Degradation Problem of the Studied System

The output impedance of the LC filter can be given as

$$Z_{out}(s) = \frac{sL + R_f}{s^2LC + sCR_f + 1} \quad (3-45)$$

where R_f is the inner resistor in series with the filter inductor.

Figure 3.12 Instability problem assessment of the converter cascaded with EMI LC filter: a) The bode plot of the closed-loop input impedance of the converter and the output impedance of the EMI LC filter; b) The Nyquist diagram of the ratio of the EMI LC filter's output impedance and converter's input impedance.

In Figure 3.12 (a), the bode plot of the closed-loop input impedance of the converter, Z_{in-FB} , and the output impedance of the EMI LC filter, Z_{out} , are depicted in the same figure. As seen, there is an intersection with $|Z_{in-FB}|$ and $|Z_{out}|$ during the frequency range (f_1, f_2) . Since Z_{in-FB} is presented as a negative resistor in this frequency range, the phase difference between $\phi(Z_{in-FB})$ and $\phi(Z_{out})$ is larger than 180° . Thus, the ratio of Z_{out} and Z_{in-FB} will encircle $(-1, j0)$ in the Nyquist diagram as shown in Figure 3.12 (b). As a result, the studied system suffers instability problems from the small-signal analysis point of view.

3.4 Virtual Impedance Regulator for Stabilizing the Bus Conversion System

3.4.1 Concept of the Proposed VIR

1) The Function of VIR

It is necessary to regulate the negative resistor characteristics of the three-phase interlinking converter to improve the stability of this cascaded system. However, the negative resistor characteristics are the embodiment of CPL, which guarantees the dynamic performance of the converter. Thus, the negative resistor characteristics cannot be removed from the converter itself.

To solve the above compromise, the VIR is proposed to regulate the input impedance of the converter only in the desired frequency range, where the converter's input impedance is intersected by the EMI LC filter's output impedance (i.e., (f_1, f_2) in Figure 3.12 (a)). In this intersection frequency range, only the phase of the input impedance is modified while the magnitude remains the same. As a result, the degraded converters' original dynamic performance can be minimized when stabilizing the whole system.

Figure 3.13 Concept of the proposed virtual impedance regulator.

The concept of the proposed VIR is presented in Figure 3.13. The traditional PLL and feedback control of the interlinking converter is also formulated in the dq frame. As seen, the proposed VIR control strategy can add special regulators to the d or q frame, i.e., $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$. Reasonable design of $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$ can suitably design the Z_{d_VIR} and Z_{q_VIR} to regulate the input impedance of the three-phase interlinking converter only in the desired frequency range.

2) The Desired Modified Input Impedance $Z_{in_M}(s)$

The virtual impedance is applied only within the intersection frequencies of $|Z_{in_FB}|$ and $|Z_{out}|$. In the intersection frequency range, the magnitude will

maintain the same, and only the frequency change from -180 degrees to 180 degrees. Then the system's stability will improve and minimize the influence on the converter's dynamic performance.

The preferred impedance Z_{in_M} can be described as

$$Z_{in_M}(s) = \begin{cases} |Z_{in_FB}(s)| & f \in [f_1, f_2] \\ Z_{in_FB}(s) & f \notin [f_1, f_2] \end{cases} \quad (3-46)$$

where Z_{in_FB} is the original closed-loop input impedance of the three-phase converter.

The desired modified input impedance of the three-phase interlinking converter is shown in Figure 3.14.

Figure 3.14 The preferred modified input impedance of the converter and the realization

3) The expression of $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$

There is a parallel virtual impedance $Z_{d_VIR}(s)$ or $Z_{q_VIR}(s)$ added into the original impedance. The parallel virtual impedance $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ can be presented as

$$Z_{d_VIR}(s) = \left[\frac{1}{Z_{in_M}(s)} - \frac{1}{Z_{in_FB}(s)} \right]^{-1} \quad (3-47)$$

$$Z_{q_VIR}(s) = \left[\frac{1}{Z_{in_M}(s)} - \frac{1}{Z_{in_FB}(s)} \right]^{-1} \quad (3-48)$$

$Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ can be deduced as

$$Z_{d_VIR}(s) = \begin{cases} \frac{|Z_{in_FB}(s)|}{2} & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (3-49)$$

$$Z_{q_VIR}(s) = \begin{cases} \frac{|Z_{in_FB}(s)|}{2} & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (3-50)$$

$Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ are typical frequency segmentation functions. The band-pass filter $G_{BPF}(s)$ is used to realize Z_{q_VIR} and Z_{d_VIR} . The second-order band-pass filter $G_{BPF}(s)$ is used to realize the close-loop impedance in d and q-axis components within the intersection frequencies of $|Z_{in_FB}|$ and $|Z_{out}|$.

A typical second-order band-pass filter $G_{BPF}(s)$ can be given as

$$G_{BPF}(s) = \frac{s^2}{s^2 + (\omega_1 / Q_H)s + \omega_1^2} \cdot \frac{\omega_2^2}{s^2 + (\omega_2 / Q_L)s + \omega_2^2} \quad (3-51)$$

where Q_H and Q_L are the quality factors of the high-pass filter and low-pass filter, which is generally chosen as 0.707 with good amplitude-frequency characteristics of $G_{BPF}(s)$; f_1 and f_2 are the characteristic frequencies of a high-pass filter and a low-pass filter, respectively.

$$f_1 = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e} \quad (3-52)$$

$$f_2 = \frac{-L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e} \quad (3-53)$$

Then the realization of $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ virtual impedance can be deduced as

$$Z_{d_VIR}(s) = \frac{|Z_{in_FB}(s)|}{2} G_{BPF}(s) \quad (3-54)$$

$$Z_{q_VIR}(s) = \frac{|Z_{in_FB}(s)|}{2} G_{BPF}(s) \quad (3-55)$$

Therefore, according to the concept of VIR, the main objective of the VIR control strategy is to design suitable $G_{q_VIR}(s)$ and $G_{d_VIR}(s)$ to realize the preferred $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$.

3.4.2 Derivation of $G_{d_VIR}(s)$

1) The Impact of $G_{d_VIR}(s)$ on $Z_{in_M}(s)$

$G_{d_VIR}(s)$ is put between the input voltage and the input current in the d-axis components of the converter. Then the duty ratio in the d-axis based on the small-signal model can also be deduced as

$$\hat{d}_d = -\hat{i}_d \cdot G_{cd}(s) - H(s) \cdot \hat{i}_q + \hat{V}_{dc} \cdot G_{d_VIR}(s) \quad (3-56)$$

Note that (3-56) together with (3-19), the small-signal model can be incorporated as

$$\begin{bmatrix} \hat{i}_{dc} \\ \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} 1/Z_{in_M} \\ G_{d_M} \\ G_{q_M} \end{bmatrix} \hat{v}_{dc} \quad (3-57)$$

where Z_{in_M} , G_{d_M} , and G_{q_M} represent the modified input impedance and the transfer function through the d and q-axis components when $G_{d_VIR}(s)$ is utilized, respectively.

The realization of the VIR in the d-axis can be shown as

$$\begin{aligned} \frac{1}{Z_{in_M}(s)} = & \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{13}(s)G_{22}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_1(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{G_{12}(s) \cdot D_d}{V_{dc}} \right) \cdot \frac{T_1(s)}{1+T_1(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{12}(s)G_{33}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_2(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{G_{13}(s) \cdot D_q}{V_{dc}} \right) \cdot \frac{T_2(s)}{1+T_2(s)} \\ & + G_{d_VIR}(s) \cdot G_{12}(s) \cdot \frac{1}{1+T_1(s)} \end{aligned} \quad (3-58)$$

Compared $Z_{in_M}(s)$ with the $Z_{in_FB}(s)$ in (3-36), the additional term can deduce the virtual impedance $Z_{d_VIR}(s)$. $1/Z_{in_M}(s)$ can be re-written as $1/Z_{in_FB}(s)$ in parallel with $1/Z_{d_VIR}(s)$

$$\frac{1}{Z_{in_M}(s)} = \frac{1}{Z_{in_FB}(s)} + \underbrace{G_{d_VIR}(s) \cdot G_{12}(s) \cdot \frac{1}{1+T_1(s)}}_{\frac{1}{Z_{d_VIR}(s)}} \quad (3-59)$$

Figure 3.15 presents the equivalent circuit of the modified input impedance of the converter through the VIR in d frame, where Z_{d_VIR} is in parallel with Z_{in_FB} to generate the Z_{in_M} .

Figure 3.15 Equivalent circuit of the modified input impedance of the converter with the proposed virtual impedance in d frame.

2) The Expression of $Z_{d_VIR}(s)$

$Z_{d_VIR}(s)$ can be expressed as

$$Z_{d_VIR}(s) = \begin{cases} \frac{1+T_1(s)}{G_{d_VIR}(s) \cdot G_{12}(s)} & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (3-60)$$

3) The Expression of $G_{d_VIR}(s)$

Therefore, the realization of the piecewise function $G_{d_VIR}(s)$ can be made by the band-pass filter $G_{BPF}(s)$. Therefore, $G_{d_VIR}(s)$ can be represented as

$$G_{d_VIR}(s) = 2 \cdot \frac{[1+T_1(s)]}{G_{12}(s) \cdot |Z_{in_FB}(s)|} \cdot G_{BPF}(s) \quad (3-61)$$

3.4.3 Derivation of $G_{q_VIR}(s)$

1) The Impact of $G_{q_VIR}(s)$ on $Z_{in_M}(s)$

Since the proposed VIR is required to damp the peak of $|Z_{in_FB}|$, it can also be realized by putting $G_{q_VIR}(s)$ between the input voltage and the input current in the q-axis components converter. The regulated impedance is Z_{in_M} . Then the duty ratio in the q-axis based on the small-signal model can also be deduced as

$$\hat{d}_q = -\hat{i}_q \cdot G_{cq}(s) + H(s) \cdot \hat{i}_d + \hat{V}_{dc} \cdot G_{q_VIR}(s) \quad (3-62)$$

Note that the (3-62) together with (3-19), the small-signal model can be deduced as

$$\begin{bmatrix} \hat{i}_{dc} \\ \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} 1/Z_{in_M} \\ G_{in_Md} \\ G_{in_Mq} \end{bmatrix} \hat{v}_{dc} \quad (3-63)$$

where Z_{in_M} represents the open-loop input impedance with the VIR in q-axis. G_{in_Md} represents the transfer function through the d-axis component, G_{in_Mq} represents the transfer function through the q-axis component, respectively.

The realization of the VIR in the q-axis can be shown as

$$\begin{aligned} \frac{1}{Z_{in_M}(s)} = & \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{13}(s)G_{22}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_1(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{G_{12}(s) \cdot D_d}{V_{dc}} \right) \cdot \frac{T_1(s)}{1+T_1(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{H(s)G_{12}(s)G_{33}(s)D_d}{V_{dc}} \right) \cdot \frac{1}{1+T_2(s)} \\ & + \left(\frac{G_{11}(s)}{2} - \frac{G_{13}(s) \cdot D_q}{V_{dc}} \right) \cdot \frac{T_2(s)}{1+T_2(s)} \\ & + G_{q_VIR}(s) \cdot G_{13}(s) \cdot \frac{1}{1+T_2(s)} \end{aligned} \quad (3-64)$$

Then the compared (3-64) with the $Z_{in_FB}(s)$ in the (3-19), there is a parallel virtual impedance $Z_{q_VIR}(s)$ added into the modified impedance.

$$\frac{1}{Z_{in_M}(s)} = \frac{1}{Z_{in_FB}(s)} + \underbrace{G_{q_VIR}(s) \cdot G_{13}(s) \cdot \frac{1}{1+T_2(s)}}_{\frac{1}{Z_{q_VIR}(s)}} \quad (3-65)$$

Figure 3.16 shows the equivalent circuit introduced parallel virtual impedance in the q-axis.

Figure 3.16 The equivalent circuit of the input impedance with VIR in q-axis

2) The Expression of $Z_{q_VIR}(s)$

Then, the virtual impedance in q-axis components $Z_{q_VIR}(s)$ can be obtained as

$$Z_{q_VIR}(s) = \begin{cases} \frac{1+T_2(s)}{G_{q_VIR}(s) \cdot G_{13}(s)} & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (3-66)$$

3) The Expression of $G_{q_VIR}(s)$

Therefore, the realization of the piecewise function $G_{q_VIR}(s)$ can be made by the band-pass filter $G_{BPF}(s)$. Therefore, the expression of $G_{q_VIR}(s)$ can be present as

$$G_{q_VIR}(s) = 2 \cdot \frac{[1 + T_2(s)]}{G_{13}(s) \cdot |Z_{in_FB}(s)|} \cdot G_{BPF}(s) \quad (3-67)$$

3.4.4 Comparison of the $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$

1) Impact of G_{d_VIR} on i_d and i_q

The transfer function G_{in_Md} and G_{in_Mq} between the disturbance of the output currents of the converter and the total voltage fed into the converter can be represented as

$$\hat{i}_d = \frac{D_d}{G_{cd} \cdot V_{dc} + s \cdot L_f + r_f} \cdot \hat{V}_{dc} + \underbrace{\frac{G_{d_VIR} \cdot V_{dc}}{G_{cd} \cdot V_{dc} + s \cdot L_f + r_f}}_{\text{impact of } i_d} \cdot \hat{V}_{dc} \quad (3-68)$$

$$\hat{i}_q = \frac{D_q}{G_{cq} \cdot V_{dc} + s \cdot L_f + r_f} \cdot \hat{V}_{dc} + \underbrace{0 \cdot \hat{V}_{dc}}_{\text{impact of } i_q} \quad (3-69)$$

where D_d and D_q is deduced based on the statistic point. Note that the impact of G_{d_VIR} on i_d is shown in (3-68) and there is no impact on i_q .

Then, the influence of the transfer function G_{in_Md} can be deduced. G_{in_Md} and G_{in_Mq} can be expressed as

$$G_{in_Md} = \frac{\hat{i}_d}{\hat{V}_{dc}} = \frac{D_d + G_{d_VIR} \cdot V_{dc}}{G_{cd} \cdot V_{dc} + s \cdot L_f + r_f} \quad (3-70)$$

$$G_{in_d} = \frac{G_{d_VIR} \cdot V_{dc}}{G_{cd} \cdot V_{dc} + s \cdot L_f + r_f} \quad (3-71)$$

2) Impact of G_{q_VIR} on i_d and i_q

The transfer function G_{in_Md} between the disturbance of the output currents of the converter in the d-axis component and the total voltage fed into the converter can be present as

$$\hat{i}_d = \frac{D_d}{G_{cd} \cdot V_{dc} + s \cdot L_f + r_f} \cdot \hat{V}_{dc} + \underbrace{0 \cdot \hat{V}_{dc}}_{\text{impact of } i_d} \quad (3-72)$$

$$\hat{i}_q = \frac{D_q}{G_{cq} \cdot V_{dc} + s \cdot L_f + r_f} \cdot \hat{V}_{dc} + \underbrace{\frac{G_{q_VIR} \cdot V_{dc}}{G_{cq} \cdot V_{dc} + s \cdot L_f + r_f} \cdot \hat{V}_{dc}}_{\text{impact of } i_q} \quad (3-73)$$

Note that there is no impact on i_d from G_{q_VIR} according to the (3-72), and the impact on i_d is shown in (3-73).

Figure 3.17 Bode plot of $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$

The bode plot of G_{d_VIR} and G_{q_VIR} is shown in Figure 3.17. As seen, there is little difference between the two transfer functions. Because there is no reactive power in this scenario and D_q is zero; therefore, the proposed virtual impedance regulator might not be effective. It is recommended that choosing the virtual impedance regulator in the d axis is more applicable in most cases. However, the proposed method can still apply to the q axis if there is reactive power. Therefore, this thesis recommends choosing $G_{d_VIR}(s)$ to modify the input impedance of the three-phase interlinking converter in this paper.

3.4.5 Robust Design for the Deviation of the LC filter

The deviation of the LC filter can be considered in the design of the virtual impedance regulator with a certain redundancy.

If the parameters of the LC filter deviated $\alpha\%$ from the original value of the inductance L , and $\beta\%$ deviated from the original value of the capacitance C . Then, based on (3-52) and (3-53) for the frequency of the band-pass filter $G_{BPF}(s)$, the corresponding frequency range of the band-pass filter can be

$$f_a = \frac{(1+\alpha\%)L - \sqrt{4(1+\alpha\%)L(1+\beta\%)CR_e^2 + (1+\alpha\%)^2 L^2}}{4\pi(1+\alpha\%)L(1+\beta\%)CR_e} \quad (3-74)$$

$$f_b = -\frac{(1+\alpha\%)L + \sqrt{4(1+\alpha\%)L(1+\beta\%)CR_e^2 + (1+\alpha\%)^2 L^2}}{4\pi(1+\alpha\%)L(1+\beta\%)CR_e} \quad (3-75)$$

where f_a and f_b are the frequency of a high-pass filter, and a low-pass filter considering the parameters deviated from the original value of the filter, α and β are the deviated percentage of the inductance L and the capacitance C compared to the original value, respectively.

Then the frequency margin of the proposed virtual impedance regulator is based on considering the parameters of the input LC filter deviating a certain percentage $\gamma\%$ from the original value. Considering all cases when the inductance L and the capacitance C deviate $\gamma\%$, $-\gamma\%$ from its original value or remain the same, there will be a total of 9 different cases (including the initial case), as shown in Table 3-2.

Table 3-2 Equations of the characteristic frequencies

	Inductance	Capacitance	Equations
Case 1	L	C	$f_{a1} = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e}$
			$f_{b1} = \frac{-L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e}$
Case 2	$(1+\gamma\%) L$	$(1+\gamma\%) C$	$f_{a2} = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi(1+\gamma\%)LCR_e}$
			$f_{b2} = -\frac{L + \sqrt{4LCR_e^2 + L^2}}{4\pi(1+\gamma\%)LCR_e}$
Case 3	$(1-\gamma\%) L$	$(1-\gamma\%) C$	$f_{a3} = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi(1-\gamma\%)LCR_e}$
			$f_{b3} = -\frac{L + \sqrt{4LCR_e^2 + L^2}}{4\pi(1-\gamma\%)LCR_e}$
Case 4	$(1+\gamma\%) L$	$(1-\gamma\%) C$	$f_{a4} = \frac{(1+\gamma\%)L - \sqrt{4(1-(\gamma\%)^2)LCR_e^2 + (1+\gamma\%)^2 L^2}}{4\pi(1-(\gamma\%)^2)LCR_e}$
			$f_{b4} = -\frac{(1+\gamma\%)L + \sqrt{4(1-(\gamma\%)^2)LCR_e^2 + (1+\gamma\%)^2 L^2}}{4\pi(1-(\gamma\%)^2)LCR_e}$

	Inductance	Capacitance	Equations
Case 5	$(1-\gamma\%) L$	$(1+\gamma\%) C$	$f_{a5} = \frac{(1-\gamma\%)L - \sqrt{4(1-(\gamma\%)^2)LCR_e^2 + (1-\gamma\%)^2 L^2}}{4\pi(1-(\gamma\%)^2)LCR_e}$
			$f_{b5} = -\frac{(1-\gamma\%)L + \sqrt{4(1-(\gamma\%)^2)LCR_e^2 + (1-\gamma\%)^2 L^2}}{4\pi(1-(\gamma\%)^2)LCR_e}$
Case 6	$(1+\gamma\%) L$	C	$f_{a6} = \frac{L - \sqrt{(4LCR_e^2)/(1+\gamma\%) + L^2}}{4\pi LCR_e}$
			$f_{b6} = -\frac{L + \sqrt{(4LCR_e^2)/(1+\gamma\%) + L^2}}{4\pi LCR_e}$
Case 7	$(1-\gamma\%) L$	C	$f_{a7} = \frac{L - \sqrt{(4LCR_e^2)/(1-\gamma\%) + L^2}}{4\pi LCR_e}$
			$f_{b7} = -\frac{L + \sqrt{(4LCR_e^2)/(1-\gamma\%) + L^2}}{4\pi LCR_e}$
Case 8	L	$(1+\gamma\%) C$	$f_{a8} = \frac{L - \sqrt{4(1+\gamma\%)LCR_e^2 + L^2}}{4\pi(1+\gamma\%)LCR_e}$
			$f_{b8} = -\frac{L + \sqrt{4(1+\gamma\%)LCR_e^2 + L^2}}{4\pi(1+\gamma\%)LCR_e}$
Case 9	L	$(1-\gamma\%) C$	$f_{a9} = \frac{L - \sqrt{4(1-\gamma\%)LCR_e^2 + L^2}}{4\pi(1-\gamma\%)LCR_e}$
			$f_{b9} = -\frac{L + \sqrt{4(1-\gamma\%)LCR_e^2 + L^2}}{4\pi(1-\gamma\%)LCR_e}$

Then in these cases, the minimum frequency of f_a can be selected as the characteristic frequency of the low-pass filter f_{amin} , it can be seen that case 2 satisfy this criterion. The maximum frequency of f_b can be selected as the characteristic frequency of the high-pass filter, and case 3 satisfies this criterion.

$$f_{a \min} = \min(f_{a1}, f_{a2}, f_{a3}, f_{a4}, f_{a5}, f_{a6}, f_{a7}, f_{a8}, f_{a9}) = f_{a2} = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi(1+\gamma\%)LCR_e} \quad (3-76)$$

$$f_{b \max} = \max(f_{b1}, f_{b2}, f_{b3}, f_{b4}, f_{b5}, f_{b6}, f_{b7}, f_{b8}, f_{b9}) = f_{b3} = -\frac{L + \sqrt{4LCR_e^2 + L^2}}{4\pi(1-\gamma\%)LCR_e} \quad (3-77)$$

The frequency of a high-pass filter f_L and a low-pass filter f_H can be deduced when considering the parameters deviating $\gamma\%$ from its original value is as

$$f_L = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi(1 + \gamma\%)LCR_e} \quad (3-78)$$

$$f_H = \frac{L + \sqrt{4LCR_e^2 + L^2}}{4\pi(1 - \gamma\%)LCR_e} \quad (3-79)$$

f_L is the minimum frequency when both the inductance and capacitance of the LC filter decrease $\gamma\%$. f_H is the maximum frequency when both the inductance and capacitance of the LC filter increased $\gamma\%$.

If considering a 50% deviation from the original value of the input LC filter, the deviated parameters of the LC filter and its corresponding characteristic frequencies of the band-pass filter f_a and f_b are as shown in Table 3-3. The selected f_L and f_H are under case 2 and case 3, respectively, which are texted in bold in Table 3-3.

Table 3-3 Parameters of the characteristic frequencies

	Inductance L	Capacitance C	f_a	f_b
Case 1: Initial case	0.6 mH	100 μF	538 Hz	784 Hz
Case 2: Increase 50% both on L and C	0.9 mH	150 μF	358 Hz	522 Hz
Case 3: Decrease 50% both on L and C	0.3 mH	50 μF	1076 Hz	1568 Hz
Case 4: Increase 50% L and decrease 50% C	0.9 mH	50 μF	543 Hz	1035 Hz
Case 5: Increase 50% C and decrease 50% L	0.3 mH	150 μF	672 Hz	836 Hz
Case 6: Only increase 50% L	0.9 mH	100 μF	421 Hz	667 Hz
Case 7: Only decrease 50% L	0.3 mH	100 μF	804 Hz	1049 Hz
Case 8: Only increase 50% C	0.6 mH	150 μF	454 Hz	618 Hz
Case 9: Only decrease 50% C	0.6 mH	50 μF	705 Hz	1196 Hz

Figure 3.18 Bode plot of the output impedance of the LC filter with 9 cases.

The virtual impedance regulator with a certain redundancy of the variation of the input filter can be calculated. The bode diagram for the output impedance of the LC filter, including from case 1 to case 9 is shown in Figure 3.18. As

mentioned before, case 2 and case 3 are selected as f_L and f_H , the bode plot for the output impedance of the LC filter, including from case 1 to case 3, and the modified input impedance of three-phase converter with $G_{d_VIR}(s)$ are shown in Figure 3.19. $Z_{out_+50\%}$ is the output impedance in case 2, and $Z_{out_ -50\%}$ is the output impedance of the LC filter in case 3. As seen, with the $G_{d_VIR}(s)$, the studied system is stable, and the input impedance of the three-phase converter is modified with 50% redundancy on the input filter.

Figure 3.19 Bode plot of the output impedance of the LC filter and the input impedance of the converter with VIR.

3.4.6 Design Example

According to the above analysis in the comparison part of VIR, $G_{d_VIR}(s)$ is selected as the VIR to shape the input impedance of the three-phase interlinking converter in this design example. The detailed parameters are presented in Table 3-1.

The bode diagram for input and output impedance with the proposed VIR is shown in Figure 3.20 (a). The Nyquist diagram of the impedance ratio of the cascaded system is shown in Figure 3.20 (b). The red dots show the test points under certain frequencies for the input impedance of the converter.

There is only a small range of frequency modified for the proposed VIR to stabilize the cascaded system, so the dynamic performance is ensured in this case. Furthermore, the design of this method is based on the magnitude and phase of the three-phase converter's input impedance. Thus, the original system is degraded, and the proposed VIR could stable the system.

Figure 3.20 Bode plot and Nyquist curve of the system with the proposed VIR: (a) Bode plot (b) Nyquist diagram of the impedance ratio.

The bode plot of G_{in_Md} with G_{d_VIR} based is shown in Figure 3.21. The VIR in d-axis will bring little influence on the system, which ensured the dynamic response of the overall system.

Figure 3.21 Bode plot of G_{in_d} and G_{in_Md}

3.5 Experimental Results

In this part, the proposed VIR is experimentally verified on a real ac cascaded system. The hardware prototype of the proposed cascaded system is shown in Figure 3.22.

A dSPACE DS1202 (MicroLabBox) is utilized as the digital controller with its internal slave Xilinx FPGA board responsible for modulation and generating the

gate signals for all the module converters. The overvoltage/overcurrent protect functions are also realized in the slave FPGA.

Figure 3.22 A photograph of the experimental platform.

3.5.1 System Stability

Figure 3.23 Instability evaluation results of the experimental system: (a) the original system; (b) the modified system with the proposed VIR (ac current and dc voltage); (c) the modified system with the proposed VIR (ac voltage and dc voltage).

The experimental results of the experimental system before and after being equipped with the proposed VIR are given in Figure 3.23.

It can be seen that the waveforms of the v_{dc} and i_a, i_b, i_c are clearly shown

the oscillation in Figure 3.23 (a). When the VIR is incorporated, the experimental system becomes stable, and no oscillation occurs in v_{dc} and i_a, i_b, i_c in Figure 3.23 (b). In Figure 3.23 (c), there is little influence on the ac voltage since the ac side is directly connected to the stiff ac bus.

3.5.2 Dynamic Performance

The dynamic performance of the experimental system with VIR is tested in Figure 3.24. The load steps from 100% to 50% full load. The system is relatively stable, and dynamic performance is desirable. The response time is about 20ms. As seen, this VIR can stabilize the system with the good dynamic performance of the three-phase converter.

Figure 3.24 The experiment results of the modified system with the proposed VIR when the dynamic load changes from 100% full-load to 50% full-load: (a) ac current; (b) ac voltage.

3.5.3 Experiments for the Robust Design

The experimental validation considering the deviation of the LC filter of the system is given in Figure 3.25. There are three cases of different values for the LC input filter, considering the increase of 50% as Case 2, and decreased 50% as Case 3 on both the inductance and capacitor.

The oscillation is more severe in Case 2 compared with all cases. The system can be stable even if the parameters of the EMI filter deviate from their original value. The selected f_L and f_H are under case 2 and case 3, respectively, which are texted in bold in Table 3-4.

Figure 3.25 Instability problem of the studied system with a variation of EMI filter: a) case 1: original (L : 0.6 mH, C : 100 μF); b) case 2: increased 50% (L : 0.9 mH, C : 150 μF); c) case 3: decreased 50% (L : 0.3 mH, C : 50 μF).

Table 3-4 Parameters of the characteristic frequencies

	Inductance L	Capacitance C	f_L	f_H
Case 1: Initial case	0.6 mH	100 μF	538 Hz	784 Hz
Case 2: Increase 50%	0.9 mH	150 μF	358 Hz	522 Hz
Case 3: Decrease of 50%	0.3 mH	50 μF	1076Hz	1568Hz

From Figure 3.23 to Figure 3.25, the effectiveness of the proposed VIR stabilization method is verified. As seen, this VIR can stabilize the system with the good dynamic performance of the converter.

3.6 Conclusions

This paper aims to provide solutions to the instability issue of the dc voltage in the interface of the bus conversion system in the hybrid ac/dc grid. In this bus conversion system, due to the interaction of the LC filter with the ac/dc interlinking converter, the stability of the dc voltage might be degraded. In this paper, the virtual impedance regulator (VIR) is proposed to remove the negative effect of the three-phase converter. The phase of the input impedance of the converter is modified only in the specific desired frequency range. Each selection of the parameters for the VIR is supported by theoretical analysis. As

a result, this method can stabilize the system with maximum retention of the dynamic performance of the interlinking converter. The proposed method can be implemented in the d-axis or q-axis control. The robust design of the proposed VIR method accounts for the deviations of the system model parameters. The experimental results verify the correctness of the proposed VIR stabilization method.

Chapter 4 Stability Control of Point-of-Load Converter in DC Microgrids

This chapter proposes a series-connected virtual impedance regulator based on the detailed small-signal model of the POL converter that feeds a linear three-phase load. The proposed method can improve the stability by modifying only the phase of the input impedance of the POL converter within the interaction frequency range. Therefore, the impact of the regulator on the converter's dynamic performance is minimized. Under different load scenarios, the experiments show that the proposed stabilization methods exhibit good performance under different load scenarios.

4.1 Introduction

Driven by environmental and energy concerns, the amount of dc distributed generation systems (DGSs), and microgrids are dramatically growing to integrate renewable energy resources[69]. As a result, the common dc interface is introduced to reduce the back and forth ac/dc or dc/ac conversions and simplify the dc microgrids [70].

Figure 4.1 Diagram of typical dc microgrids

A typical dc microgrid is shown in Figure 4.1. The power sources are connected to the common dc bus (i.e., dc/dc converters for solar panels and fuel

cells, ac/dc converters for wind turbines). Then, the point-of-load (POL) converters are deployed to deliver the common DC bus power to a variety of possible loads [71]. However, these high switching frequency converters will introduce electromagnetic interference (EMI) noise to the common dc bus [9]. Thus, to suppress the EMI noise, an input EMI filter (usually an LC filter) is usually required [72].

The input EMI filter and the POL converter form a typical cascaded system, where the input filter is defined as the source subsystem, and the POL converter is defined as the load subsystem. The tightly regulated POL converter can be deemed constant power loads (CPLs) [12]. Since CPL has incremental negative input impedance characteristics, it may deteriorate the stability of the dc voltage in the interface of the bus conversion system. Furthermore, the interaction between the input LC filter with the POL converter may bring instability issues and even lead to complex oscillation problems in the whole system.

Inspired by the active damping method, a series-connected virtual impedance regulator (VIR) is proposed based on improving the stability of the dc microgrid in Figure 4.1. The VIR is deduced according to the detailed impedance model of the POL converter. The design of the VIR is based on modifying the phase of the input impedance only in the oscillation frequency range. Hence, the proposed VIR can minimize the regulator's impact on the POL converter's closed-loop performances while stabilizing the dc-link voltage. The derivation and verification of the VIR are all carefully analyzed and discussed. The proposed VIR can be implemented either on the d-axis or q-axis, and its influence is discussed and analyzed.

4.2 The Degradation Problem of the Studied Systems

As shown in Figure 4.2, the studied system with POL converter feeds a local load is presented. The input LC filter is integrated with the three-phase VSC to suppress the EMI noise. The output currents and the voltages of the converter are represented as i_a, i_b, i_c and, v_a, v_b, v_c . The local load voltages are shown as v_{La}, v_{Lb}, v_{Lc} ; i_{dc} and v_{dc} and the total current and voltage fed into the VSC, respectively.

To demonstrate the instability problem of the studied system, this instability case is simulated by MATLAB/Simulink, whose results are presented. This section will present the design example with the corresponding system parameters from Table 4-1.

Figure 4.2 Circuit topology of VSC fed on the dc-link bus and supplying an ac load

Figure 4.3 DC bus voltage and three-phase output current performance with different DC bus capacitance: (a) electrolytic capacitor; (b) LC EMI filter

The time-domain simulation results of the converter system are shown in Figure 3.3. As seen in Figure 4.3 (a), the three-phase converter is working very well individually. In Figure 4.3 (b), the simulation results of the system are presented, where the three-phase converter is cascaded with an LC filter. Then, both the DC bus voltage and the three-phase converter output current will incur

a large oscillation. Therefore, the overall cascaded system suffers from instability problems.

Table 4-1 Parameters of the stand-alone DGS

Parameter	Value
DC bus voltage V_{dc}	300 V
AC bus voltage	110 V (RMS)
DC filter inductance L	0.6 mH
DC filter capacitance C	100 μF
Grid fundamental frequency	10 Hz
Switching frequency	10 kHz
AC filter inductance L	6 mH
AC filter capacitance C	100 μF
Resistance Load R	12 Ω

4.3 Modeling of the Three-Phase Voltage Source Converter

4.3.1 Modeling of the VSC with Its Equivalent Circuit

The small-signal model impedance is deduced based on linearizing the average model around the steady-state operation point and neglecting the high-order terms.

1) System Modelling

Figure 4.4 Schematic diagram of the grid-connected three-phase converters

Figure 3.2 shows the schematic diagram of the stand-alone three-phase converters controlled as ac voltage sources. A dc can feed the dc-link voltage source with v_{dc} or a well-regulated capacitor; for simplification, it is assumed as

a dc voltage source. The semiconductor switches are noted as $S_1 \sim S_6$, operate at switching frequency f_0 . The EMI filter on the dc side is L and C , and on the ac side are L_f and C_f which are used for shunting the high-frequency harmonics generated by the converter side. The load voltage and current noted as v_{La} , v_{Lb} , v_{Lc} , and i_{La} , i_{Lb} , i_{Lc} . The output currents and voltages of the converter are i_a , i_b , i_c and v_a , v_b , v_c . The local load is noted as z in this chapter. For simplification analysis, the load is considered balanced.

The mathematical model of the grid-connected converter is derived based on the steady-state and dynamic characteristics of the three-phase converters. The voltage between node N and n is noted as v_{Nn} .

$$L_f \frac{d}{dt} \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} = \begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} - \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \cdot r_f - \begin{bmatrix} v_{La}(t) \\ v_{Lb}(t) \\ v_{Lc}(t) \end{bmatrix} - \begin{bmatrix} v_{Nn}(t) \\ v_{Nn}(t) \\ v_{Nn}(t) \end{bmatrix} \quad (4-1)$$

$$C_f \frac{d}{dt} \begin{bmatrix} v_{La}(t) \\ v_{Lb}(t) \\ v_{Lc}(t) \end{bmatrix} = \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} - \begin{bmatrix} i_{La}(t) \\ i_{Lb}(t) \\ i_{Lc}(t) \end{bmatrix} \quad (4-2)$$

$$i_{dc}(t) = [s_a(t) \ s_b(t) \ s_c(t)] \cdot \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (4-3)$$

$$\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \begin{bmatrix} s_a(t) \\ s_b(t) \\ s_c(t) \end{bmatrix} v_{dc} \quad (4-4)$$

Noted that $v_{Nn}(t) = \frac{1}{3} \cdot (v_a(t) + v_b(t) + v_c(t))$ for the three-phase three-wire systems.

Inserting (3-4) in (3-1), it can be obtained,

$$L_f \frac{d}{dt} \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 \cdot v_a(t) - v_b(t) - v_c(t) \\ 2 \cdot v_b(t) - v_a(t) - v_c(t) \\ 2 \cdot v_c(t) - v_a(t) - v_b(t) \end{bmatrix} - \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \cdot r_f - \begin{bmatrix} v_{La}(t) \\ v_{Lb}(t) \\ v_{Lc}(t) \end{bmatrix} \quad (4-5)$$

The semiconductor switches functions are noted as S_a , S_b , S_c , corresponds to the on and off state with 1 and 0. Following this definition $v_x(t) = v_{dc} \cdot S_x(t)$, ($x = a, b, c$) can be satisfied, which yields

$$L_f \frac{d}{dt} \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} = \frac{1}{3} v_{dc} \begin{bmatrix} 2 \cdot S_a(t) - S_b(t) - S_c(t) \\ 2 \cdot S_b(t) - S_a(t) - S_c(t) \\ 2 \cdot S_c(t) - S_a(t) - S_b(t) \end{bmatrix} - \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \cdot r_f - \begin{bmatrix} v_{La}(t) \\ v_{Lb}(t) \\ v_{Lc}(t) \end{bmatrix} \quad (4-6)$$

2) Average Modelling

The model needs to convert from the symmetrical components into the d - q coordinate system. The model can be simplified and converted into dc quantities under the d - q reference frame in the steady-state.

The switching function can describe the state of the switches to model the power converters. The switching function can be defined by its algebraic average over a switching cycle. Therefore, in each switching cycle T_s , the three-phase converter model in the d - q frame system ground on the generalized state-space averaging method can be described as

$$L_f \frac{d}{dt} \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} = \begin{bmatrix} d_d \\ d_q \\ d_0 \end{bmatrix} \langle v_{dc} \rangle_{T_s} - \begin{bmatrix} r_f & -\omega L_f & 0 \\ \omega L_f & r_f & 0 \\ 0 & 0 & r_f \end{bmatrix} \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} - \begin{bmatrix} \langle v_{Ld} \rangle_{T_s} \\ \langle v_{Lq} \rangle_{T_s} \\ \langle v_{L0} \rangle_{T_s} \end{bmatrix} \quad (4-7)$$

$$C_f \frac{d}{dt} \begin{bmatrix} \langle v_{La} \rangle_{T_s} \\ \langle v_{Lb} \rangle_{T_s} \\ \langle v_{Lc} \rangle_{T_s} \end{bmatrix} = \begin{bmatrix} \langle i_a \rangle_{T_s} \\ \langle i_b \rangle_{T_s} \\ \langle i_c \rangle_{T_s} \end{bmatrix} - \begin{bmatrix} \langle i_{La} \rangle_{T_s} \\ \langle i_{Lb} \rangle_{T_s} \\ \langle i_{Lc} \rangle_{T_s} \end{bmatrix} \quad (4-8)$$

$$\langle i_{dc} \rangle_{T_s} = [d_d \quad d_q \quad d_0] \cdot \begin{bmatrix} \langle i_d \rangle_{T_s} \\ \langle i_q \rangle_{T_s} \\ \langle i_0 \rangle_{T_s} \end{bmatrix} \quad (4-9)$$

where $\langle \cdot \rangle_{T_s}$ denotes the algebraic mean over a switching cycle T_s , ω is the rotor frequency of the grid, the subscript d, q, 0 indicates the value of variables in the d-q reference frame, d_s , d_q , d_0 are the duty ratios of the switch in the d-q reference frame, respectively.

3) The Steady-State Operating Point

The nonlinear averaged model can be linearized by given by the steady-state analysis. The three-phase converter model, together with the quiescent operating point, can reach the quiescent values. According to the state-space model, the steady-state operating point are described below,

$$D_d = \frac{V_d}{V_{dc}} \quad (4-10)$$

$$D_q = \frac{V_q}{V_{dc}} \quad (4-11)$$

where D_d, D_q are the switch duty ratio at the quiescent operating point, V_{dc} is the total voltage fed into the converter.

$$V_d = \sqrt{\frac{3}{2}} V_m \quad (4-12)$$

$$V_q = 0 \quad (4-13)$$

where V_d and V_q are the imaginary output voltages of the converter at the quiescent operating point. V_m is the maximum voltage fed into the converter.

$$V_{Ld} = \sqrt{\frac{3}{2}} V_m \cos(\theta) \quad (4-14)$$

$$V_{Lq} = \sqrt{\frac{3}{2}} V_m \sin(\theta) \quad (4-15)$$

where V_{Ld} and V_{Lq} are the imaginary load voltages at the quiescent operating point. θ is the angle between the voltage fed into the converter.

$$I_d = \frac{(V_{dc} D_q - V_{Lq}) L_f \omega + (V_{dc} D_d - V_{Ld}) r}{(L_f \omega)^2 + r^2} \quad (4-16)$$

$$I_q = \frac{(V_{dc} D_q - V_{Lq}) r - (V_{dc} D_d - V_{Ld}) L_f \omega}{(L_f \omega)^2 + r^2} \quad (4-17)$$

where I_d and I_q are the imaginary line currents at the quiescent operating point.

4) The Small-Signal Model

The relationship of the small variations around the operation point can deduce the small-signal model. Also, the equivalent model can be transferred into the Laplace domain to facilitate the controller design.

The formulas with the perturbed state can be transformed into

$$(sL_f + r_f) \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} = V_{dc} \begin{bmatrix} \hat{d}_d \\ \hat{d}_q \end{bmatrix} + \hat{v}_d \begin{bmatrix} D_d \\ D_q \end{bmatrix} - \begin{bmatrix} 0 & -\omega L_f \\ \omega L_f & 0 \end{bmatrix} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} - \begin{bmatrix} \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} \quad (4-18)$$

$$C_f s \begin{bmatrix} \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} = \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} - \begin{bmatrix} \hat{i}_{Ld} \\ \hat{i}_{Lq} \end{bmatrix} - \begin{bmatrix} 0 & -\omega C_f \\ \omega C_f & 0 \end{bmatrix} \begin{bmatrix} \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} \quad (4-19)$$

$$\hat{i}_{dc} = [D_d \ D_q] \cdot \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} + [I_d \ I_q] \cdot \begin{bmatrix} \hat{d}_d \\ \hat{d}_q \end{bmatrix} \quad (4-20)$$

where $\hat{\cdot}$ denotes the disturbance of each quantity.

Therefore, the equivalent circuit for the small-signal model of the three-phase VSC is shown in Figure 4.5. The block diagram of plant models is shown in Figure 4.6, where the cross-coupling terms $(-\omega L_f i_d)$, $(\omega L_f i_q)$ and $(-\omega C_f v_{Ld})$, $(\omega C_f v_{Lq})$ are introduced by the abc to $dq0$ transformations can be observed.

Figure 4.5 The equivalent circuit of the small-signal model

Figure 4.6 Block diagram of the system plant of the three-phase converter

4.3.2 The Open-Loop Input Impedance

The small-signal model can be rewritten as

$$\begin{bmatrix} \hat{v}_{dc} \\ \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} = \begin{bmatrix} Z_{in0} & G_{12} & G_{13} \\ G_{21} & G_{22} & G_{23} \\ G_{31} & G_{32} & G_{33} \end{bmatrix} \begin{bmatrix} \hat{i}_{dc} \\ \hat{d}_d \\ \hat{d}_q \end{bmatrix} = \begin{bmatrix} \frac{num_{11}}{den} & \frac{num_{12}}{den} & \frac{num_{13}}{den} \\ \frac{num_{21}}{den} & \frac{num_{22}}{den} & \frac{num_{23}}{den} \\ \frac{num_{31}}{den} & \frac{num_{32}}{den} & \frac{num_{33}}{den} \end{bmatrix} \begin{bmatrix} \hat{i}_{dc} \\ \hat{d}_d \\ \hat{d}_q \end{bmatrix} \quad (4-21)$$

For resistance load R , the transfer function can be derived as:

$$num_{11} = L_f^2 (\omega^2 + (G_a C_f R + s)^2) + 2RL_f (G_b C_f R + s) + R^2 \quad (4-22)$$

$$\begin{aligned} num12 = & -\left(2G_b I_d C_f L_f + V_{dc}(sD_d + \omega D_q)C_f + I_d\right)R^2 + I_d L_f^2 G_a (C_f^2 \omega^2 R^2 + (C_f sR + 1)^2) \\ & + \left(2sI_d L_f + D_d V_{dc}\right)R + V_{dc} L_f (C_f R + 1)^2 (sD_d - \omega D_q) \end{aligned} \quad (4-23)$$

$$num13 = -I_d \left(\left((C_f sR + 1)^2 + (C_f R\omega)^2 \right) G_a L_f^2 + 2RL_f (G_b RC_f + s) + R^2 \right) \quad (4-24)$$

$$num21 = R \left(D_d (1 + L_f C_f G_b + sL_f) + D_q L_f \omega (2C_f Rs + 1) \right) \quad (4-25)$$

$$num22 = R \left(D_d \left(\left((1 + L_f C_f G_b)R + sL \right) I_d \right) + D_q \left(\left(RC_f (2L_f sI_d + D_d V_{dc}) + L_f I_d \right) \omega \right) \right. \\ \left. - (C_f Rs + 1) D_q V_{dc} \right) \quad (4-26)$$

$$num23 = -I_d R \left(\left(C_f (D_d G_b + 2D_q s\omega) L_f + D_d \right) R + L_f (D_d s + D_q \omega) \right) \quad (4-27)$$

$$num31 = R \left(D_d L_f \omega (2C_f Rs + 1) - D_q (1 + L_f C_f G_b + sL_f) \right) \quad (4-28)$$

$$num32 = R \left(D_d \left(\left((1 + L_f C_f G_b)R + sL \right) I_d \right) - D_d (L_f I_d \omega (2C_f Rs + 1) - (C_f Rs + 1) D_q V_{dc}) \right) \quad (4-29)$$

$$num33 = -RI_d \left(\left((s^2 - \omega^2) D_q - 2D_d \omega s \right) L_f + D_q \right) RC_f + L_f (D_q s - D_d \omega) \quad (4-30)$$

$$G_a = s^2 + \omega^2 \quad (4-31)$$

$$G_b = s^2 - \omega^2 \quad (4-32)$$

$$\text{Noticed that } \text{den}(s) = (D_d^2 + D_q^2) \left(L_f s \left((C_f R\omega)^2 + (C_f Rs + 1)^2 \right) + (sC_f R + 1)R \right)$$

Noted Z_{ino} represents the open-loop input impedance for the VSC. Hence, the transfer functions of the small-signal model for the VSC includes (4-21): Z_{ino} , G_{12} , G_{13} , G_{21} , G_{22} , G_{23} , G_{31} , G_{32} and G_{33} .

The open-loop input impedance $Z_{ino}(s)$ can be derived based on (4-21) when the local load is resistive (R), inductive (R and L), and capacitive (R and C):

$$Z_{ino_R}(s) = \frac{\left(L_f^2 (\omega^2 + (C_f G_a R + s)^2) + 2RL_f (G_b C_f R + s) + R^2 \right)}{(D_d^2 + D_q^2) \left(L_f s \left((C_f R\omega)^2 + (C_f Rs + 1)^2 \right) + (sC_f R + 1)R \right)} \quad (4-33)$$

$$Z_{ino_RL}(s) = \frac{\left(\left((C_f L_f G_a)^2 + 1 \right) \cdot ((Ls + R)^2 + \omega^2) + L_f (Ls^2 + Rs + \omega^2) \right. \\ \left. + 2C_f \left(L_f^2 G_a (Ls^2 + Rs - \omega^2) \right) + L_f G_b (2LRs + R^2 - \omega^2) \right) + L_f^2 G_a}{\left(C_f^2 L_f s G_a ((Ls + R)^2 + \omega^2) + (L_f + L)s + R \right) (D_d^2 + D_q^2)} \\ \left(+ C_f s ((Ls + R)^2 + 2(Ls^2 + Rs - \omega^2)L + \omega^2) \right) \quad (4-34)$$

$$Z_{ino_RC}(s) = \frac{(G_a(C_f + C)RL_f + sL_f)^2 + R^2 + 2G_b(C_f + C)L_fR^2 + 2sRL_f + L^2\omega^2}{\left(s(C_f + C)(1 + G_a(C_f + C)L_f)R^2 + (1 + 2L_f s^2(C_f + C))R + L_f s \right) \cdot (D_d^2 + D_q^2)} \quad (4-35)$$

where $Z_{ino_R}(s)$, $Z_{ino_RL}(s)$ and $Z_{ino_RC}(s)$ are the open-loop input impedance under the three-phase resistive load, inductive load, and capacitive load, respectively.

The characteristics of the open-loop input impedance can be analyzed based on its bode diagram, for the resistive load case is shown in Figure 4.7. The converter's open-loop input impedance with resistive load can be regarded as an inductor L_{eR} connected in series with a resistor R_{eR} . The equivalent LR circuit is presented in Figure 4.8.

Figure 4.7 Bode diagram for the open-loop input impedance of the converter with resistive load

Figure 4.8 The equivalent circuit of the open-loop input impedance of the converter with resistive load

The expressions of the equivalent resistance and equivalent inductor are as follows:

$$R_{eR} = \frac{R^2 (C_f L_f \omega^2 + 1)^2 + L_f^2 \omega^2}{R(D_d^2 + D_q^2)} \quad (4-36)$$

$$L_{eR} = \frac{L_f}{D_d^2 + D_q^2} \quad (4-37)$$

where R_{eR} is the equivalent resistor, L_{eR} is the equivalent inductor.

As shown in Figure 4.9, the open-loop input impedance of the converter under inductive load can be regarded as an inductor L_{eL} connected in series with a capacitor C_{eL} . The equivalent LC circuit is presented in Figure 4.10.

Figure 4.9 Bode diagram for the open-loop input impedance of the converter with inductive loads

Figure 4.10 The equivalent circuit of the open-loop input impedance of the converter with inductive load

The expressions of the equivalent capacitor and equivalent inductor are as follows:

$$C_{eL} = \frac{\left((C_f \omega^2 - 1)^2 L_f^2 - 2L_f C_f (\omega^2 - 1) + 1 \right) \omega^2 + R^2 (C_f \omega^2 - 1)}{R(D_d^2 + D_q^2)} \quad (4-38)$$

$$L_{eL} = \frac{L_f}{D_d^2 + D_q^2} \quad (4-39)$$

where C_{eL} is the equivalent capacitor, L_{eL} is the equivalent inductor.

The open-loop input impedance under the capacitive load can be regarded as an inductor L_{eC} connected in series with a resistor R_{eC} , as shown in Figure 4.11. The equivalent LR circuit is presented in Figure 4.12.

The expressions of the equivalent resistance and equivalent inductor are as follows:

$$R_{eC} = \frac{R^2 \left((C_f + C) L_f \omega^2 + 1 \right)^2 + L_f^2 \omega^2}{R (D_d^2 + D_q^2)} \quad (20(a))$$

$$L_{eC} = \frac{L_f}{D_d^2 + D_q^2} \quad (20(b))$$

where R_{eC} is the equivalent capacitor, L_{eC} is the equivalent inductor.

Figure 4.11 Bode diagram for the open-loop input impedance of the converter with the capacitive load

Figure 4.12 The equivalent circuit of the open-loop input impedance of the converter with the capacitive load

Therefore, for the high-frequency band (higher than the cut-off frequency), the closed-loop input impedance behaves as the same equivalent inductor for the three loads mentioned above types.

4.3.3 Close-Loop Input Impedance

The double-loop control strategy for the three-phase VSC is given. $G_{id}(s)$ and $G_{iq}(s)$ are the current regulator. The cross-coupling effects need to be handled by adding the voltage and current feedforward terms to separate the controller in the d and q channels. The time delay is also considered with $G_{SPWM}(s)$. In summary, this chapter establishes a dual-loop control strategy with the capacitive current feedback as the inner loop.

1) Current Inner Loop Decoupling Control

First, taking the capacitor current loop as an example, the current decoupling control strategy is given, as shown in Figure 4.13.

Figure 4.13 Capacitance current loop decoupling control block diagram

$G_{id}(s)$ and $G_{iq}(s)$ are the current regulator, which can be implemented as a proportional-integral (PI). The PI controller adjusts the deviation between the actual capacitor current value and the controller reference value. Considering the coupling effect of the filter inductor, the feedforward cross-coupling terms $(-\omega L_f i_d)$, $(\omega L_f i_q)$ are introduced and subtracted to achieve the decoupling effect. The cross-coupling effects need to be ignored by adding the voltage and current feedforward terms to design the controller independently in the d and q axis separately. Therefore, according to the decoupled model of the current inner loop, the duty ratio which can be described as

$$\hat{d}_d = \frac{G_{SPWM}(s)}{V_{dc_ref}} \cdot (\hat{v}_{Ld} - G_{id}(s) \hat{i}_d - \omega L_f \hat{i}_q) \quad (4-40)$$

$$\hat{d}_q = \frac{G_{SPWM}(s)}{V_{dc_ref}} \cdot (\hat{v}_{Lq} - G_{iq}(s) \hat{i}_q + \omega L_f \hat{i}_d) \quad (4-41)$$

The modulated reference value is obtained after the decoupling mitigation. Then, the converter's output voltage is obtained by PWM modulation, and the actual output capacitance current value as well, and finally completes the feedback control design of the current inner loop.

Noted that the $G_{SPWM}(s)$ presents the small inertia and delay loop in Sinusoidal PWM (SPWM) modulation control. The modulation ratio is 1 for the dc voltage feedback loop is involved in SPWM modulation. The sampling period is noted as T .

$$G_{SPWM}(s) = \frac{1}{1.5Ts + 1} \quad (4-42)$$

Figure 4.14 Simplified current loop control block diagram

The simplified capacitor current loop block diagram after decoupling is shown in Figure 4.14. It can be seen that the coupling terms are mitigated by each other; the decoupled dq axes are independent of each other.

Then the closed-loop transfer function of the capacitor current inner loop in the d and q axis can be deduced as $G_{id_c}(s)$ and $G_{iq_c}(s)$ separately.

$$G_{id_c}(s) = \frac{G_{id}(s)G_{SPWM}(s)}{sL_f + G_{id}(s)G_{SPWM}(s)} \quad (4-43)$$

$$G_{iq_c}(s) = \frac{G_{iq}(s)G_{SPWM}(s)}{sL_f + G_{iq}(s)G_{SPWM}(s)} \quad (4-44)$$

2) Voltage Outer Loop Decoupling Control

Figure 4.15 Voltage outer loop decoupling control block diagram

Similar to the decoupling control strategy of the capacitor current loop, the voltage outer loop decoupling control block diagram is shown in Figure 4.15.

In Figure 4.15, $G_{vd}(s)$ and $G_{vq}(s)$ are the voltage regulator, which can be implemented as a proportional-integral (PI). The voltage loop PI controller adjusts the difference between the actual capacitor voltage value and the reference. At the same time, considering the coupling term of the capacitor $(-\omega C_f v_{Ld})$ and $(\omega C_f v_{Lq})$, decoupling is achieved by involving the feedforward terms, and finally, the current inner loop reference is obtained. The simplified voltage loop control block diagram is given in Figure 4.16.

Figure 4.16 Simplified voltage loop control block diagram

Therefore, according to the decoupled model of voltage loop, the output currents of the converter which can be described as

$$\hat{i}_d = -G_{idq-c}(s) \cdot (G_{vd}(s) \hat{v}_{Ld} + \omega C_f \hat{v}_{Lq}) \quad (4-45)$$

$$\hat{i}_q = G_{idq-c}(s) \cdot (\omega C_f \hat{v}_{Ld} - G_{vq}(s) \hat{v}_{Lq}) \quad (4-46)$$

3) Voltage and Current Double-Loop Control

Figure 4.17 Voltage and current double-loop control block diagram

Based on the previous analysis, the coupling terms are mitigated, and the decoupled d - q axes are independent of each other, and the control loops between the d and q axes are the same. Therefore, the double loop control is shown in Figure 4.17. In addition, the capacitor current feedforward is introduced to eliminate the influence of the output voltage when the load changes.

The single closed-loop control of the deduces the input impedance current in

the d - q frame. According to the decoupled model of the duty ratio which can be described as

$$\hat{d}_d = G_{de1}(s)\hat{v}_{Ld} - G_{de2}(s)\hat{v}_{Lq} \quad (4-47)$$

$$\hat{d}_q = G_{de2}(s)\hat{v}_{Ld} + G_{de1}(s)\hat{v}_{Lq} \quad (4-48)$$

$$G_{de1}(s) = (1 + G_{idq_c}(s)(G_i(s)G_v(s) - \omega^2 L_f C_f))G_{SPWM}(s) \quad (4-49)$$

$$G_{de2}(s) = (G_{idq_c}(s)(\omega L_f G_v(s) + \omega C_f G_i(s)))G_{SPWM}(s) \quad (4-50)$$

4) Close-Loop Input Impedance

Note that (4-47) to (4-50) together with the original small-signal model (4-21), it can be incorporated with the closed-loop control as

$$\begin{bmatrix} \hat{v}_{dc} \\ \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} = \begin{bmatrix} G_{inc} \\ G_{dC} \\ G_{qC} \end{bmatrix} \hat{i}_{dc} \quad (4-51)$$

where G_{inc} represents the open-loop input impedance, G_{dC} and G_{qC} are the transfer function from d and q channel load voltage to dc side current, respectively.

The detailed closed-loop input impedance transfer function can be deduced as below.

Figure 4.18 Close-loop Impedances of the three-phase VSC

Suppose

$$G_{denc}(s) = (G_{22}G_{33} - G_{23}G_{32})(G_{de1}^2 + G_{de2}^2) - (G_{22} + G_{33})G_{de1} + (G_{32} - G_{23})G_{de2} + 1 \quad (4-52)$$

Then the input impedance is

$$Z_{inC} = \frac{1}{G_{denc}} \left((G_{13}G_{22} - G_{12}G_{23})G_{31} + G_{21}(G_{12}G_{33} - G_{13}G_{32}) + G_{ap} \right) (G_{de1}^2 + G_{de2}^2) + \frac{1}{G_{denc}} \left((G_{12}G_{21} + G_{13}G_{31} - (G_{22} + G_{33}))G_{de1} + ((G_{32} - G_{23}) + G_{12}G_{31} - G_{13}G_{21})G_{de2} + 1 \right) \quad (4-53)$$

The bode plot of the closed-loop input impedance is presented in Figure 4.18. The characteristics of the input impedance are summarized as follows:

- a) In the low-frequency range (lower than the cut-off frequency), the closed-loop input impedance is a negative resistor R_e with all loads types.
- b) In the high-frequency band (higher than the cut-off frequency), the closed-loop input impedance behaves like an equivalent inductor L_e .

Figure 4.19 The equivalent Thévenin circuit of the closed-loop impedance

(a)

(b)

(c)

(d)

Figure 4.20 Instability problem assessment of the cascaded system with LC filter: a) Bode plot of Z_{inC_R} , Z_{inC_L} , Z_{inC_C} , and Z_{out} ; The Nyquist diagram of the impedance ratio of b) Z_{inC_R} with Z_{out} , c) Z_{inC_L} with Z_{out} , d) Z_{inC_C} with Z_{out}

Note that the high-frequency band is inductive, which is the same as the

open-loop. The input impedance of the load converters is necessary to perform as a negative resistor in order to meet the dynamic requirement.

$$L_e = \frac{L}{D_d^2 + D_q^2} \quad (4-54)$$

where R_e is the equivalent resistor, L_e is the equivalent inductor.

4.3.4 Degradation Problem of the Studied System

In Figure 4.20 (a), the bode plot of the closed-loop input impedance of the VSC, Z_{inC} , and the output impedance of the LC filter, Z_{out} , are depicted in the same figure. As seen, during the frequency range (f_1, f_2) , $|Z_{inC}|$ and $|Z_{out}|$ have an intersection. Since Z_{inC} is presented as a negative resistor at this frequency range, the difference between $\varphi(Z_{inC})$ and $\varphi(Z_{out})$ is larger than 180° . Thus, the ratio of Z_{out} and Z_{inC} at different loads, including resistive, inductive, and capacitive loads, will encircle $(-1, j0)$ in the Nyquist diagram as shown in Figure 4.20 (b), (c), and (d). As a result, the studied system suffers instability problems from the small-signal analysis point of view. Therefore, it is necessary to compensate for the negative impedance characteristic and improve the system's stability.

4.4 Stabilization via VIR Control Strategies

4.4.1 Concept of the Proposed VIR

The negative impedance characteristics of the well-regulated VSC should be compensated to solve the system instability problem. However, the negative resistor characteristics are the embodiment of CPL, which guarantees the dynamic performance of the three-phase VSC. Thus, the negative resistor characteristics cannot be entirely removed. It is a trade-off.

Therefore, to deal with the compromise mentioned above, the VIR is proposed to regulate the input impedance only in the oscillation frequencies range where the converter's input impedance is intersected by the LC filter's output impedance (i.e., (f_1, f_2) in Figure 4.20 (a)). Within this frequency range, the magnitude of the input impedance remains the same, and only the phase is changed to improve the stability. As a result, the impact of the VIR on the original

dynamic performance of the POL converter can be minimized.

Figure 4.21 The small-signal control block diagram with the VIR

As shown in Figure 4.21, the design concept of the proposed VIR is presented. The proposed VIR can be added in both d and q frames, i.e., $G_{d-VIR}(s)$ and $G_{q-VIR}(s)$. The function of $G_{d-VIR}(s)$ and $G_{q-VIR}(s)$ are applied to the virtual impedance Z_{d-VIR} or Z_{q-VIR} in series with the VSC to regulate the converter's input impedance, respectively. Reasonable design can suitably design the virtual impedance to regulate the input impedance only in the required frequency range.

1) The Preferred Stabilized Input Impedance Z_{in_M}

The preferred modified input impedance of the three-phase VSC is shown in Figure 4.22.

Figure 4.22 The preferred input impedance of the three-phase grid-connected converter and the realization

Therefore, the desired input impedance Z_{in_M} can be described based on

Figure 4.22.

$$Z_{in_M}(s) = \begin{cases} |Z_{inC}(s)| & f \in [f_1, f_2] \\ Z_{inC}(s) & f \notin [f_1, f_2] \end{cases} \quad (4-55)$$

where Z_{inC} is the initial input impedance with closed-loop control, and Z_{in_M} is the desired closed-loop input impedance.

2) The Expression of $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$

The VIR is connected in series with the input port of the three-phase VSC, as shown in Figure 4.22. Thus, the input impedance will be modified with the virtual impedance $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$, which can be obtained as

$$Z_{d_VIR}(s) = Z_{in_M}(s) - Z_{in_FB}(s) \quad (4-56)$$

$$Z_{q_VIR}(s) = Z_{in_M}(s) - Z_{in_FB}(s) \quad (4-57)$$

Then, $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ can be further deduced as

$$Z_{d_VIR}(s) = \begin{cases} 2 \cdot |Z_{in_FB}(s)| & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (4-58)$$

$$Z_{q_VIR}(s) = \begin{cases} 2 \cdot |Z_{in_FB}(s)| & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (4-59)$$

Z_{q_VIR} and Z_{d_VIR} are typical frequency segmentation functions. Thus, the band-pass filter $G_{BF}(s)$ is utilized to implement Z_{q_VIR} and Z_{d_VIR} . A second-order band-pass filter $G_{BF}(s)$ can be presented as

A typical second-order band-pass filter $G_{BPF}(s)$ can be given as

$$G_{BF}(s) = \frac{s^2}{s^2 + (2\pi f_1 / Q_H)s + \omega_1^2} \cdot \frac{(2\pi f_2)^2}{s^2 + (2\pi f_2 / Q_L)s + \omega_2^2} \quad (4-60)$$

where ω_1 and ω_2 are the characteristic angular frequencies of a high-pass filter and a low-pass filter Q_H and Q_L are the quality factors of the high-pass filter and low-pass filter; respectively.

Then, the preferred $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$ can be deduced

$$Z_{d_VIR}(s) = 2 \cdot |Z_{in_FB}(s)| G_{BPF}(s) \quad (4-61)$$

$$Z_{q_VIR}(s) = 2 \cdot |Z_{in_FB}(s)| G_{BPF}(s) \quad (4-62)$$

Therefore, according to the concept of VIR, the main task of the VIR control strategy is to design suitable $G_{q_VIR}(s)$ and $G_{d_VIR}(s)$ to realize the preferred $Z_{d_VIR}(s)$ and $Z_{q_VIR}(s)$.

4.4.2 Derivation of $G_{d_VIR}(s)$

1) The Impact of $G_{d_VIR}(s)$ on $Z_{in_M}(s)$

According to Figure 4.22, $G_{d_VIR}(s)$ is the transfer function added to the original d channel control system to mimic $Z_{d_VIR}(s)$. Hence, the duty ratio in the d channel with VIR can be expressed as:

$$\hat{d}_d = G_{de1}(s)\hat{v}_{Ld} - G_{de2}(s)\hat{v}_{Lq} + G_{d_VIR}(s)\hat{i}_{dc} \quad (4-63)$$

Note that (4-63) together with (4-21), the small-signal model can be incorporated with VIR in the d channel. Then Z_{in_M} represents the modified input impedance of the VSC when $G_{d_VIR}(s)$ is utilized. The small-signal model can be rearranged as

$$\begin{bmatrix} \hat{v}_{dc} \\ \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} = \begin{bmatrix} Z_{in_M} \\ G_{in_Md} \\ G_{in_Mq} \end{bmatrix} \hat{i}_{dc} \quad (4-64)$$

Compared with the original $Z_{inC}(s)$ in (4-51), $Z_{in_M}(s)$ can be re-written as $Z_{in_FB}(s)$ in series with $Z_{d_VIR}(s)$.

$$Z_{in_M}(s) = Z_{inC}(s) + \underbrace{\frac{1}{G_{denc}} \left((G_{13}G_{32} - G_{12}G_{33})G_{de1} + (G_{13}G_{22} - G_{12}G_{23})G_{de2} + G_{12} \right)}_{Z_{d_VIR}(s)} \cdot G_{d_VIR} \quad (4-65)$$

Figure 4.23 shows the equivalent circuit of the modified input impedance with series-connected VIR in the d channel, where $Z_{d_VIR}(s)$ is in series with $Z_{inC}(s)$ to generate the $Z_{in_M}(s)$.

Figure 4.23 The equivalent circuit of the input impedance with VIR

2) The Expression of $Z_{d_VIR}(s)$

Based on (4-65), $Z_{d_VIR}(s)$ can be expressed as

$$Z_{d_VIR}(s) = \begin{cases} \frac{G_{d_VIR}}{G_{denc}} \left((G_{13}G_{32} - G_{12}G_{33})G_{de1} + (G_{13}G_{22} - G_{12}G_{23})G_{de2} + G_{12} \right) & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (4-66)$$

3) The Expression of $G_{d_VIR}(s)$

Also, the realization of the piecewise function $G_{d_VIR}(s)$ can be deduced according to the band-pass filter $G_{BF}(s)$. Therefore, $G_{d_VIR}(s)$ can be represented as

$$G_{d_VIR}(s) = 2 \cdot \frac{G_{denc} |Z_{inC}|}{\left((G_{13}G_{32} - G_{12}G_{33})G_{de1} + (G_{13}G_{22} - G_{12}G_{23})G_{de2} + G_{12} \right)} \cdot G_{BF} \quad (4-67)$$

4.4.3 Derivation of $G_{q_VIR}(s)$

1) The Impact of $G_{q_VIR}(s)$ on $Z_{in_M}(s)$

$G_{q_VIR}(s)$ is the transfer function added to the original q channel control system to mimic $Z_{q_VIR}(s)$. Hence, the duty ratio in the d channel with VIR can be expressed as:

$$\hat{d}_q = G_{de2}(s)\hat{v}_{Ld} + G_{de1}(s)\hat{v}_{Lq} + G_{q_VIR}(s)\hat{i}_{dc} \quad (4-68)$$

Note that (4-68) together with (4-21), the small-signal model can be incorporated. Then Z_{in_M} represents the modified input impedance of the VSC when $G_{q_VIR}(s)$ is utilized.

$$\begin{bmatrix} \hat{v}_{dc} \\ \hat{v}_{Ld} \\ \hat{v}_{Lq} \end{bmatrix} = \begin{bmatrix} Z_{in_M} \\ G_{in_Md} \\ G_{in_Mq} \end{bmatrix} \hat{i}_{dc} \quad (4-69)$$

where Z_{in_M} represents the open-loop input impedance with the VIR in q-axis. G_{in_Md} represents the transfer function through the d-axis component, G_{in_Mq} represents the transfer function through the q-axis component, respectively.

Compared with $Z_{in_FB}(s)$ in (4-51), $Z_{in_M}(s)$ can be re-written as $Z_{in_FB}(s)$ in series with $Z_{q_VIR}(s)$.

$$Z_{in_M}(s) = Z_{inC}(s) + \underbrace{\frac{1}{G_{denc}} \left((G_{12}G_{23} - G_{13}G_{22})G_{de1} + (G_{13}G_{32} - G_{12}G_{33})G_{de2} + G_{13} \right)}_{Z_{q_VIR}(s)} \cdot G_{q_VIR}(s) \quad (4-70)$$

Figure 4.24 shows the equivalent circuit of the modified input impedance series connected with VIR in the q channel, where $Z_{q_VIR}(s)$ is in series with $Z_{in_FB}(s)$ to generate the $Z_{in_M}(s)$.

Figure 4.24 The equivalent circuit of the input impedance with VIR in q-axis

2) The Expression of $Z_{q_VIR}(s)$

The virtual impedance in q-axis components $Z_{q_VIR}(s)$ can be obtained as

$$Z_{q_VIR}(s) = \begin{cases} \frac{G_{q_VIR}}{G_{den}} \left((G_{12}G_{23} - G_{13}G_{22})G_{de1} + (G_{13}G_{32} - G_{12}G_{33})G_{de2} + G_{13} \right) & f \in [f_1, f_2] \\ -\infty & f \notin [f_1, f_2] \end{cases} \quad (4-71)$$

3) The Expression of $G_{q_VIR}(s)$

Therefore, the realization of the piecewise function $G_{q_VIR}(s)$ can be made by the band-pass filter $G_{BPF}(s)$. The expression of $G_{q_VIR}(s)$ can be present as

$$G_{q_VIR}(s) = 2 \cdot \frac{G_{denc} |Z_{inC}|}{\left(\left((G_{12}G_{23} - G_{13}G_{22})G_{de1} + (G_{13}G_{32} - G_{12}G_{33})G_{de2} + G_{13} \right) \right)} \cdot G_{BF} \quad (4-72)$$

4.4.4 Comparison of the G_{d_VIR} and G_{q_VIR}

As discussed before, both $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$ can be treated as the VIR to shape the input impedance for stabilizing the whole cascaded system. The comparison of the $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$ is carried out here.

The bode plot of G_{d_VIR} and G_{q_VIR} is shown in Figure 4.25. As seen, there is

little difference in the frequency response between the two regulators. However, considering the D_q is much smaller in the normal case, the same impact on the q-axis is more significant than that in the d-axis. In other words, the relationship between D_d and D_q represents that D_q is more sensitive to the variation, which is a higher tendency to involve additional influence on the operation of the system.

Figure 4.25 Bode plot of $G_{d_VIR}(s)$ and $G_{q_VIR}(s)$

Therefore, it is recommendable to choose G_{d_VIR} to modify the input impedance of the VSC.

4.4.5 Robust Design for the Deviation on The LC Filter

The deviation on the LC filter in the dc side can be considered in the design of the VIR with a certain redundancy that the LC's value can vary on the dc side. The frequency margin can be included when designing the virtual impedance regulator.

Assume the value of the inductor L and the capacitor C for the input filter will be varied from the original state, f_1 and f_2 of the high-pass filter and the low-pass filter can be settled according to this change. Therefore, the frequency of a high-pass filter f_1 and a low-pass filter f_2 are as follows

$$f_1 = \frac{L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e} \quad (4-73)$$

$$f_2 = \frac{-L - \sqrt{4LCR_e^2 + L^2}}{4\pi LCR_e} \quad (4-74)$$

Then, for the LC filter, if the parameters deviated $\alpha\%$ from the original value of the inductance L , and $\beta\%$ deviated from the original value of the capacitance C , the corresponding frequency range of the band-pass filter can be

$$f_1 = \frac{L - \sqrt{4(1 + \beta\%)LCR_e^2 / (1 + \alpha\%) + L^2}}{4\pi L(1 + \beta\%)CR_e} \quad (4-75)$$

$$f_2 = -\frac{L + \sqrt{4(1 + \beta\%)LCR_e^2 / (1 + \alpha\%) + L^2}}{4\pi L(1 + \beta\%)CR_e} \quad (4-76)$$

where α and β are the deviated percentage of the inductance L and the capacitance C compared to the original value, respectively.

4.4.6 Design Example

A dc microgrid system with an input LC filter and a three-phase VSC is analyzed here to illustrate the instability problem of the studied system. Table 4-2 gives detailed parameters.

Table 4-2 Parameters of the stand-alone DGS

Parameter	Value
DC bus voltage V_{dc}	300 V
AC bus voltage	110 V (RMS)
DC filter inductance L	0.6 mH
DC filter capacitance C	100 μF
Grid fundamental frequency	50 Hz
Switching frequency	10 kHz
AC filter inductance L	6 mH
AC filter capacitance C	100 μF
Resistance Load R	12 Ω

1) Stability Analysis

A three-phase VSC in stand-alone operation with three different load types, including resistive, inductive, and capacitive load, and an input LC input filter is designed. According to the above analysis in the comparison part, G_{d_VIR} is selected as the VIR to shape the input impedance of the VSC in this design example.

The bode diagram for the output impedance of the LC filter and the modified input impedance of the VSC at different loads are shown in Figure 4.26 (a),(c), and (e). The Nyquist diagram of the impedance ratio at different loads is shown in Figure 4.26 (b), (d), and (e). As seen, the overall system is stabilized at different loads with proper design $G_{d_VIR}(s)$.

Figure 4.26 Bode plot and Nyquist curve of the proposed VIR for the POL converter: Bode plot of Z_{inC_R} , Z_{inM_R} and Z_{out} with resistive load (a), inductive load (c) and capacitive load (e); Nyquist diagram of the impedance ratio of Z_{inM_R} and Z_{out} with resistive load (b), inductive load (d) and capacitive load (f).

2) Comparison of the G_{dC} and G_{dC_Md}

As discussed before, $G_{d_VIR}(s)$ only affects v_{Ld} . To evaluate its impact, two transfer functions $G_{dC}(s)$ and $G_{dC_Md}(s)$, are defined as:

$$G_{dC}(s) = \frac{\hat{v}_{Ld}(s)}{\hat{i}_{dc}(s)} = \frac{1}{G_{denc}} \cdot \left((G_{23}G_{31} - G_{21}G_{33})G_{de1} + G_{21} \right) + (G_{21}G_{32} - G_{22}G_{31})G_{de2} \quad (4-77)$$

$$G_{dC_Md}(s) = \frac{\hat{v}_{Ld}(s)}{\hat{i}_{dc}(s)} = G_{dC}(s) + \frac{G_{d_VIR}}{G_{denc}} \cdot \left((G_{12}G_{33} - G_{13}G_{32})G_{de1} - G_{12} \right) + (G_{12}G_{23} - G_{13}G_{22})G_{de2} \quad (4-78)$$

where $G_{dC}(s)$ and $G_{dC_Md}(s)$ represent the transfer function without VIR and with VIR, respectively.

Figure 4.27 shows the bode plot of $G_{dC}(s)$ and $G_{dC_Md}(s)$. $G_{d_VIR}(s)$ only bring little influence on the VSC's dynamic performance, which ensured the dynamic response of the overall system. The magnitude of both G_{dC} and G_{dC_Md} are below zero, therefore the disturbance of i_{dc} to the v_{Ld} of the converter tends to converge.

Figure 4.27 Bode plot of $G_{dC}(s)$ and $G_{dC_Md}(s)$

3) Robust Design

The virtual impedance regulator with a certain redundancy of the variation on the LC input filter can be calculated. The corresponding parameters of the LC filter and characteristic frequencies of the band-pass filter are shown in Table 4-3.

Table 4-3 Parameters of the characteristic frequencies

	Inductance L_f	Capacitance C_f	f_1	f_2
Case 1: Initial case	0.6 mH	100 μF	538 Hz	784 Hz
Case 2: Increase 50%	0.9 mH	150 μF	358 Hz	522 Hz
Case 3: Decrease 50%	0.3 mH	50 μF	1706 Hz	1568 Hz

The bode diagram for the output impedance of the LC filter and the modified input impedance of the converter with $G_{d_VIR}(s)$ are shown in Figure 4.28. As seen, with the $G_{d_VIR}(s)$, the input impedance of the three-phase VSC has been

modified with 50% redundancy on the input filter the system is still stable.

Figure 4.28 Bode plot of the system with the proposed VIR considered the deviation on LC filter

4.5 Experimental Results

In this section, the proposed VIR stabilization strategy was verified experimentally. The system parameters are listed in Table 4-1. In this part, the proposed VIR is experimentally verified on a real ac cascaded system. The hardware prototype of the proposed cascaded system is shown in Figure 4.29.

Figure 4.29 A photograph of the main circuit of the experimental platform.

A dSPACE DS1202 (MicroLabBox) is utilized as the digital controller with its internal slave Xilinx FPGA board responsible for modulation and generating the gate signals for all the module converters. In addition, the overvoltage/overcurrent protect functions are also realized in the slave FPGA.

4.5.1 Experiments for the Stability Analysis

The schematic diagram of the experimental system is given in Figure 4.2. To evaluate the performance, the experiments are conducted under three different load cases, including under resistive load ($R = 12 \Omega$), inductive load ($R = 12 \Omega, L = 0.6\text{mH}$), and capacitive load ($R = 12 \Omega, L = 0.6\text{mH}, C = 470\mu\text{F}$) scenarios.

Figure 4.30 Instability evaluation results of the experimental system with resistive load: (a) experimental waveforms of the original system; (b) experimental waveforms of the modified system with the proposed VIR.

Figure 4.31 Instability evaluation results of the experimental system with inductive load: (a) experimental waveforms of the original system; (b) experimental waveforms of the modified system with the proposed VIR.

The dc voltage v_{inv} and the resistive load currents i_a, i_b, i_c transient response with and without the proposed VIR stabilization strategy are represented in Figure 4.30. The experimental result for the inductive and capacitive loads are also presented in Figure 4.31 and Figure 4.32. It shows that the proposed VIR stabilization strategy can effectively suppress the oscillation with v_{inv} and $i_a, i_b,$

i_c under three different loads of scenarios.

Figure 4.32 Instability evaluation results of the experimental system with capacitive load: (a) experimental waveforms of the original system; (b) experimental waveforms of the modified system with the proposed VIR.

The dynamic performance for the experimental system with VIR under resistive load is tested in Figure 4.33. The system is stable before and after the load steps from 50% to 100% full-load, and the dynamic performance is desirable.

Figure 4.33 The experiment results of the modified system with the proposed VIR when dynamic load change: from 50% full-load to 100% full-load.

4.5.2 Experiments for the Robust Design

The experimental validation considering the deviation on the LC filter of the system is given in Figure 4.34. The parameters are shown in Table 4-2.

There are three cases of different values for the LC input filter, considering increased 50% as Case 2, and decreased 50% as Case 3 on both the inductance and capacitor. The oscillation is more severe in Case 2 compared with all cases. The system can be stable even if the parameter of the EMI filter deviates from its original value.

In summary, the experimental results shown in Figure 4.30 to Figure 4.34

verified the effectiveness of the proposed VIR stabilization method. It can be seen that the proposed VIR can stabilize the system with the good dynamic performance of the VSC.

Figure 4.34 Instability problem of the studied system with a variation on LC filter: a) case 1: original(L: 0.6 mH, C: 100 μF); b) case 2: increased 50%(L: 0.9 mH, C: 150 μF); c) case 3: decreased 50%(L: 0.3 mH, C: 50 μF).

4.6 Conclusions

In this paper, the instability problem of dc-link voltage in the dc microgrid is investigated and discussed. The interaction between the input LC filter with the POL converter may degrade the dc-link voltage stability and even lead to complex oscillation problems in the whole system. Thus, to solve this instability issue, a series-connected virtual impedance regulator (VIR) is proposed to remove the negative effect of the VSC only in the intersected frequency range. Therefore, this VIR stabilization method can stabilize the dc-link voltage with maximum retention of the dynamic characteristics of the VSC. The modeling of the input impedance of the VSC, which feeds linear three-phase loads, and the deviation of the proposed VIR are carefully discussed in this paper. The experimental results verify the correctness of the proposed VIR stabilization method.

Chapter 5 Robust Dichotomy Solution-Based Model Predictive Control for the Grid-Connected Inverters with Disturbance Observer

This chapter presents a robust dichotomy solution-based model predictive control (DS-MPC) with a fixed switching frequency for the grid-connected inverter (GCI). The proposed fast dichotomy algorithm can select and deduce the optimal voltage vector dynamically through the space vector plane. Therefore, the proposed DS-MPC strategy could ensure dynamic performance and steady-state performance as well. Also, the current control robustness can be improved through DS-MPC with disturbance observer (DO) based on the extended Kalman filter (EKF). The novelty of this control is that the current control with a fast dynamic response can be realized in the weak grid, even if the grid voltages are significantly distorted.

5.1 Introduction

Grid-connected inverters (GCIs) are nowadays highly popular in various applications, such as the interface for connecting the distributed power generations with the utility grid and motor drives [73]. Hence, the topology and control strategy of the GCI is becoming significant as the increasing use in electrical systems and several interviews have been reported focusing on this field [74]. In particular, a weak grid can lead to a large impedance in the grid side, which will be overlapped in the output filter parameters. However, the detailed information of the plant model influences the performance of MPC. Meanwhile, considering the grid voltage could be unbalanced and distorted, the amplitude and frequency of the grid voltage are varied as well [57]. Therefore, the control performance could be influenced by the disturbance of equivalent grid impedance.

This chapter proposes a robust dichotomy solution-based method model predictive control (DS-MPC) based on the disturbance observer of the grid voltage through the EKF for the three-phase inverters. The fast dichotomy

solution approach is employed in the proposed DS-MPC to dynamically choose and compute the finest voltage vector for every sampling period. Using the FPGA controller, the high-frequency control can be executed in a short time. Besides, the proposed DS-MPC strategy essentially contains the basic FS-MPC and the PWM modulator with constant frequency. Therefore, compared with the GCI based on the conventional FS-MPC strategy, it can be simplified for the design of the output filter. Furthermore, with the proposed DS-MPC, the total harmonic distortion (THD) can also be reduced. Moreover, based on the disturbance observer, the model accuracy used in MPC is compensated. Hence, the proposed method for control of the inverter keeps the dynamic characteristics of FS-MPC and presents excellent steady-state performance and high robustness to disturbances of grid voltages at the same time.

5.2 Modeling and Control for the Studied Systems

5.2.1 Mathematical Model of the GCIs

The system under study in this paper is a three-phase GCI system, as displayed in Figure 5.1. The related parameters are summarized in Table 5-1. The renewable energy sources are simplified as a stiff dc voltage source, then through the GCI and the output filter, connected to the grid at the point of common coupling (PCC).

The GCI is connected with the supply voltages with the EMI filter. The general network dynamic models can be deduced as

$$\begin{bmatrix} \frac{di_a}{dt} \\ \frac{di_b}{dt} \\ \frac{di_c}{dt} \end{bmatrix} = -\frac{R}{L} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} u_{aN} - u_{ga} - u_{nN} \\ u_{bN} - u_{gb} - u_{nN} \\ u_{cN} - u_{gc} - u_{nN} \end{bmatrix} \quad (5-1)$$

where i_a, i_b, i_c are output currents of the inverter, u_{aN}, u_{bN}, u_{cN} are the inverter output voltages, u_{nN} is the voltage between the dc-link negative point and the grid voltage neutral point, u_{ga}, u_{gb}, u_{gc} are the ac-link grid voltages, respectively.

Figure 5.1 The topology of the studied three-phase grid-connected inverter.

Table 5-1 Specifications of the studied GCI

Parameter	Value
Rated power P_N	1.5 kW
Fundamental frequency f	50 Hz
Switching frequency f_s	10 kHz
Sampling Frequency	50 kHz
Filter inductance L	10 mH
Line resistance R	0.01 Ω
DC-link voltage U_{dc}	300 V
AC bus voltage U_{ga}, U_{gb}, U_{gc}	115 V (RMS)

In this thesis, the three-phase voltage is supposed to be balanced, then the dynamics equation of the current can be deduced based on the $\alpha\beta$ frame in the vector equation

$$L \frac{d\mathbf{i}_s}{dt} = \mathbf{u}_s - R\mathbf{i}_s - \mathbf{u}_{gs} \quad (5-2)$$

where $\mathbf{i}_s$ is the output current vector, $\mathbf{u}_s$ is the voltage vector and $\mathbf{u}_{gs}$ is the grid-side voltage vector.

Then, the current vector can be deduced according to the phase currents

$$\mathbf{i}_s = \frac{2}{3} (i_a + \alpha i_b + \alpha^2 i_c) \quad (5-3)$$

where $\alpha = e^{j(2\pi/3)}$. The voltage vector is determined the same

$$\mathbf{u}_s = \frac{2}{3} (u_{aN} + \alpha u_{bN} + \alpha^2 u_{cN}) \quad (5-4)$$

The voltage $\mathbf{u}_s$ is deduced based on the inverter switching state and the dc-link voltage, which can be described as

$$\mathbf{u}_s = \mathbf{S} U_{dc} \quad (5-5)$$

where U_{dc} is the dc voltage and $\mathbf{S}$ is the inverter gating signals vector which can

be determined as

$$S = \frac{2}{3}(S_a + \alpha S_b + \alpha^2 S_c) \quad (5-6)$$

where S_a, S_b, S_c are the gating signals of each inverter in Figure 5.1, and when for the value of S_x is 0 present OFF and 1 is ON, respectively ($x = a, b, c$).

Therefore, the predicted current can be derived according to the discrete-time model below

$$i_s(k+1) = \left(1 - \frac{RT_s}{L}\right)i_s(k) + \frac{T_s}{L}[\mathbf{u}_s(k) - \mathbf{u}_{gs}(k)] \quad (5-7)$$

The equation (5-7) is deduced by the discrete equation (5-2) based on a sampling period T_s . According to the forward Euler equation, the discretized model can be established based on estimating the derivative for the difference for each sampling period

$$\frac{d\mathbf{i}_s}{dt} \approx \frac{\mathbf{i}_s(k+1) - \mathbf{i}_s(k)}{T_s} \quad (5-8)$$

5.2.2 Traditional FS-MPC Strategy

In Figure 5.2 (a), the schematics of the traditional FS-MPC strategy are displayed.

Note that the entire control scheme operates in the stationary frame through Clarke's transformation. The discrete-time equation is contained in the "Predictive Model" model part. It can predict one sampling step on eight current values, which means for every eight voltage vectors in the following sampling period $\mathbf{i}_s(k+1)$ is deduced through this model. Therefore, the error compared with the predicted currents and the reference of the following sampling period can be deduced according to the cost function (5-9) in the "Minimization of cost function" model part.

$$G = |i_\alpha^* - i_\alpha(k+1)| + |i_\beta^* - i_\beta(k+1)| \quad (5-9)$$

The current error can be minimized by the selected voltage vector and implemented for each corresponding switching state.

(a)

(b)

Figure 5.2 Schematics for MPC for the GCI: a) the traditional FS-MPC; b) the proposed DS-MPC.

This is basic predictive current control strategy presents a variable switching frequency through the sampling frequency and the operating situation. The

dynamic response was chiefly selected through the sampling periods. This is because the absolute current error becomes smaller when the sampling frequency increases. However, the maximum switching frequency is restricted by the power of the electronic device in practical applications. Because of this, the sampling frequency of the controller is generally decided by the maximum value of the switching frequency. Therefore, especially with the three-phase inverters, the dynamic response of the traditional predictive current control is generally not desired compared to several traditional control strategies with the fixed switching frequency.

5.2.3 Proposed DS-MPC Strategy

Figure 5.2 (b) shows that the proposed DS-MPC strategy with the voltage observer based on EKF is deduced for the three-phase inverter. Note that the three-phase currents and voltages are transferred to the rectangular coordinate system. Then through the model predictive equations, the predicted currents are deduced. The EKF observer calculates the observed grid voltage. The details of the EKF observer are explained below. Besides, the cost functions based on the dichotomy solution are conveyed to the PWM modulator, then the optimal voltage vector can be determined compared to the reference current. Therefore, it is quasi-continuous for the selected optimal voltage vector. Finally receives the switching state of each inverter is collected, and the currents are well-regulated as anticipation.

The cost function in current control is determined according to the absolute error compared with the predicted current and the reference, which can be manifested as

$$G = |i_{\alpha}^* - i_{\alpha}(k+1)| + |i_{\beta}^* - i_{\beta}(k+1)| + g_I \quad (5-10)$$

$$g_I = \begin{cases} 0 & i_s < I_{s,\max} \\ \infty & \text{others} \end{cases}$$

where g_I is added for the overcurrent protection.

5.2.4 Challenge of DS-MPC under the Weak Grid

To realize the function optimization algorithm, the dichotomy is applied in the control strategy. The flowchart of the voltage vector selection iteration procedure

is presented in Figure 5.3. The search area for the voltage vector is shown in Figure 5.4, which is determined by a circular plane with the radius as the maximum available voltage U_m . Hence, through the internal circle of the hexagon, which is integrated by the active voltage vectors, the circle plane is deduced. The amplitude $U_m/2$ and phase $\pi/4$ are different in every switching period, the eight voltage vectors are chosen at the initialization. Then, one of the eight vectors is selected according to the cost function as the optimal vector in the initial step of the iteration. Therefore, with the updated amplitude and phase angles by half of the final step optimal vector, 18 new vectors were selected in the final step to calculate the cost function. The search area narrows to half of the last step. The optimal vector is chosen after 14 steps iteration based on the reference vector for the PWM modulator.

Figure 5.3 Optimization algorithm for the cost function.

Figure 5.4 Flowchart of the voltage vector v_i iteration procedure.

Figure 5.5 The three-phase voltage in the PCC with 5 mH grid inductance in the weak grid.

According to the characteristics of the weak grid, the grid has a large inductive impedance, and the grid voltage is heavily affected by harmonics. Due to MPC being designed through the equivalent model of the three-phase GCI, the distorted voltage cannot be considered. Therefore, the performance can be degraded with conventional MPC.

Figure 5.5 presents the simulation results of distorted grid voltage for the existence of grid impedances. In the simulation, the topology of GCI is presented

in Figure 5.1; the parameters are shown in Table 5-1. The grid system is considered very weak when $SCCR = 1$, and weak when $2 < SCCR < 3$ [75]. The equivalent grid resistance and inductance are set to 0.05Ω and 5 mH , and $SCR=2.6$, respectively. Currently, the output current of GCI in the PCC controlled by DS-MPC is shown in Figure 5.6. Because the actual voltage deviates from the voltage in the control system, the real current is distorted.

Figure 5.6. DS-MPC controls the three-phase output current of GCI in the PCC with 5 mH grid inductance variation.

5.3 Grid Voltage Estimation through the Extended Kalman Filter

5.3.1 Fundamentals of the Extended Kalman Filter

It is well known that the Kalman filter (KF) is an optimal recursive algorithm for a time-variant linear system that proposes the minimum variance state estimation. It aims to reduce the error compared with the predicted state and the actual state of a system only using the previous state. The extended Kalman filter (EKF) is the extension for the KF in the non-linear system is. The applications of EKF could usually achieve fast convergence and robustness [76], [77].

The prediction process can be deduced according to the forward Euler method

$$x_{k/k-1} = x_{k-1} + T_s \mathbf{A} x_{k-1} + T_s \mathbf{B} u_{k-1} \quad (5-11)$$

where T_s is the sampling time, x_k is the state variable, u_k is the actuating variable, $\mathbf{A}$ and $\mathbf{B}$ are the system matrix and input matrix, respectively. Besides, k is the sampling step, and $k/(k-1)$ indicates a priori estimate.

Let the transition function be $f(x)$,

$$x_{k/k-1} = x_{k-1} + T_s f(x_{k-1}, u_{k-1}) \quad (5-12)$$

The equation (5-12) is the prediction or transition equation where a new temporary state (a priori) is deduced based on the old state, the input, and the transition function. Thus, the Jacobian matrix of the plant model can be obtained as follows

$$\mathbf{J}_k = \frac{\partial (x_{k-1} + T_s f(x_{k-1}, u_{k-1}))}{\partial x_{k-1}} \quad (5-13)$$

The derived Jacobian matrix is used to deter the error covariance matrix $\mathbf{P}$ using the following equation

$$\mathbf{P}_{k/k-1} = \mathbf{J} \mathbf{P}_{k-1} \mathbf{J}^T + \mathbf{Q} \quad (5-14)$$

where $\mathbf{Q}$ is the process noise covariance matrix. Then the

Kalman gain can be calculated as

$$\mathbf{K}_k = \mathbf{P}_{k/k-1} \mathbf{H}_k^T (\mathbf{H}_k \mathbf{P}_{k/k-1} \mathbf{H}_k^T + \mathbf{R})^{-1} \quad (5-15)$$

where $\mathbf{R}$ is the measurement noise covariance matrix. The error covariance $\mathbf{P}$ is then updated using the following equation

$$\mathbf{P}_k = \mathbf{P}_{k/k-1} - \mathbf{K}_k \mathbf{H}_k \mathbf{P}_{k/k-1} \quad (5-16)$$

The Jacobian matrix of the measurement matrix is

$$\mathbf{H}_k = \frac{\partial h(x)}{\partial x} \quad (5-17)$$

where $h(x)$ is the measurement matrix.

In conclusion, the Kalman gain can be calculated based on the error covariance matrix, the Jacobian for the measurement matrix, and the measurement error covariance matrix using the equation (5-15). This Kalman

gain is applied to correct the temporary estimated state and to update the error covariance matrix through the equation (5-16). Finally, the EKF calculation process is summarized in Table 5-2.

Table 5-2 The equations of the EKF calculation process

Prediction	$x_{k/k-1} = x_{k-1} + T_s f(x_{k-1}, u_{k-1})$
Jacobian matrix	$J_k = \frac{\partial (x_{k-1} + T_s f(x_{k-1}, u_{k-1}))}{\partial x_{k-1}}$
Covariance matrix	$P_{k/k-1} = J_k P_{k-1} J_k^T + Q$
Measurement Jacobian	$H_k = \frac{\partial h(x)}{\partial x}$
Kalman gain	$K_k = P_{k/k-1} H_k^T (H_k P_{k/k-1} H_k^T + R)^{-1}$
Covariance update	$P_k = P_{k/k-1} - K_k H_k P_{k/k-1}$
Correction	$x_k = x_{k/k-1} + K_k (y_k - H_k x_{k/k-1})$

Figure 5.7 EKF algorithm flow chart.

The EKF is a recursive predictor-corrector type estimator. To demonstrate the EKF algorithm, its algorithm flow chart is presented in Figure 5.7. It can be seen that EKF mainly includes two steps of iteration. The first step is called filtering. The system's state is transitioned forward using the plant model, and at the same time, a Kalman gain is calculated. Then the forwarded state is corrected based on the Kalman gain and the measurement error. Simultaneously, the error covariance matrix is updated as well. Thus, the inputs of the system are the initial state and the initial error covariance matrix.

5.3.2 Observer Formulation Model

Figure 5.8 shows that the equivalent grid impedance model of the weak grid. In the tested system, the PCC voltages and line currents of the GCI are the measured quantities for the control algorithm. The three-phase system can be transformed into a two-phase system by the Clarke transformation.

Figure 5.8 Equivalent grid impedance system model.

The dynamic equations of the circuit in the $\alpha\beta$ reference frame can be described as

$$\frac{d}{dt} \begin{bmatrix} i_{pcc,\alpha} \\ i_{pcc,\beta} \end{bmatrix} = \begin{bmatrix} -\frac{R_g}{L_g} & 0 \\ 0 & -\frac{R_g}{L_g} \end{bmatrix} \begin{bmatrix} i_{pcc,\alpha} \\ i_{pcc,\beta} \end{bmatrix} + \begin{bmatrix} -\frac{1}{L_g} & 0 \\ 0 & -\frac{1}{L_g} \end{bmatrix} \begin{bmatrix} u_{pcc,\alpha} \\ u_{pcc,\beta} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_g} & 0 \\ 0 & \frac{1}{L_g} \end{bmatrix} \begin{bmatrix} u_{g,\alpha} \\ u_{g,\beta} \end{bmatrix} \quad (5-18)$$

$$\underbrace{\begin{bmatrix} i_{pcc,\alpha}(k+1) \\ i_{pcc,\beta}(k+1) \\ u_{pcc,\alpha}(k+1) \\ u_{pcc,\beta}(k+1) \\ u_{g,\alpha}(k+1) \\ u_{g,\beta}(k+1) \\ R_g(k+1) \\ l_g(k+1) \end{bmatrix}}_{:=x(k+1)} = \underbrace{\begin{bmatrix} \gamma & 0 & -T_s l_g & 0 & T_s l_g & 0 & 0 & 0 \\ 0 & \gamma & 0 & -T_s l_g & 0 & T_s l_g & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}}_{:=f} \underbrace{\begin{bmatrix} i_{pcc,\alpha}(k) \\ i_{pcc,\beta}(k) \\ u_{pcc,\alpha}(k) \\ u_{pcc,\beta}(k) \\ u_{g,\alpha}(k) \\ u_{g,\beta}(k) \\ R_g(k) \\ l_g(k) \end{bmatrix}}_{:=x(k)} + \underbrace{\begin{bmatrix} q_{ipcc,\alpha}(k) \\ q_{ipcc,\beta}(k) \\ q_{upcc,\alpha}(k) \\ q_{upcc,\beta}(k) \\ q_{ug,\alpha}(k) \\ q_{ug,\beta}(k) \\ q_{Rg}(k) \\ q_{lg}(k) \end{bmatrix}}_{:=f} \quad (5-19)$$

$$y(k) = \begin{bmatrix} i_{pcc,\alpha}(k) \\ i_{pcc,\beta}(k) \\ u_{pcc,\alpha}(k) \\ u_{pcc,\beta}(k) \end{bmatrix} = \underbrace{\begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 \end{bmatrix}}_{:=C} x(k) + \begin{bmatrix} r_{ipcc,\alpha}(k) \\ r_{ipcc,\beta}(k) \\ r_{upcc,\alpha}(k) \\ r_{upcc,\beta}(k) \end{bmatrix}$$

According to Taylor series expansion, the discrete EKF observer formulation can be derived as

$$\gamma = \exp\left(-\frac{R_g}{L_g} T_s\right) \approx 1 - \frac{R_g}{L_g} T_s \quad (5-20)$$

The divide operation is not preferred for the simplification of calculating the Jacobian matrix $\mathbf{J}_k$. Thus, the following substituted equation is implied

$$l_g = \frac{1}{L_g} \quad (5-21)$$

Note that it is necessary to define the state matrix initial values for EKF, including the measured noise covariance $\mathbf{R}$ and the process noise covariance matrix $\mathbf{Q}$.

However, the general method to define these values is optimized based on the online tuning due to no mathematical proof yet regarding the matrices' optimal values. The measurement error can influence the $\mathbf{R}$ matrix, so it can be calculated according to the sensors' measuring data or tuned using the trial and error method. The $\mathbf{Q}$ and $\mathbf{R}$ matrices are supposed to be diagonal to simplify the calculation and on the off-diagonal terms, no statistical data is available.

5.4 Experimental Results

In this section, the performance of the proposed DS-MPC with a grid voltage observer with EKF through a real GCI system test platform is evaluated, as shown in Figure 5.9. The FPGA-based controller consists of an A/D converter, PWM interface board, and Altera DE2-115 development board.

The control strategy is utilized Through the FPGA-based controller. Then according to CORDIC (Coordinate Rotation Digital Computer) algorithm, complex trigonometric and quadratic functions are computed. Note that the clock frequency is 80 MHz and every voltage vector assessment demands a 40-step pipeline, and this procedure requires 0.5 μS . Then 18 voltage vectors are assessed for every iteration, and the iteration step is arranged to 14 in this experiment. Hence, 10.15 μS is needed for the proposed method in this entire design. For the implementation of EKF based grid voltage observer, there are many matrix operations, which means many multipliers need to be used. However, the resource of the multiplier in FPGA is limited. So, many FPGA optimization methods should be considered. The resource sharing technique is used to decrease the utilization of multipliers. Meanwhile, all the data type in EKF is the same which is realized by parameter normalization. The bit width of

all parameters in EKF is 18-bit; the lower 14 bits are for the fraction part and the most significant bit (MSB) is for the sign. In Table 5-3, the resource employment of FPGA EP4CE115F29C7 is depicted.

Figure 5.9 Experimental testbench.

Table 5-3 FPGA resources employment of proposed DS-MPC

Resource	Amount Used/Total Available
PLLs	1/4 (25%)
Embedded Multiplier 9-bit elements	234/532 (44%)
Memory bits	292,872/3,981,312 (8%)
Pins	61/529 (12%)
Dedicated logic registers	14,056/114,480 (12%)
Logic elements	28,468/114,480 (25%)
Combinational functions	28,036/114,480 (25%)

Hence, the performance is evaluated according to a three-phase GCI with a 1.5 kW power level. The conventional FS-MPC, the traditional linear strategies, and the proposed DS-MPC are verified and compared through the experimental testbench. Note that the conventional FS-MPC sampling frequency is 40 kHz to ensure the highest switching frequency is 20 kHz. This switching frequency is usually the highest switching frequency for IGBT in industrial applications. For the proposed DS-MPC and the traditional linear method, the PWM switching frequency is 10 kHz.

The steady-state and dynamic response of the traditional FS-MPC, the proposed DS-MPC, and the linear method are also verified based on the active current step test. In Figure 5.10 (a), (b), and (c), the active is 3 A initially. Then, the active current is claimed up to 5 A. The experimental results indicated that all control strategies could achieve satisfying dynamic performance. However, compared to the linear methods and conventional FS-MPC method, the

experimental results from the DS-MPC method have a less current ripple. The results demonstrate the high-quality dynamic performance of the proposed DS-MPC method as, unlike those for the closed-loop PI control scheme, less ripple

(a)

(b)

(c)

Figure 5.10 Experimental results of the GCI: (a) With the FS-MPC method, (b) with the proposed DS-MPC method, (c) with the traditional linear method.

The current THD (total harmonics distortion) of the proposed DS-MPC is 2.74%, which is much less than 4.12% for conventional FS-MPC and 3.26% for the linear method, as shown in Figure 5.11. Moreover, the control of DS-MPC is the more straightforward design of the filter to diminish the harmonic for the harmonics regularly spread around the switching frequency, compared with the traditional FS-MPC, with which more current harmonics exist. This can be explained by the low bandwidth nature of the main PI-based current controllers.

Since the grid interface is an isolation transformer, the grid is fundamentally a weak grid and the grid voltage is to some extent also distorted by the filter resonance. However, upon the activation of the cost function, the resonance can be suppressed.

(a)

(b)

Figure 5.11 The PCC's one-phase output voltage and current performance with 5 mH equivalent grid impedance in the weak grid: (a) Voltage and current waveforms without disturbance observer, (b) Voltage and current waveforms with EKF grid voltage observer.

In Figure 5.11, the experimental results of one-phase output voltage and current performance in the PCC with 5 mH equivalent grid impedance in the weak grid are presented. When the measured distorted grid voltages are directly used in the DS-MPC, the output currents are also distorted. The reason is that

the selected optimal voltage vector cannot represent the real distorted voltage vector precisely. In Figure 5.11 (b), after using the DS-MPC switching over the EKF observer, the ac voltage without distortion could be obtained. Therefore, the robustness of the current control is verified under the grid voltage disturbance.

(a)

(b)

Figure 5.12 Experiments of the proposed DS-MPC strategy switching over the EKF observer. (a) The ac current, (b) The ac voltage.

The dynamic performance with the DS-MPC switching over the EKF observer is shown in Figure 5.12. Under a weak grid scenario, EKF can still quickly converge on the voltage observer with high robustness. It can be shown in Figure 5.12 that once the EKF observer is applied, the output current rapidly recurs to the sinusoidal waveforms.

Therefore, the proposed robust DS-MPC strategy can ensure the dynamic performance of the current control for the GCI in the weak grid.

5.5 Conclusions

This chapter proposes a robust DS-MPC with an EKF observer for grid

voltage for the three-phase inverter system with the output L filter.

With the voltage vector dichotomy solution method, the ideal voltage vector could be dynamically chosen in a short period through the space vector plane. Therefore, the PWM modulator can be combined based on the output voltage vector with quasi-continuous characteristics. Because the THD spread in a specific range with the fixed switching frequency, the GCIs' output current filter design can be simplified than the traditional FS-MPC. Moreover, the proposed control method shows fast convergence and high robustness to grid voltage disturbance. Based on the well-designed and turned EKF observer, the grid voltage can be estimated precisely. The control algorithm is implemented based on an FPGA controller through hardware optimization methods such as pipeline and resource sharing. The high-performance current control with the weak grid can be realized with the proposed control method. The simulation and experimental results under the weak grid case based on the real prototype verified the effectiveness of the proposed method.

Chapter 6 Conclusion and Recommendations

This chapter provides further research that will be developed next, apart from work presented in this thesis.

6.1 Conclusions

In this thesis, existing stabilization methods for the power conversion system have been reviewed in detail. Although these stabilization methods can stabilize the system, there are several drawbacks. For example, most of the need trade-off with the stability and the dynamic performance of the cascaded system, they did not consider the detailed model of the power conversion system, and so on.

The proposed VIR control techniques will improve the stability of the power conversion system in both grid-connected and stand-alone operations. The modeling of the grid-connected converter with the output L filter and its equivalent circuit is deduced. Afterward, a control strategy by VIR to enhance the stability of the interlinking converter is presented. Simulation and experimental results have verified the effectiveness of the control strategy. This is a new approach that has never been investigated. Hence, the future course of action will revolve around the impedance-based method of such modulation techniques to further enhance the performance of the power conversion system.

Then, the VIR control strategies are applied for the dc-link voltage instability problem in the dc microgrid. The modeling of the converter with the stand-alone operation and its equivalent circuit is described. According to the voltage and current double-loop control, VIR is series connected to enhance the system's stability. This method is established based on the equivalent input impedance of the three-phase converter. There is a small range of frequency modified to minimize the influence on the system's dynamic performance.

With the voltage vector dichotomy solution method, the ideal voltage vector could be dynamically chosen in a short period through the space vector plane. Therefore, the PWM modulator can be combined based on the output voltage vector with quasi-continuous characteristics. Because the THD spread in a

specific range with the fixed switching frequency, the design of the GCIs' output current filter can be simplified than the traditional FS-MPC. Moreover, the proposed control method shows fast convergence and high robustness to grid voltage disturbance. Based on the well-designed and turned EKF observer, the grid voltage can be estimated precisely. The control algorithm is implemented based on an FPGA controller through hardware optimization methods such as pipeline and resource sharing. Thus, the high-performance current control with the weak grid can be realized with the proposed control method.

Therefore, the stabilization methods proposed in this thesis overcame the drawbacks of the existing solutions and provided the optimal methods. Simulation results are provided based on MATLAB/Simulink environment to validate all the performance and robustness of the VIR and FCS-MPC method. The experiments are also applied to prove the proposed methods' performance and robustness, which could emulate delays and errors.

6.2 Future Work

As discussed above, many impedance-based methods have been proposed to stabilize the power conversion system, which can be classified into two types: passive methods [17] and active methods [78]–[80]. However, though these stabilization methods can stabilize the cascaded system, they have unavoidable drawbacks, e.g., they do not have enough load stabilization methods, they degrade the original performance of the power conversion system, and so on. To overcome these drawbacks, a family of impedance-based stabilization methods will be proposed in future work.

Until now, the research focus has been on the power conversion system with active damping methods. The future is to investigate and improve the performance of three-phase converters cascaded with advanced passive filters, and multiple-inputs and multiple-outputs (MIMO) systems.

To further stabilize the power conversion system, several specific research directions should be developed.

1. Applying the stability control strategies into the power conversion system

cascaded with advanced passive output filters in the ac side, i.e., the integrated *LCL* filter [81]–[83], the integrated *LLCL* filter [84], series-parallel-resonant *LCL* (*SPRLCL*) filter [85], and *LT-C-L* filter [86], which can provide high attenuation and robustness. The study on the detailed model with these advanced passive filters is limited. The impedance model and the stability criteria need further discussion to apply this method.

2. Establishing the detailed model and applying the stability control strategies into the multiple-inputs and multiple-outputs (MIMO) systems. With the development of the microgrid, MIMO systems are gaining increased attention to supply high-quality and reliable power [87]. However, MIMO systems present complex dynamics due to their structure and several sources behaving as CPL [88]–[90]. To ensure the stability and the dynamic performance of the MIMO systems, future research efforts should focus on realizing these virtual impedances on the MIMO systems.
3. Establishing the detailed model and applying the stability control strategies into the multilevel converter system. The multilevel converter system attracts growing attention because it can provide a series voltage sharing of power devices and low voltage stress.[91]. Therefore, this has been widely applied in the field of high-power transformers.[92] However, multilevel converter systems present complex dynamics due to their structure and operation. Future research efforts should focus on realizing these stability control to enhance the stability and the dynamic performance of the multilevel converter systems.
4. The adaptive stability control strategy for the power conversion system can be further discussed. In [42], the adaptive stability control is proposed for a single impedance interaction frequency range. The detailed model is hard to develop for large-scale power systems, and there will be multiple interaction frequency ranges [89]. Hence, to improve the power quality and eliminate the harmonics, the adaptive stability control strategy is needed for the self-organizing microgrid clusters in the future.
5. The online stability assessment of the power conversion system is critical considering the increasing practical power system complexity [93]. The fast and accurate assessment methods can diagnosis the operation state of the power conversion system before the oscillation or even collapse of the

voltage [94], [95]. Few studies have explored the real-time implementation of the online stability assessment of large-scale power systems. This can be the future study to enhance the stability and safety of the power system.

List of Publications

Journal Papers:

- [1] **Huanyue Liao**, Xin Zhang*, Hao Ma, "Impedance-Shaping Based Stability Control of Point-of-Load Converter Integrated with EMI Filter in DC Microgrids," *IEEE Access*, 2021, early access.
- [2] **Huanyue Liao**, Xin Zhang*, Zhixun Ma, "Robust dichotomy solution-based model predictive control for the grid-connected inverters with disturbance observer," in *CES Transactions on Electrical Machines and Systems*, vol. 5, no. 2, pp. 81-89, June 2021.
- [3] **Huanyue Liao**, Wenjian Cai*, Fanyong Cheng, "Automated Fault Diagnosis in Air Handling Units Through Rule and Convolutional Neural Networks," *Sensors*, vol. 21, no. 13, p. 4358, Jun. 2021.
- [4] Fanyong Cheng, Wenjian Cai*, Xin Zhang, **Huanyue Liao**, Can Cui, "Fault detection and diagnosis for Air Handling Unit based on multiscale convolutional neural networks", *Energy and Buildings*, vol. 236, p. 110795, Apr. 2021.

Conference Papers:

- [1] **Huanyue. Liao** and Xin. Zhang, "Stabilization of Inverter–Based Distributed Generation System via Virtual Impedance Regulator," in *2019 IEEE Energy Conversion Congress and Exposition (ECCE)*, USA, 2019, pp. 609–615.
- [2] **Huanyue. Liao**, Xin. Zhang*, Jinsong. He, "Virtual Impedance Regulator for a Three-Phase Inverter of Stand-Alone Distributed Generation System," in *2020 IEEE 9th International Power Electronics and Motion Control Conference (IPEMC2020-ECCE Asia)*, 2020, pp. 417-421.
- [3] **Huanyue. Liao**, X. Zhang*, and Xiaochao. Hou, "A Decentralized Control of Series-Connected PV-ES Inverters with MPPT and Virtual Inertia Functionality", in *2020 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2020, pp. 3221-3224.

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