

POLYTECHNIQUE MONTRÉAL

affiliée à l'Université de Montréal

**Robust control of voltage source converters under weak and unbalanced grid
conditions**

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Thèse présentée en vue de l'obtention du diplôme de *Philosophiæ Doctor*
Génie électrique

Novembre 2022

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Cette thèse intitulée :

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conditions**

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en vue de l'obtention du diplôme de *Philosophiæ Doctor*
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DEDICATION

*This thesis is dedicated to my family
for their endless love, support and encouragement. . . .*

*Cette thèse est dédié à ma famille
pour leur amour, leur soutien et leurs encouragements sans fin... .*

ACKNOWLEDGEMENTS

I would like to express my full gratitude to Professor Houshang Karimi for his invaluable advice and guidance throughout my Ph.D. program. I would also like to thank Professor Masoud Karimi Ghartemani, whose expertise and support were invaluable in completing my Ph.D. program. I am also greatly thankful to Professor Keyhan Sheshyekani for his support and accepting the supervisory of this research. Many thanks to Professor Jean Mahseredjian, Professor Tarek Ould-Bachir, and Professor Shamsodin Taheri for their careful review of this thesis and for attending my defence.

RÉSUMÉ

La prolifération des énergies renouvelables a conduit à une augmentation de l'utilisation des convertisseurs basés sur l'électronique de puissance intégrés au réseau. En raison d'avantages tels qu'une fiabilité élevée, une dynamique rapide et une faible distorsion harmonique, les convertisseurs de source de tension (VSC) sont le premier choix pour les applications d'interface réseau parmi d'autres topologies. L'interfaçage des VSC contrôlés par vecteur dans des systèmes de réseau faibles peut cependant entraîner une dégradation des performances et des problèmes de stabilité. Ceci est attribué aux interactions entre les boucles de contrôle et la boucle à verrouillage de phase (PLL) qui est couramment utilisée dans les méthodes de contrôle de courant vectoriel (VCC). Récemment, ce problème a attiré plus d'attention en raison de la croissance rapide des ressources énergétiques distribuées (DER) installées dans des endroits éloignés avec des interconnexions de réseau à très haute impédance. Pour tenter de résoudre ce problème, certaines publications ont suggéré des méthodes sans PLL où le VCC est mis en œuvre à l'aide d'une PLL en boucle ouverte. Cependant, l'exactitude d'une telle affirmation n'est pas claire.

Ainsi, le premier objectif de cette thèse porte sur l'évaluation de la méthode sans PLL et la comparaison de ses performances avec les méthodes basées sur PLL dans des conditions de réseau anormales et faibles. Dans cette thèse, il est montré que la dynamique des filtres passe-bande utilisés dans l'approche PLL-Less est très similaire à celle d'une PLL et donc, de telles méthodes ne peuvent pas offrir des performances améliorées dans des conditions de réseau faible.

Par conséquent, le problème de la dégradation de la stabilité des VSC dans des conditions de réseau faible reste à réfléchir, ce qui est le deuxième objectif de cette thèse. La dynamique d'une PLL améliorée (ePLL) est intégrée dans les équations caractéristiques d'un système de réseau faible en référentiel fixe pour obtenir un modèle linéaire invariant dans le temps (LTI). Un contrôleur multivariable robuste est ensuite appliqué sur le modèle augmenté qui inclut des gains de rétroaction sur les variables d'état du système ainsi que sur les états de l'ePLL. Mathématiquement, il est démontré que le contrôleur proposé a une marge de stabilité plus élevée dans des conditions de réseau faible, puis ses performances améliorées sont validées à l'aide de simulations logicielles et de tests de matériel-dans-le-boucle (HIL).

Dans des conditions de réseau déséquilibré, la tension continue du VSC est polluée par une oscillation de second ordre qui est induite par la puissance oscillatoire aux bornes du VSC. Bien que ce problème ait été bien étudié dans le référentiel synchrone (SRF) par la mise en

œuvre de contrôleurs basés sur la résonance sur la tension cc, il est resté une question ouverte pour le référentiel fixe. L'équation d'équilibre de puissance de VSC devient une équation non linéaire variant dans le temps dans le référentiel fixe, ce qui complique à son tour la mise en œuvre des contrôleurs à base de résonance en boucle fermée. S'attaquer à ce problème est le troisième objectif de cette thèse où un contrôleur multivariable robuste est proposé qui élimine complètement les oscillations de tension de liaison courant continu même sous de graves défauts déséquilibrés.

Ainsi, le résultat de cette recherche est un contrôleur qui non seulement améliore les performances et la stabilité du système dans des conditions de réseau faible, mais fournit également un contrôle de tension cc amélioré dans des conditions de réseau déséquilibré.

ABSTRACT

The proliferation of renewable energies has led to a surge in use of grid-integrated power electronics-based converters. Due to advantages such as high reliability, fast dynamics, and low harmonic distortion, voltage source converters (VSCs) are the foremost choice for grid-interfacing applications among other topologies. The interfacing of vector-controlled VSCs into weak grid systems however, can cause performance degradation and stability issues. This is attributed to the interactions between the control loops and the phase-locked loop (PLL) which is commonly employed in the vector current control (VCC) methods. Recently, this problem has gained more attention due to the rapid growth of distributed energy resources (DERs) installed in remote locations with very high impedance grid interconnections. Trying to address this problem, some publications have suggested PLL-less methods where the VCC is implemented using an open-loop PLL. However, the accuracy of such a claim is not clear.

Thus, the first objective of this thesis deals with evaluating the PLL-less method and comparing its performance with the PLL-based methods under abnormal, as well as weak grid conditions. In this thesis it is shown that the dynamics of the band-pass filters employed in the PLL-Less approach are much similar to those of a PLL and thus, such methods can not offer improved performance under weak grid conditions.

Hence the problem of stability degradation for VSCs under weak grid conditions remains to be pondered, which is the second objective of this thesis. The dynamics of an enhanced PLL (ePLL) are integrated into the characteristic equations of a weak grid system in stationary frame to obtain a linear time-invariant (LTI) model. A robust multivariable controller is then applied on the augmented model which includes feedback gains on system state variables as well as the states of the ePLL. Mathematically, it is shown that the proposed controller has a higher stability margin under weak grid conditions and then its improved performance is validated using software simulations and hardware-in-the-loop (HIL) tests.

Under unbalanced grid conditions, the dc voltage of the VSC becomes polluted with second-order oscillation which is induced by oscillatory power at the terminals of the VSC. Although this problem has been well studied in synchronous reference frame (SRF) by implementation of resonant-based controllers on dc-link voltage, it has remained an open question for stationary reference frame. The dc-link power balance equation of the VSC becomes a non-linear time varying one in stationary frame, which in turn complicates implementation of the closed-loop resonant-based controllers. Tackling this problem is the third objective of this thesis where a robust multivariable controller is proposed which completely eliminates the

dc-link voltage oscillations even under severe unbalanced faults.

Thus, the outcome of this research is a controller which not only improves system performance and stability under weak grid conditions but also provides enhanced dc voltage control under unbalanced grid conditions.

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LIST OF SYMBOLS AND ABBREVIATIONS

BPF	Band-Pass Filter
BTB	Back To Back
CSC	Current Source Converter
DER	Distributed Energy Resources
DFIG	Doubly-Fed Induction Generator
DG	Distributed Generation
DPC	Direct Power Control
DSO	Digital Storage Oscilloscope
DSP	Digital Signal Processor
ePLL	Enhanced Phase-Locked Loop
GFL	Grid Following
GFM	Grid Forming
GPIO	General Purpose Input/Output
HIL	Hardware-in-the-Loop
LQR	Linear Quadratic Regulator
LTI	Linear Time-Invariant
MIMO	Multiple Input-Multiple Output
PCC	Point of Common Coupling
PI	Proportional Integral
PLL	Phase-Locked Loop
PV	Photovoltaics
SCR	Short Circuit Ratio
SISO	Single Input-Single Output
SRF	Synchronous Reference Frame
THD	Total Harmonic Distortion
VCC	Vector Current Control
VSC	Voltage Source Converter
VUF	Voltage Unbalance Factor

CHAPTER 1 INTRODUCTION

1.1 Context and Motivations

During past few decades, integration of distributed generations (DGs) into the power system has faced a rapid growth as the renewable energy market has been flourishing, particularly at distribution level, where DGs are increasingly utilized [8]. By encompassing various forms of generation units, microgrids provide the basis for integration of DGs into the power grid.

Such application usually rely on power electronics-based converters to convert and exchange power between the ac-grid and the dc-bus. Owing to characteristics such as high reliability, fast dynamics and low harmonic distortion, voltage source converters (VSCs) are a feasible solution for such power conversion applications [9–16].

The control of VSCs is usually implemented in synchronous reference frame (SRF) using a phase-locked loop (PLL), where frequency and phase angle of the voltage is estimated at point of common coupling (PCC) and are used to transform ac system variables into dc values. A VSC-based DER unit can operate in grid-connected or in the islanded mode of operation [17]. In the grid-connected mode, the main objective is to control the exchange of real and reactive powers while withstanding the extreme grid and source conditions and disturbances.

Real and reactive power control in a VSC-based grid-connected DER unit is normally carried out based on the well-known vector current control (VCC) in the synchronous reference frame (SRF), where two separate feedback loops are closed on the direct and quadrature (dq)-currents using two proportional-integral (PI) compensators [2]. To perform the transformations between abc-frame and SRF, an SRF-based PLL is commonly employed [18, 19] which extracts the phase angle of the PCC voltage. The interactions between control loops and PLL however, can adversely affect the performance of the overall system when the inductance between the PCC and the grid becomes larger in size (a condition also known as weak grid) [20, 21]. Recently, this problem has been in the spotlight on account of increasing numbers of grid-integrated DGs in remote areas with extremely weak grid interconnections [22].

To address this problem, different approaches are suggested and studied in the literature which can be categorized as follows

- Reducing the Bandwidth of the PLL (e.g., [23, 24]).
- Impedance reshaping approaches (e.g., [25–28]).

- PLL-Less methods, also known as direct power control (DPC) methods (e.g., [29–31]).
- Integrating the PLL dynamics within the control system (e.g., [6, 32, 33]).

While reducing the PLL bandwidth improves system stability, it limits the system dynamics and may violate performance criteria. Impedance reshaping approaches usually are developed based on the linearized models of the system and commonly deploy the grid impedance value in their feed-forward structures which imposes two problems: i) the lack of knowledge on the precise value of grid inductance which may lead to steady-state errors and ii) the fact that such linearized model are developed within a specific range of system parameters and set-points. Thus, performance accuracy could not be guaranteed if system set-points are substantially changed.

The PLL-less methods are recently introduced in an effort to improve the weak-grid performance of the conventional VCCs by using some band-pass filters instead of a PLL. However, despite the popularity of such approaches, the accuracy of their claim is unclear, specially considering the fact that some enhanced PLLs (e.g., the ePLL introduced in [3]) have similar structure of a band-pass filter.

Integration of PLL dynamics into the system model is another interesting approach which has been recently studied and has shown promising results [32, 33]. Such approaches, which are implemented in SRF, however require linearization around a specific operating point which may impose some limitation on system dynamic performance. Furthermore, the method presented in [32] is developed for single phase systems and thus, not suitable for three-phase VSCs, unless separate PLLs are used with each phase voltage, which is not a common practice. While most of the works in the literature belong to the SRF-based controllers, only a few have been conducted on investigating the impact of PLL on current controllers in stationary frame (also known as $\alpha\beta$ -frame). In [34–36], however, it is shown that resonant-based current controllers implemented in stationary reference frame are less prone to stability degradation caused by PLL dynamics under weak grid conditions. This, in particular, calls for conducting more research on development of enhanced current controllers in stationary frame to improve system performance under weak grid conditions. Such resonant-based current controllers can also benefit from robust ePLLs implemented in stationary frame. For instance, the ePLL presented in [3], not only has a robust performance but also results in an linear time-invariant (LTI) model in stationary-reference frame and thus, integration of ePLL dynamics into the grid-connected VSC model becomes much simpler in comparison with SRF-based control approaches. This, in turn, can further improve system performance under weak grid conditions.

On the other hand, some control objectives are more difficult to achieve in stationary reference frame, e.g., dc-voltage control is particularly challenging under unbalanced grid conditions considering the time varying dynamics of the dc-link. While the mean-value control of dc-link voltage can be achieved in stationary reference frame using PI controllers in cascaded multi-loop structures, elimination of second-order harmonics on the dc voltage is far more complicated.

The dc-link voltage oscillations in a VSC reduce the life-span of dc-capacitors, affect the stability margin of the control system and deteriorate the reliability of the overall power conversion mechanism [37]. Although selecting the dc-link capacitors with larger capacitance improves the dc-link voltage ripples to some degree, it does so at the price of increased total expense of the system. There are different methods found in the literature, trying to address this problem, which are dominantly implemented in SRF. This shed light on an existing gap in the current literature where dc-link voltage control under unbalanced grid conditions has not been fully addressed in stationary reference frame. Achieving such control tasks in stationary reference frame, however, could be specially beneficial for VSCs under weak grid conditions since the PLL has a smaller impact on the resonant-based current controllers [34–36].

1.2 Thesis Objectives

As discussed in the previous section, several gaps are identified in the literature which require deeper evaluation. Consequently, the objectives of this research are shaped based on these identified gaps and are summarized as follows:

1. Rigorous evaluation of PLL-less VCCs and comprehensive comparison between these methods with the PLL-based VCCs.
2. presenting a comprehensive PLL-integrated model of a grid-connected VSC in stationary reference frame in order to: i) analyze the impact of PLL dynamics on the closed-loop system stability, and ii) proposing an enhanced multivariable control scheme which incorporates PLL state variables in its state-feedback mechanism and improves system performance and stability under weak grid conditions.
3. Implementation of dc voltage resonant control within the proposed PLL-integrated control scheme to not only improve system performance under weak grid condition but also under system voltage imbalance.

1.3 Existing Work

In this section an overview of the latest studies in enhanced performance of VSCs connected to weak and unbalanced grid is provided.

1.3.1 PLL-less Control Method

In a VSC-based grid-connected DER unit, real and reactive power control is typically implemented using the well-known vector current control (VCC) in the SRF, where two separate feedback loops are closed on the direct and quadrature (dq)-currents using two proportional-integral (PI) compensators [2]. A SRF-based phase-locked loop (PLL) is commonly used to carry out the transformations between the abc-frame and SRF. However, it is widely known that when the point of grid connection is weak, the dynamics of PLL can negatively impact the performance of the overall system [20, 21]. As a result, the concept of a PLL-less VCC, also known as the direct power control (DPC) in some publications, has been put forth in a number of recently published papers [4, 38, 39]. In such publications, the goal is to produce the synchronisation signals directly from the grid voltage without employing a PLL and only using a simple filtering stage; e.g., In [39] this is accomplished by using two conventional second-order band-pass filters (BPFs). In [40], the idea of PLL-less VCC is extended to a back-to-back (BTB) converter in the doubly fed induction generator (DFIG). However, the proposed method in [40] employs the same BPFs when the DFIG is connected to a weak grid. Further, to enable PLL-less VCC to operate under voltage imbalance, a parallel compensator is introduced in [41], which controls the negative-sequence current through a closed-loop structure. This approach however, requires a technique to extract the negative-sequence components.

1.3.2 Enhanced Performance of VSC under Weak Grid Conditions

The relationship between PLL dynamics and closed-loop system instability has been well studied in the literature. The work in [22] demonstrates that the ac-bus voltage control is highly correlated with PLL dynamics. This problem exacerbates as the system becomes weaker and degrades the stability of the VSC at lower short circuit ratios (SCR). According to the study in [36], PLL affects the output impedance of grid-tied inverters for both synchronous and stationary (i.e., dq and $\alpha\beta$) reference frames. It also shows that VCCs implemented in $\alpha\beta$ -reference frame have less significant negative damping than those realised in SRF.

Due to implementation of renewable energies in remote areas, however, connection of VSCs to weak and extremely weak grids is inevitable. In [42], it is recommended that PLL gains

be decreased in order to boost VSC's power capacity. However, the extent to which the PLL gains can be decreased is constrained by the dynamic performance criteria. Supplementary active compensators to reduce the negative effects of PLL in a weak grid-connected current source converter (CSC) is proposed in [43]. This method although provides a wider stable operation area for the system, does not consider the nonlinear nature of the PLL and its impact on system behaviour.

In [23], an impedance reshaping technique based on double-PLLs is suggested. However, the bandwidth of the inner current control loop in [23] is assumed to be infinite in order to study the destabilising factors, making the improvements only effective for the outer control loops. A frequency coupling suppression control approach is provided in [44] to enhance the performance of single-phase grid-tied inverters and improve the stability under weak and distorted grid conditions. This approach however is only applicable for single-phase systems. Control of a grid-connected photovoltaic (PV) power plant in stationary reference frame without a PLL is discussed in [45]. However, the resonance controllers deployed in [45] only attenuate around a particular frequency which limits the system to operate only very close to its nominal frequency. Furthermore, to extract voltage symmetrical components, some filtering stages are deployed which can introduce nonlinear dynamics to the system.

In [32] the dynamics of an enhanced PLL (ePLL) [46], are included into the single-phase inverter's extended current control loop. As a result, the converter stability margins have improved under weak grid conditions. However, this approach cannot be readily extended to a three-phase system, unless two independent single-phase PLLs are used for the α and β axes, which is not a common practice. In [33] the linearized model of an enhanced PLL (ePLL) [46] is integrated into the system model in synchronous frame and then the closed-loop system is shaped using an optimized multivariable feedback approach. The PLL-integrated model in [33], however, entails some nonlinear terms which in turn requires linearization of system equation. As a result, the dynamic response of the optimally designed controller might be affected by changes in system operating points.

1.3.3 Enhanced Performance of VSC under Unbalanced Conditions

Although resonant-based controllers generally perform better under weak grid situations, they are less frequently used in applications where dc voltage control is needed in the presence of voltage imbalance. The reason for this, is that since realization of dc-link dynamics in stationary frame produces a nonlinear time-varying term, it cannot be employed directly with linear control strategies. In order to achieve dc-voltage control under voltage imbalance, multi-loop nested control structures with voltage and current sequence decomposition

are extensively researched in the literature [45, 47, 48]. These techniques, however, require sequential component extractors and notch filters, which complicate the control algorithm, introduce nonlinear dynamics to the system, and degrade system stability in weak grid conditions.

In an effort to obtain imbalanced power regulation in a single stationary frame, [49] suggests employing a vector quadrature component (i.e., a variable vector delayed by 90° electrical degrees) to calculate current reference signals. The similar strategy has been used in [50, 51], which employs voltage quadrature components for the implementation of DPC techniques under voltage imbalanced. Despite their enhanced performance however, these techniques still involve control effort (i.e., rectifier output voltages) for obtaining current reference signals, which require precise estimation of system parameters and may result in steady state errors if not done so. All of these techniques also involve quite complicated structures that acquire reference currents by resolving matrix equations. When the voltages are severely distorted under poor grid conditions, this may in particular result in errors and cause stability issues. The relationship between current asymmetry and instantaneous power components in a VSC is examined in [52], where the power reference signals are purposefully oscillated to produce sinusoidal currents for various control objectives. The method in [52] merely regulates the average component of real power and thus, is unable to completely compensate the oscillations in the dc-link voltage.

1.4 Thesis contribution

The main contributions of the thesis are summarized below.

1. Comparison between PLL-less and PLL-based VCC is carried out and it is shown that the PLL-based method is almost, always, the preferred choice.
2. By incorporating the dynamics of an enhanced PLL, an LTI model of the system is developed which facilitates the design of linear controllers in grid following VSCs.
3. By proposing a new multivariable controller, the performance of the VSCs under weak grid conditions has been significantly improved.
4. The closed-loop control of dc-link voltage resonant is achieved for the first time.
5. The performance of VSCs under weak and unbalanced conditions has been significantly improved.

The results of this research has been published in two IEEE journals [1, 6].

1.5 Thesis Structure

The performance of the PLL-based VCC and the PLL-free (or PLL-less) VCC introduced as DPC in [4, 38, 39] are thoroughly evaluated and investigated in Chapter 2 of this thesis. The PLL-less approach is first revisited and it is shown that in fact, this method has the same structure as the PLL-based VCC but employs a simple form of synchronization unit accompanied with some filtering stage. In this chapter it is shown that there are several major issues with the PLL-less VCC and subsequently, an approach is introduced to overcome some of these shortcomings.

Chapter 3 presents a new PLL-integrated model of the VSC in stationary reference frame. Based on this model, a multivariable controller with optimally designed gains is developed which incorporates feedbacks from PLL state variables in its structure and thus, improves the performance and stability of the system under weak grid conditions. In order to show the effectiveness of the proposed method, stability analysis, simulation case studies as well as experimental tests are provided which validate the elevated performance of the proposed controller. Moreover, the concept of short-circuit ratio is revisited and the relationship between reactive power and voltage stability is revealed in this chapter.

Chapter 4, presents a new controller in stationary frame for enhanced performance of VSCs under weak and unbalanced grid conditions. The proposed controller incorporates a nested multi-loop structure consisting of an inner phase-locked loop (PLL)-integrated current controller as well as an outer dc-link voltage control loop based on integral-resonant control mechanism. The presented controller neutralizes second-order oscillations of the dc-link voltage as well as third-order harmonics in the output currents under voltage imbalance while improving system performance and stability under weak grid conditions.

An application of the proposed controller is presented in Chapter 5 where the ac-bus voltage is controlled under weak grid conditions. The proposed multivariable controller in Chapter 3 is employed as the inner current control loop while the ac-bus voltage is adjusted through a nested control loop. In Chapter 6, the summary of this thesis, contributions, and recommendations for further research are presented.

CHAPTER 2 EVALUATION OF PLL-LESS VCC

2.1 Background

Historically, the control of VSCs is implemented in synchronous reference frame (SRF) using a phase-locked loop (PLL), where frequency and phase angle of the voltage is estimated at point of common coupling (PCC) and are used to transform ac system variables into dc values [1]. Despite the simplicity of SRF-based control methods, thriving electricity market and high penetration of electronically-interfaced distributed energy resources (DERs) have led to a steady growth of interest in voltage-source converters (VSCs) [1, 9–13]. Due to advantages such as high reliability, fast dynamics and low harmonic distortion, the VSCs are the popular choice for electronically interfacing the DERs, e.g., photovoltaic (PV) and wind turbine generators [14–16]. A VSC-based DER unit can operate in grid-connected and islanded mode of operation [17]. Controlling the exchange of real and reactive powers while enduring the harsh grid and source conditions and disruptions is the primary goal in the grid-connected mode.

In a VSC-based grid-connected DER unit, real and reactive power control is typically implemented using the well-known vector current control (VCC) in the synchronous reference frame (SRF), where two separate feedback loops are deployed on the direct and quadrature (dq)-currents using two proportional-integral (PI) compensators [2]. A SRF-based phase-locked loop (PLL) is frequently used to carry out the transformations between the abc-frame and SRF [18, 19]. It is commonly acknowledged that when the point of grid connection is weak, the dynamics of PLL can negatively impact the performance and stability of the entire system [20, 21].

Consequently, the PLL-less VCC, also known as the direct power control (DPC), has been proposed in a number of recent promoted papers, including [4, 38, 39]. Their objective is to produce the synchronisation signals directly from the grid voltage without engaging a PLL and just by using a simple filtering stage. In [39], two common second-order band-pass filters (BPFs) are employed for this purpose [1].

The objective of this chapter is to comprehensively investigate the performances of the two VCC approaches, i.e., the PLL-based VCC and the PLL-free VCC known as DPC in the publications [4, 38, 39]. For this approach two steps are taken: i) The VCC method is reviewed first, and an optimal multivariable controller is developed for it. This controller can be works with PLL-based and PLL-less VCCs equally well. With this optimal controller, both VCC

performances are improved beyond what is reported in the existing literature [1]. 2) The two VCCs are then put to the test in both normal and severe working conditions. These severe conditions consist of weak grid conditions, grid voltage and frequency fluctuations, as well as dramatic jumps in DER power. Finally, the outcomes of various simulations as well as some laboratory-scale experimental tests are collected, presented, and analysed [1].

The following conclusions are drawn from the investigation in this chapter. 1) When there are grid frequency offsets, the PLL-less VCC produces steady-state offset errors (in tracking the real and reactive power reference signals). 2) The grid frequency offsets can cause the PLL-less VCC to become unstable when accompanied with weak grid conditions. 3) Under weak grid conditions with grid low-voltage disturbances, the PLL-less VCC cannot ensure a pure sinusoidal output current. 4) All of the aforementioned complications are successfully resolved by using the PLL-based VCC [1]. 5) Both VCCs operate equally well under normal situation and when tracking the rapid changes in the reference signals of DER power amid weak grid conditions. The chapter concludes that the PLL-based VCC (if designed well and optimally, as proposed in this chapter) still outperforms the PLL-free VCC in severe operating conditions [1].

In order to mathematically explain the results of this chapter, it is observed that a single-phase PLL is theoretically similar to a second-order BPF if its frequency-update law is deactivated [18, 53]. Correspondingly, in the absence of a frequency loop, a three-phase PLL with both positive and negative-sequence loops is theoretically similar to two BPFs [3]. By using a PLL, One more first-order dynamics is added thanks to the frequency-update loop. Consequently, employing BPFs in the DPC, which is necessary due to the presence of grid voltage disturbances and distortions, is mathematically equivalent to using a PLL. However, because the BPF is unable to adjust the frequency, new problems in the DPC emerge [1].

Finally, in order to define the scope of this chapter, it is important to note that several controllers have been presented in the literature under the label "direct power control" (DPC). These include rectifier applications [54–56], active power filters [57], and doubly-fed induction generators (DFIGs) [41, 58–60]. The approaches of [41, 58–60] can operate under unbalanced and harmonically distorted grid conditions. Model predictive control (MPC) based DPCs have been proposed in [55, 61, 62]. The voltage-modulated DPC described in [4, 38, 39] is, however, the focus of the study in this chapter [1].

2.2 System Description

Figure 2.1 shows the schematic diagram of a grid-connected DER consisting of 1) a DER source modeled by a constant dc voltage source, 2) a two-level three-phase VSC, and an interface L -filter denoted by L_t with internal resistance R_t , 3) the grid model of an ideal voltage source with an inductance L_g , and 4) the schematic of the entire control system. The gating signals of the VSC switches are generated based on sinusoidal pulse-width modulation (SPWM) method [1].

The three-phase voltages at the point of common coupling (PCC), v_{sabc} , and the VSC three-phase output currents, i_{tabc} , are measured and used in the feedback controller. The control system of the VSC generates a set of three-phase voltages (v_{tabc}) in order to regulate (the dc-link voltage if need be and) the exchanged real/reactive powers with the grid [1].

In Fig. 2.1, we use and optimally design a multi-input multi-output (MIMO) controller instead of the common single-input single-output (SISO) controllers. The block marked by "Sync. Method" generates the synchronizing signals $\cos(\theta)$ and $\sin(\theta)$ which are used for abc/dq and dq/abc transformations [1].

The system parameters are given in Table 2.1 which are chosen according to [4,39] to facilitate the comparisons.

Table 2.1 Parameters of the Study Systems

Parameter	Symbol	Value	
		Simulation	Experiment
DC side voltage	V_{dc}	730 V	380 V
Filter resistance	R_t	30 m Ω	30 m Ω
Filter inductance	L_t	6 mH	5 mH
Grid resistance	R_g	60 m Ω	60 m Ω
Grid inductance	L_g	22 mH	20 mH
Grid phase voltage	V_g	110 V	60 V
System frequency	f	50 Hz	60 Hz
VSC power rating	S	20 kVA	20 kVA
Switching frequency	f_{sw}	10 kHz	5 kHz

2.3 Principles of Operation

The difference between the PLL-based VCC and the PLL-less VCC is only in "Sync. Method". While the PLL-based VCC uses a PLL to generate the synchronizing signals, the PLL-

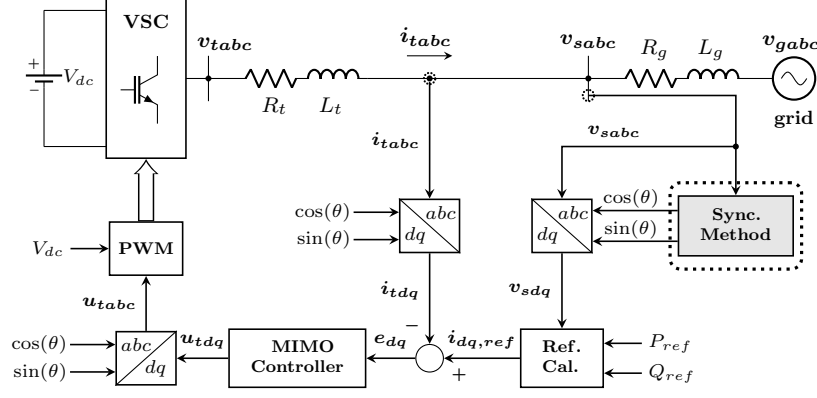


Figure 2.1 Single-line diagram of a grid-connected VSC with its control system [1]

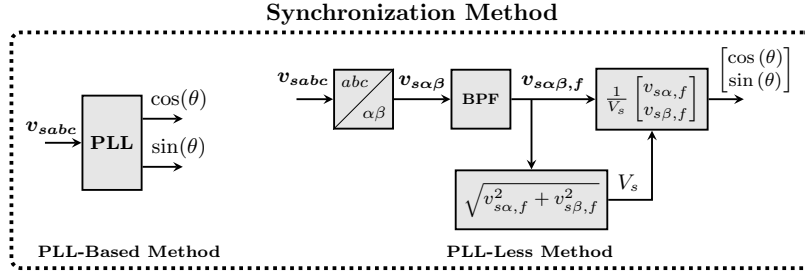


Figure 2.2 Synchronization methods, Left) PLL-based method, [2, 3], and Right) PLL-less method of [4] (Clarke transformation) [1]

less VCC either directly uses the grid voltage or just uses some filtered version of the grid voltage [1].

As shown in Fig. 2.2, the PLL-less VCC uses two standard second-order band-pass filter with the transfer function

$$G_{\text{BPF}}(s) = \frac{2\zeta\omega_o s}{s^2 + 2\zeta\omega_o s + \omega_o^2} \quad (2.1)$$

where ω_o is the nominal grid frequency in rad/s and ζ is a damping ratio. Employing filters is necessary in practical conditions due to the fact that the grid voltage at the point of common coupling (PCC) is noisy and has low-frequency harmonics as well [1].

Considering a set of balanced voltages at the PCC as

$$v_{sa} = V_s \cos(\theta), \quad v_{sb} = V_s \cos(\theta - \frac{2\pi}{3}), \quad v_{sc} = V_s \cos(\theta + \frac{2\pi}{3}),$$

where V_s is the voltage amplitude, and using the Clarke's transformation to obtain the voltage in the $\alpha\beta$ -frame lead to

$$v_{s\alpha} = V_s \cos(\theta), \quad v_{s\beta} = V_s \sin(\theta). \quad (2.2)$$

Therefore, the synchronizing signals, i.e., $\cos(\theta)$ and $\sin(\theta)$, can simply be obtained by dividing the $\alpha\beta$ signals to their magnitude. The magnitude is calculated according to

$$V_s = \sqrt{v_{s\alpha}^2 + v_{s\beta}^2}. \quad (2.3)$$

As for the PLL, there are multiple structures available in the literature [19]. The most basic and standard structure, called the dq-PLL or the SRF-PLL, comprises a simple feedback loop with a PI compensator that estimates the angle θ [18]. This PLL has been improved to address the unbalanced voltages [3, 18, 63]. In our study, we use the improved PLL (also called the UTSP) of [3] with the parameter setting of $\mu_1 = \mu_2 = 300$ and $\mu_3 = 9000$ [1].

2.4 Mathematical Control Model

Applying the KVL to the ac-side of the VSC of Fig. 2.1

$$v_{tabc} = R_t i_{tabc} + L_t \frac{di_{tabc}}{dt} + v_{sabc}.$$

Using Clarke's transformation, we obtain

$$v_{t\alpha\beta} = R_t i_{t\alpha\beta} + L_t \frac{di_{t\alpha\beta}}{dt} + v_{s\alpha\beta}, \quad (2.4)$$

where $v_{t\alpha\beta} = v_{t\alpha} + jv_{t\beta}$ and $i_{t\alpha\beta} = i_{t\alpha} + ji_{t\beta}$ are the representations of the VSC output voltages v_{tabc} and currents i_{tabc} in the $\alpha\beta$ -frame. Subsequently, transforming them into the dq -frame using the relation $x_{\alpha\beta} = x_{dq}e^{j\theta}$ results in [1]

$$\begin{aligned} \frac{di_{td}}{dt} &= -\frac{R_t}{L_t}i_d + \omega i_q + \frac{1}{L_t}(v_{td} - v_{sd}) \\ \frac{di_{tq}}{dt} &= -\omega i_d - \frac{R_t}{L_t}i_q + \frac{1}{L_t}(v_{tq} - v_{sq}) \end{aligned} \quad (2.5)$$

where θ is the phase-angle of the PCC voltages as indicated in (2.2), $x_{dq} = x_d + jx_q$, and $\omega = \frac{d\theta}{dt}$. It should be noted that this selection of θ leads to $v_{sd} = V_s$ and $v_{sq} = 0$.

The model in the state-space form is represented by [1]

$$\dot{x}_p = A_p x_p + B_p u + Fw, \quad y = C_p x_p, \quad (2.6)$$

where $x_p = [i_{td} \ i_{tq}]^T$, $u = [u_d \ u_q]^T = [v_{td} \ v_{tq}]^T$, $y = [P \ Q]^T$, and $w = [v_{sd} \ v_{sq}]^T$, and

$$A_p = \begin{bmatrix} -\frac{R_t}{L_t} & \omega \\ -\omega & -\frac{R_t}{L_t} \end{bmatrix}, B_p = \frac{1}{L_t} I_2, C_p = I_2, F = -B_p.$$

where I_2 is the 2×2 identity matrix. The system of (2.6) is an MIMO (2-input 2-output) LTI system which is open-loop stable, controllable, observable and minimum-phase [1].

2.5 Control System Structure and Design

Fig. 2.3 shows the structure of the proposed controller which comprises two subsystems: i) a servo-compensator or output controller, and ii) a stabilizing state feedback controller. The servo-compensator provides zero steady-state error for all desired outputs (or set-points), and for all disturbance signals with known dynamics. Moreover, the servo-compensator ensures steady-state decoupling between the control channels. The stabilizing controller is a state feedback gain and guarantees robust stability and robust performance of the closed-loop system. The servo-compensator is designed so that it includes dynamics of the set-points as well as those of the disturbance signals [64]. The stabilizing state feedback controller can be designed by any pole placement methods. In this chapter, a new approach to linear quadratic regulator (LQR) discussed in [5] is used to design the stabilizing state feedback controller [1].

2.5.1 Control Objectives

The controller is required to accomplish the following objectives.

1. Asymptotic stability of the closed-loop system.
2. Steady-state asymptotic tracking and regulation for constant set-points $y_{ref} = [P^*, Q^*]^T$ and disturbances w .
3. Robust closed-loop stability for as wide as possible a range of L-filter and grid impedance parameters.
4. Fast and smooth dynamic response with low interaction between the two control channels.

The above conditions can be achieved by designing a robust controller consisting of a servo-compensator whose structure and dynamics are determined based on the reference and disturbance signals, and a stabilizing state feedback gain which can be optimally designed, e.g., using the LQR approach. The proposed controller is explained below [1].

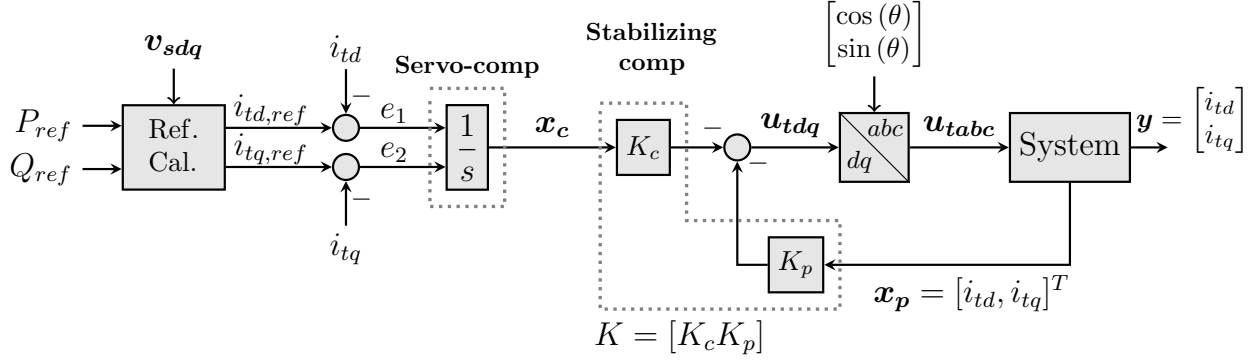


Figure 2.3 Structure of the proposed controller [1]

2.5.2 Controller Design

Since the reference and disturbance signals are stepwise functions, then according to [64], the core unit of the servo-compensator is an integral control. Moreover, since the error in the system, defined as $e = y_{ref} - y$, is a 2×1 vector, the servo-compensator becomes a 2-input 2-output system with two state variables. More specifically, the transfer function matrix of the servo-compensator is $diag(\frac{1}{s}, \frac{1}{s})$, and its minimal realization in state-space domain is [1]

$$\dot{x}_c = A_c x_c + B_c e = A_c x_c + B_c y_{ref} - B_c C_p x_p \quad (2.7)$$

where $A_c = 0_{2 \times 2}$ and $B_c = I_2$. Consequently, the state-space representation of the augmented system can be written as

$$\begin{aligned} \dot{x} &= A_{aug} x + B_{aug} u + B_{ref} y_{ref} + F_{aug} w, \quad y = C_{aug} x \\ A_{aug} &= \begin{bmatrix} A_c & -B_c C_p \\ 0 & A_p \end{bmatrix}, \quad B_{aug} = \begin{bmatrix} 0 \\ B_p \end{bmatrix} \\ B_{ref} &= \begin{bmatrix} B_c \\ 0 \end{bmatrix}, \quad F_{aug} = \begin{bmatrix} 0 \\ F \end{bmatrix}, \quad C_{aug} = [0 \quad C_p]. \end{aligned} \quad (2.8)$$

Equation set (2.8) describes an augmented system with the state variable vector $x = [x_c^T, x_p^T]^T$, which includes the plant's state variables as well as dynamics of the error vector. Furthermore, the controllability of (2.6) conveys the feasibility of arbitrary pole placement of the closed-loop system [64]. Therefore, a state feedback law

$$u = -Kx \quad (2.9)$$

is employed to assure the robustness of the closed-loop system. As shown in Fig. 2, the

controller gain $K_{2 \times 5}$ is a combination of the gain matrix $K_{c2 \times 2}$ which operates on the added control states (x_c) and the feedback controller gain matrix, $K_{p2 \times 3}$, such that $K = [K_c \ K_p]$. To meet the design requirements, an optimal control solution is provided by the Linear Quadratic Regulator (LQR) theory to minimize the cost function of J , which is extracted from (2.8) and (2.9), and is given by

$$J = \int_0^\infty (q_{01}e_1^2 + q_{02}e_2^2 + q_{p1}x_{p1}^2 + q_{p2}x_{p2}^2 + r_1v_1^2 + r_2v_2^2)dt. \quad (2.10)$$

In (2.10), the cost function, J , weights the error signals e_1 and e_2 with q_{01} and q_{02} , and the plant state variables x_{p1} and x_{p2} with q_{p1} and q_{p2} , respectively. It also weights the controller inputs, v_1 and v_2 by r_1 and r_2 , respectively. By defining a proper value for the cost factors, minimization of J will yield to a gain matrix K that ensures convergence of error signals in a fast and efficient manner.

The error, control, and state costs are initially defined as $q_{01} = q_{02} = \epsilon$, $r_1 = r_2 = 1$, $q_{p1} = q_{p2} = 0$, where ϵ is a small, positive nonzero real number. By successively increasing q_{01} and q_{02} , and q_{p1} and q_{p2} , and by plotting the eigenvalues of the closed-loop augmented system, an optimum design can be achieved. Fig. 2.4 represents the graphical approach for designing the gain matrix K . The design is based on the numerical values given in Table I. Eventually, the gain matrix K is obtained as

$$K = \left[\underbrace{\begin{bmatrix} -3684.5 & 741.5 \\ -741.5 & -3684.5 \end{bmatrix}}_{K_c} \mid \underbrace{\begin{bmatrix} 8.7 & -0.127 \\ -0.127 & 10 \end{bmatrix}}_{K_p} \right].$$

This gain matrix sets the eigenvalues of the closed-loop system at: $-967 \pm j402$ and $-594 \pm j70$. Based on such eigenvalues, a fast and smooth dynamic response is secured for the closed-loop system.

2.6 Performance Evaluation

The designed controller of Section 2.5.2 and the system parameters given in Table 2.1 are used to digitally simulate the grid-connected VSC in Matlab/Simulink. The simulations are performed at 50 Hz to allow better comparison with existing results in the literature pertaining to European system. The experimental setup, however, is based on an actual 60 Hz North American grid. The experimental setup parameters are given in Table 2.1.

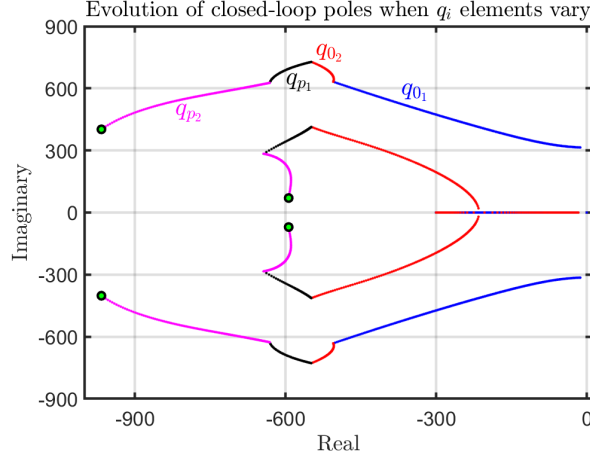


Figure 2.4 Root locus representing the design procedure of the controller using the *LQT* method of [5] [1]

2.6.1 Simulation Results

It is shown that both methods perform equally robust under normal and weak grid conditions but the PLL-less VCC shows errors and instability in two severe conditions: grid frequency offsets, and unbalanced grid voltage disturbances [1].

Stiff Grid

In this case study, the grid inductance is $L_g = 440 \mu\text{H}$ which represents a stiff grid, i.e., $SCR \approx 13.1$ at 20 kVA inverter power. Fig. 2.5 and 2.6, respectively, show the responses of the PLL-based and the PLL-less VCC when set-points change. As expected, they show identical transient and steady-state responses.

Weak Grid

In this simulation study, the grid inductance is increased to $L_g = 22 \text{ mH}$ for which $SCR \approx 2.6$ (at 2 kVA inverter operation), and the grid becomes extremely weak. Figs. 2.7 and 2.8 respectively show the responses of both PLL-based and PLL-less VCC systems to a step change in the real power set-point. As it is observed from Fig. 2.7, both systems respond but the PLL-based VCC offers slightly better/smoothen transient response.

To further verify the performance of the proposed methods, sudden set-point changes with higher magnitudes in the real and reactive powers are applied. While the VSC initially injects no current, the set-point of reactive power is increased from 0 to 2 kVar at $t = 0.1 \text{ s}$, and the real power reference value steps up from 0 to 4 kW at $t = 0.3 \text{ s}$. (The SCR will be

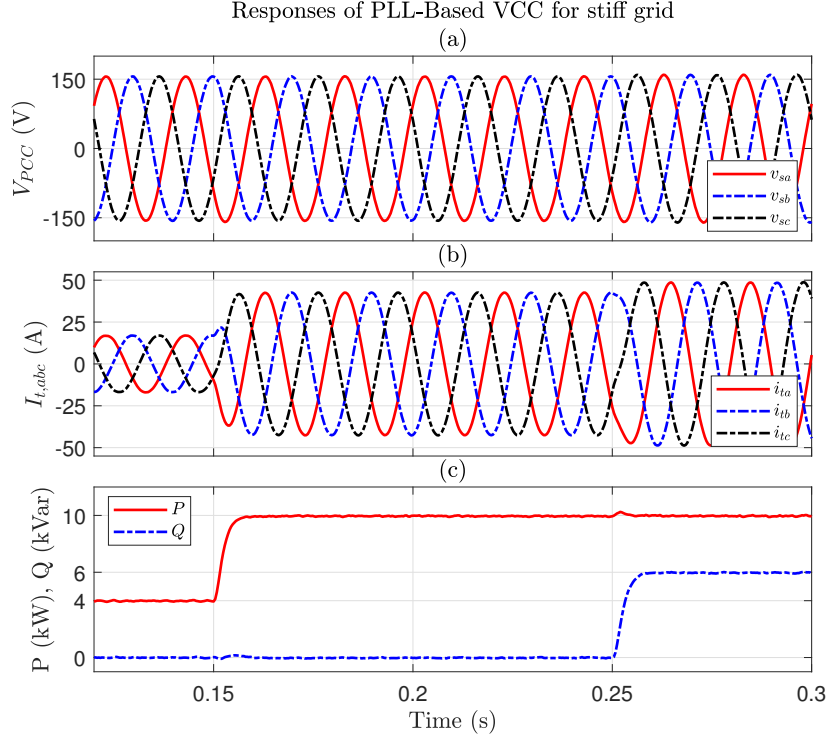


Figure 2.5 Dynamic response of the MIMO PLL-based VCC when the grid is stiff. (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

around 1.2 at this operating condition.) Figs. 2.9 and 2.10 show the responses of both PLL-based and PLL-less VCC systems, respectively. As observed, both methods show almost the same dynamic responses. It should be mentioned that our simulation studies show that both systems will be destabilized due to a step change in real power set-point higher than 4 kW. However, in order to achieve a higher real power injection by the VSC without jeopardizing closed-loop stability, we can increase the injected reactive power to a maximum value of 11 kVar. In this case, the maximum real power supplied can reach up to 8 kW. In other words, increasing the reactive power allows increased level of real power transfer.

Grid Frequency Drift

In this study, it is assumed that, during some abnormal conditions, the grid frequency could deviate from its nominal value by 1 – 5 Hz, and this may last several seconds. Both positive and negative frequency drift scenarios are considered when the grid is very weak ($SCR = 1.2$). While the set-points of the real and reactive powers are set to 4 kW and 2 kVar, respectively, the system frequency steps up from 50 Hz to 53 Hz (for PLL-based) and to 52 Hz (for PLL-less) at $t = 0.5$ s. Figs. 2.11 and 2.12 respectively show the responses of the PLL-based and PLL-less VCCs. As it is observed from Fig. 2.12, although the PLL-less VCC remains stable

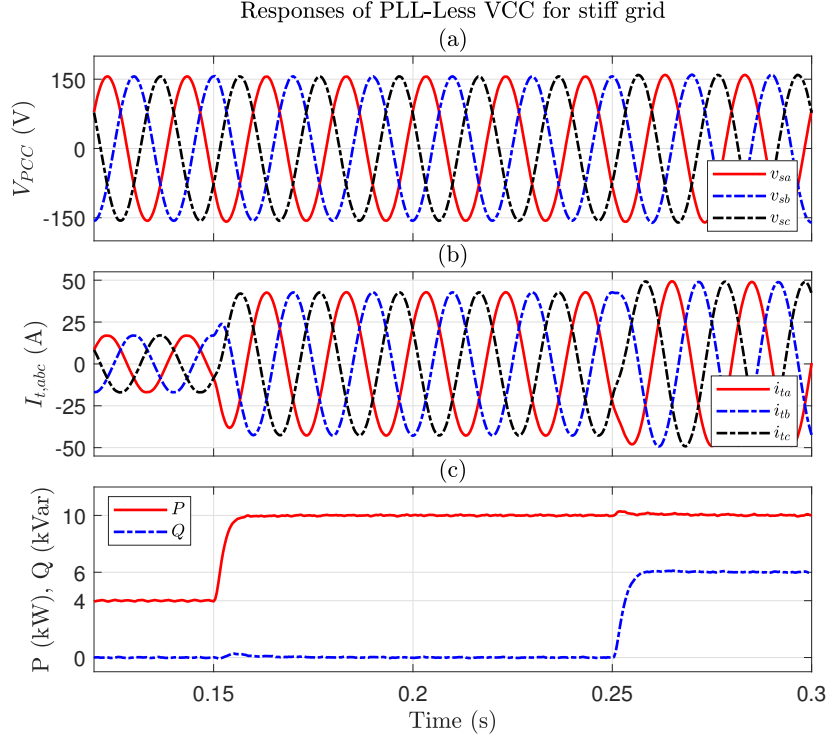


Figure 2.6 Dynamic response of the MIMO PLL-less VCC when the grid is stiff. (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

after this frequency excursion, steady-state errors in the real and reactive powers occur. The PLL-based VCC method has no difficulty in robustly tracking the reference values.

In the next scenario, while the VSC supplies 4 kW real power and 2 kVar reactive power to the grid, the system frequency is stepped down to 47 Hz (for PLL-based) and to 49 Hz (PLL-less) at $t = 0.5$ s. Figs. 2.13 and 2.14 show their responses. As observed from Fig. 2.12, the PLL-less VCC becomes unstable after this frequency excursion, whereas the PLL-based VCC is robust against the negative frequency drift and tracks the reference signals without any steady-state error.

Unbalanced Faults

In this case study, a single-phase-to-ground fault occurs at middle of the line at $t = 0.5$ s. The line between the PCC and the host grid is modeled by two identical series-connected inductors, each equal to 5.5 mH. This fault makes the PCC voltages unbalanced with an imbalance voltage factor of 17.5%. Figs. 2.15 and 2.16 respectively show the responses of the PLL-based and PLL-less VCCs due to this unbalanced fault. It is observed that the PLL-based VCC generates sinusoidal currents while the PLL-less VCC fails to generate sinusoidal currents during this voltage disturbance.

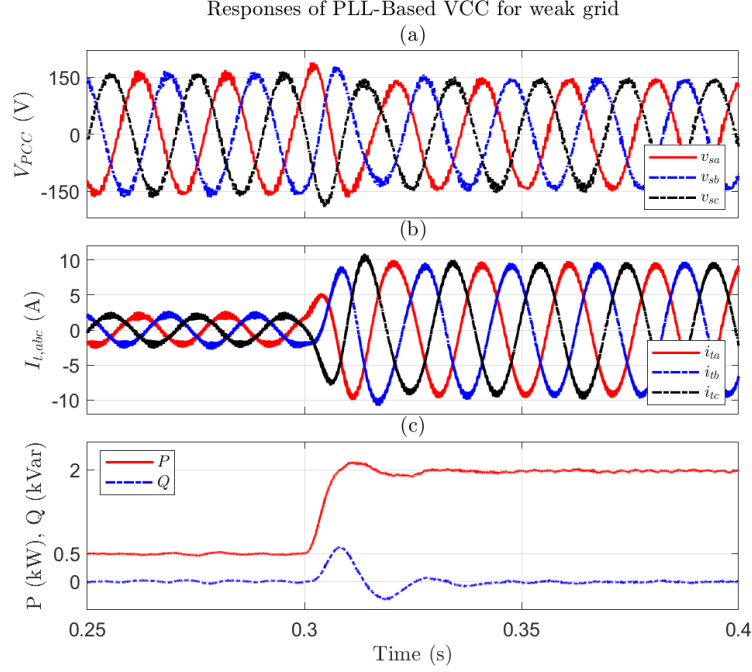


Figure 2.7 Dynamic response of the PLL-based VCC to set-point change in real power when the grid is weak ($SCR = 2.6$). (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

2.6.2 Experimental Results

The experimental setup consists of an MTM series IGBT inverter with power rating of 20 kW from GUASCH, an eZdspTMf28335 digital signal processor (DSP) for implementing the controller, and a PV emulator from Chroma as dc power source (See Fig. 2.17). The grid voltage is stepped down 60 V using a three-phase transformer. The experimental data are collected using an OP8660 current/voltage sensor, and are processed using OP5700 real-time simulator. The data are then stored on a PC and are plotted in MATLAB. For each method, two scenarios are tested under extremely weak grid (i.e., $L_g = 20$ mH).

Scenario I

In the first scenario, the grid is balanced and while the VSC injects some reactive power into grid, a step change in the real power is applied. Figs. 2.18 and 2.19 show the dynamic responses of the PLL-based and PLL-less VCC, respectively. Notice that the step changes are applied manually and this is why the responses are somewhat slow. We performed a number of tests which all confirm that both methods have almost the same real power injection capacity for any given reactive power injection.

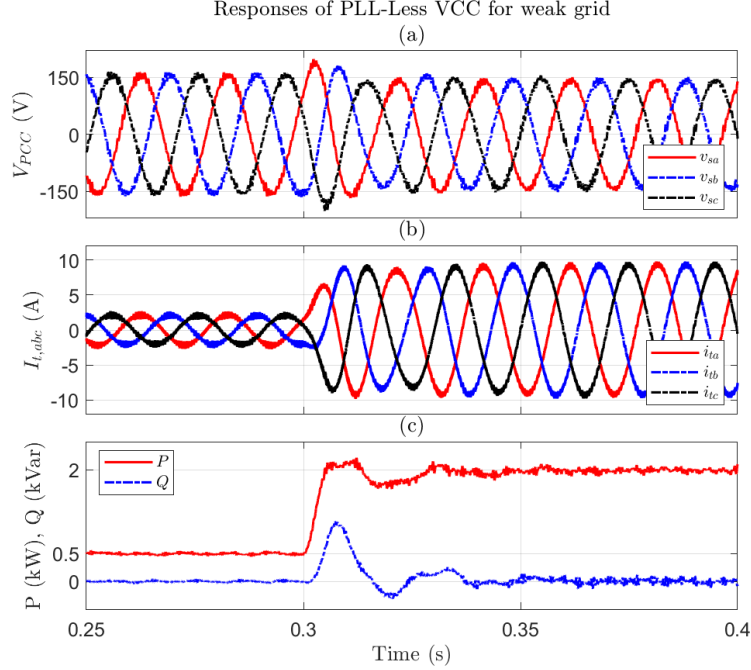


Figure 2.8 Dynamic response of the PLL-less VCC to set-point change in real power when the grid is weak $SCR = 2.6$. (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

Scenario II

In the second scenario, while the VSC injects both real and reactive powers, a non-zero impedance single-phase-to-ground fault occurs which adversely affects the power quality. Due to this single-phase fault, the PCC voltage becomes unbalanced with the imbalance factor of about 16%. Figs. 2.20 and 2.21 show that the PLL-based VCC offers much better power quality in terms of current THD.

2.6.3 Analogy between PLL and BPF dynamics

The above results show the incorrectness of the claim that the PLL-less VCC improves the performance in weak grid conditions. In this section, we show that a BPF introduces dynamics which are very similar to a PLL. Therefore, contrary to what is claimed, the BPFs cause the same problems that the PLL does (in weak grid conditions).

Structural block diagram of the enhanced PLL (ePLL) [18] is shown in Fig. . The ePLL receives the grid voltage $v_s(t)$ as the input, and provides the synchronizing unit functions $\cos \theta$, $\sin \theta$, the magnitude A , the frequency ω , and the phase angle θ .

The three controlling parameters of the ePLL are: μ_1 , μ_2 , and μ_3 . Assuming $\mu_1 = \mu_2 = \mu$,

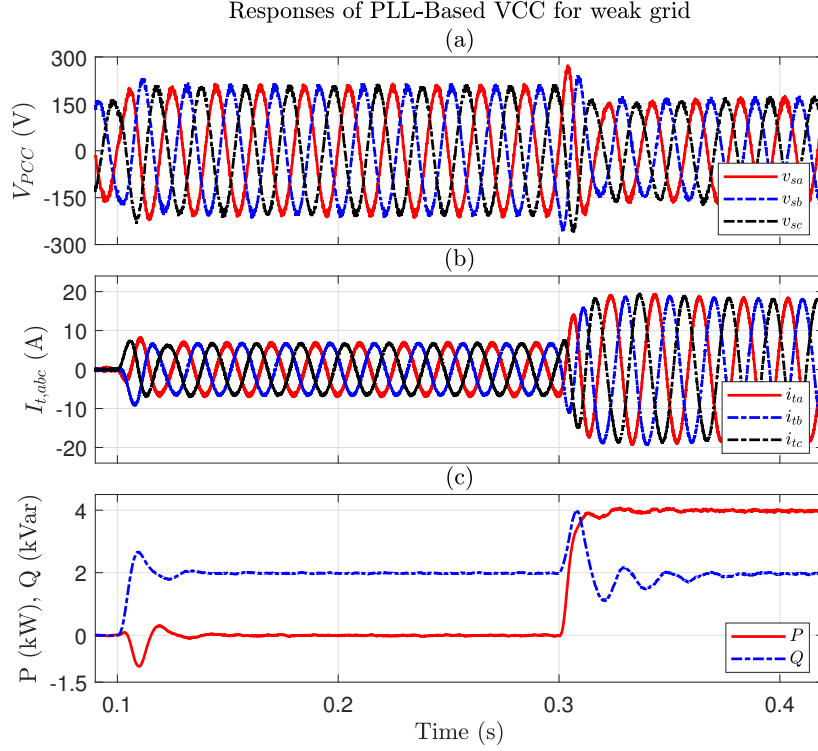


Figure 2.9 Dynamic response of the PLL-based VCC to set-point changes in real and reactive powers when the grid is weak ($SCR = 1.2$). (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

the two main equations from Fig. 2.22 are derived by inspection as

$$\dot{A} = \mu e \cos \theta, \quad \dot{\theta} = \omega - \mu \frac{e}{A} \sin \theta \quad (2.11)$$

where, $e = v_s - A \cos \theta$. If we define $x_1 = A \cos \theta$ and $x_2 = A \sin \theta$, it is readily seen that

$$\begin{aligned} \dot{x}_1 &= \dot{A} \cos \theta - A \dot{\theta} \sin \theta = \mu e \cos^2 \theta - A \left(\omega - \mu \frac{e}{A} \sin \theta \right) \sin \theta \\ &= -\omega x_2 + \mu e \\ \dot{x}_2 &= \dot{A} \sin \theta + A \dot{\theta} \cos \theta = \mu e \sin \theta \cos \theta + A \left(\omega - \mu \frac{e}{A} \sin \theta \right) \cos \theta \\ &= \omega x_1 \end{aligned} \quad (2.12)$$

If the dynamics of ω are neglected, the two equations in (2.12) describe an LTI system. Taking the Laplace transform of (2.12) and setting $\omega = \omega_0$ to be a constant, yields

$$X_1(s) = \frac{\mu s}{s^2 + \omega_0^2} E(s), \quad X_2(s) = \frac{\mu \omega_0}{s^2 + \omega_0^2} E(s) \quad (2.13)$$

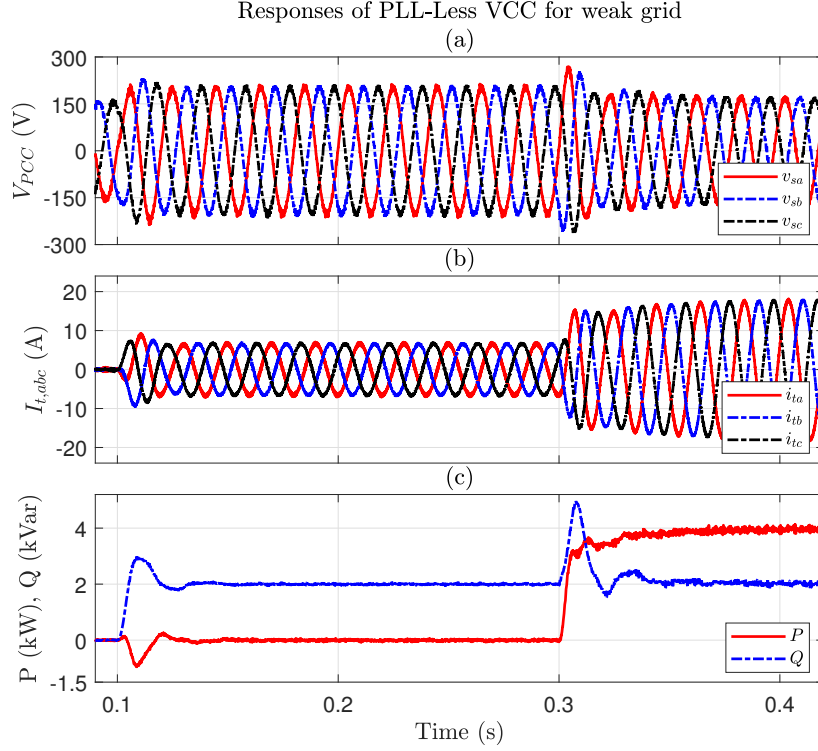


Figure 2.10 Dynamic response of the PLL-less VCC to set-point changes in real and reactive powers when the grid is weak $SCR = 1.2$. (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

and given that $E(s) = V_s(s) - X_1(s)$, one can obtain

$$X_1(s) = \frac{\mu s}{s^2 + \mu s + \omega_0^2} V_s(s), \quad X_2(s) = \frac{\mu \omega_0}{s^2 + \mu s + \omega_0^2} V_s(s). \quad (2.14)$$

The PLL and BPF exhibit the same dynamics for $\mu = 2\zeta\omega_0$ when comparing (2.14) with (2.1). The simulations demonstrate that in weak grid conditions, the PLL-less VCC (which uses BPF) introduces the same dynamics as the PLL-based VCC. This is shown in the simulation results.

2.7 Improving PLL-less BPFs for unbalanced grids

The PLL-less VCC's two BPFs use $v_{s\alpha}$ and $v_{s\beta}$ to operate. Under voltage imbalance, these two parameters can be represented as

$$v_{s\alpha} = V_p \cos \theta + V_n \cos(\theta + \delta), \quad v_{s\beta} = V_p \sin \theta - V_n \sin(\theta + \delta) \quad (2.15)$$

where δ is the relative phase-angle of the negative-sequence component with respect to the positive-sequence component and V_p and V_n are the magnitudes of the positive- and negative-

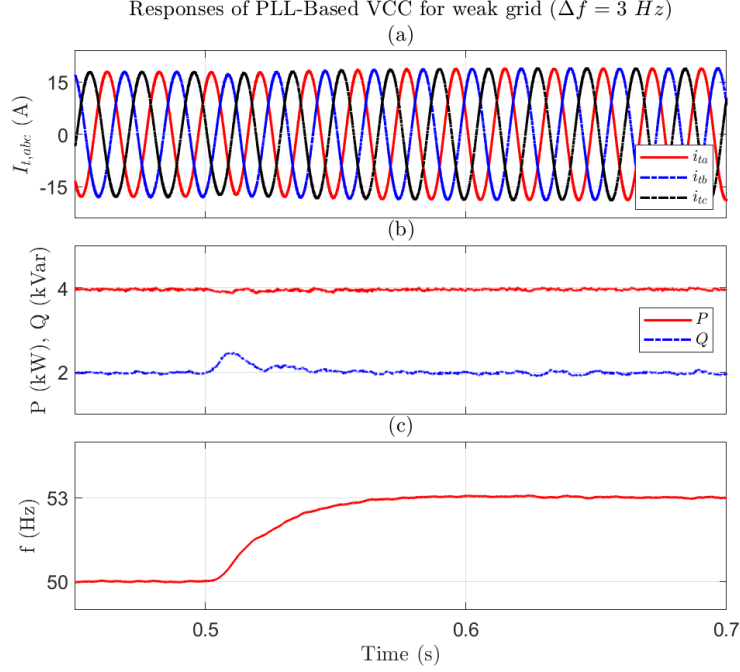


Figure 2.11 Dynamic response of the PLL-based VCC when the grid frequency steps up from 50 Hz to 53 Hz. (a) PCC voltages, (b) real and reactive powers, and (c) system frequency extracted by the PLL [1]

sequence components, respectively. As a result, the negative-sequence component taints the BPFs' outputs. In this section, we suggest a technique for precisely extracting the terms $\cos \theta$ and $\sin \theta$.

Take the BPF implementation in Fig. 2.23. These are its differential equations

$$\dot{y}_1 = -\mu y_1 - y_2 + \mu u, \quad \dot{y}_2 = \omega_0^2 y_1 \quad (2.16)$$

which readily results in

$$Y_1(s) = \frac{\mu s}{s^2 + \mu s + \omega_0^2} U(s), \quad Y_2(s) = \frac{\mu \omega_0^2}{s^2 + \mu s + \omega_0^2} U(s). \quad (2.17)$$

Accordingly, $y_1(t)$ represents the filtered version of the input, and $(1/\omega_0)y_2(t)$ represents the 90° delayed version of $y_1(t)$. Consequently, the two BPFs will produce the following four

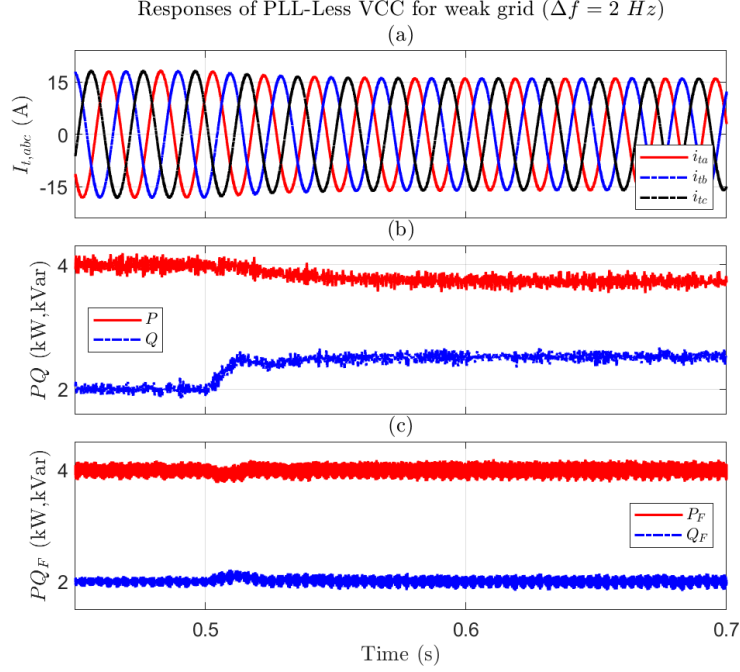


Figure 2.12 Dynamic response of the PLL-less VCC when the grid frequency steps up from 50 Hz to 52 Hz. (a) PCC voltages, (b) actual measured real and reactive powers of the VSC, and (c) real and reactive powers computed/used by control system [1]

outputs

$$\begin{aligned}
 y_{1\alpha} &= V_p \cos \theta + V_n \cos(\theta + \delta) \\
 \frac{1}{\omega_0} y_{2\alpha} &= V_p \sin \theta + V_n \sin(\theta + \delta) \\
 y_{1\beta} &= V_p \sin \theta - V_n \sin(\theta + \delta) \\
 \frac{1}{\omega_0} y_{2\beta} &= -V_p \cos \theta + V_n \cos(\theta + \delta).
 \end{aligned} \tag{2.18}$$

Now let's define

$$z_1 = \frac{1}{2}(y_{1\alpha} - \frac{1}{\omega_0} y_{2\beta}), \quad z_2 = \frac{1}{2}(y_{1\beta} + \frac{1}{\omega_0} y_{2\alpha}). \tag{2.19}$$

Which yields

$$\cos \theta = \frac{z_1}{\sqrt{z_1^2 + z_2^2}}, \quad \sin \theta = \frac{z_2}{\sqrt{z_1^2 + z_2^2}}. \tag{2.20}$$

The simulation scenario used in Section 2.6.1 is reproduced here to confirm the performance of the enhanced PLL-less technique under imbalanced grid voltage. The results, which reveal a considerable increase in power quality in the PCC voltages and VSC currents, are presented in Fig. 2.24.

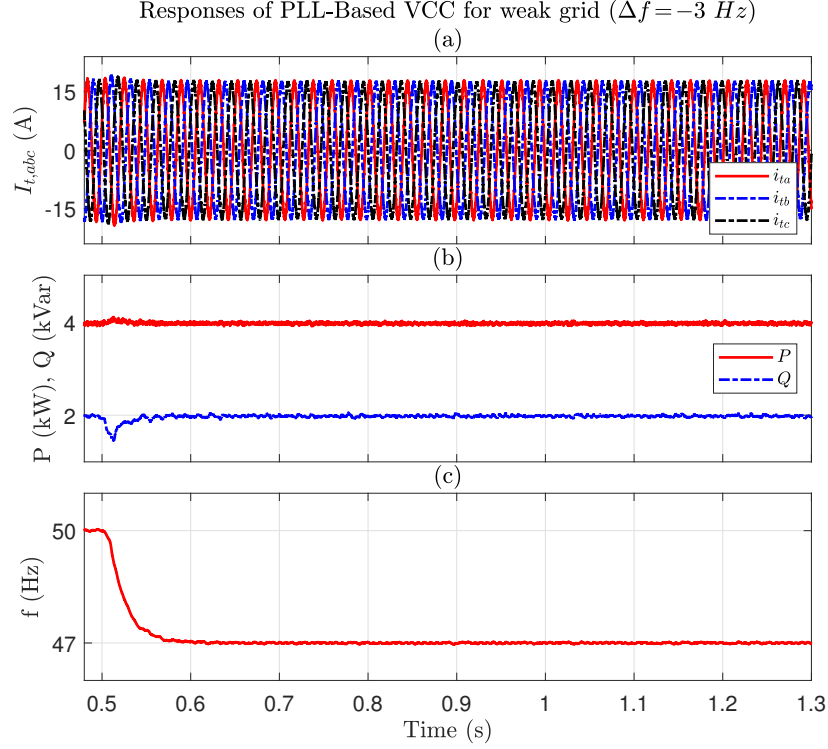


Figure 2.13 Dynamic response of the proposed PLL-based VCC when the grid frequency steps down from 50 Hz to 47 Hz. (a) PCC voltages, (b) real and reactive powers of VSC, and (c) system frequency extracted by the PLL [1]

2.8 Conclusion

The study in this chapter compared the performance of the widely used VCC strategy based on the use of a PLL with a recently advocated alternative approach that does not utilize a PLL but instead employs two second-order bandpass filters (BPFs). To begin with, both PLL-less and PLL-based methods are improved in the following regards: using a multivariable control structure, and optimal design of its feedback gains. The performance of both approaches in weak grid conditions was then demonstrated through comprehensive simulations and testing. This was explained mathematically by the fact that the BPFs produce dynamics that are very similar to the PLL, ultimately making it impossible to enhance the performance under weak grid circumstances. The PLL-less technique, however, falls short under two circumstances: (i) grid frequency offsets and (ii) imbalanced grid voltage breakdowns. A solution to the latter problem then was proposed in this study. Finally, the conclusion is that, if the controller and PLL are well-designed, as demonstrated here, the PLL-based VCC technique is still a better option.

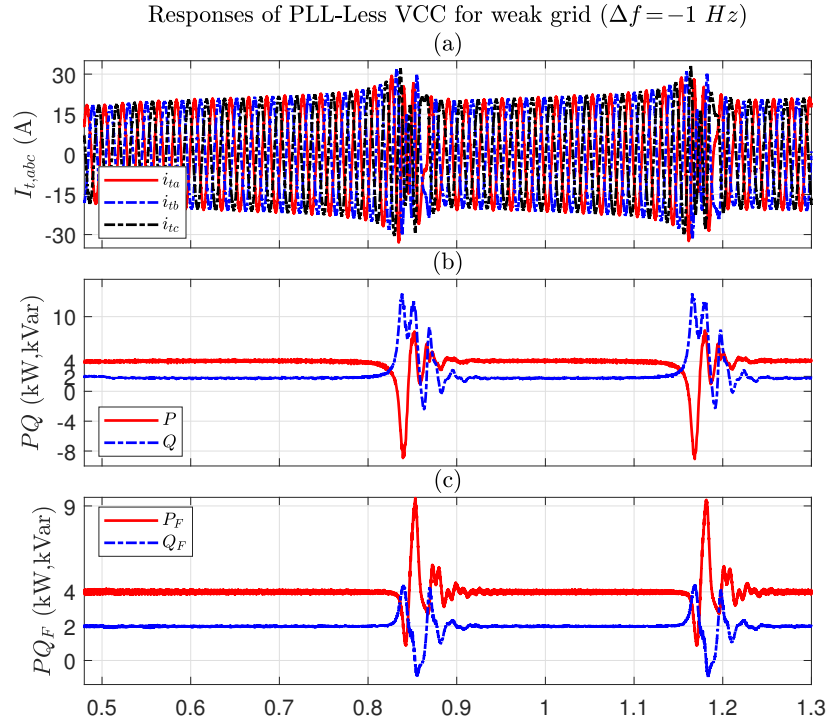


Figure 2.14 Dynamic response of the proposed PLL-less VCC when the grid frequency steps down from 50 Hz to 49 Hz. (a) PCC voltages, (b) real and reactive powers of the VSC, and (c) real and reactive powers computed by control system (filtered powers) [1]

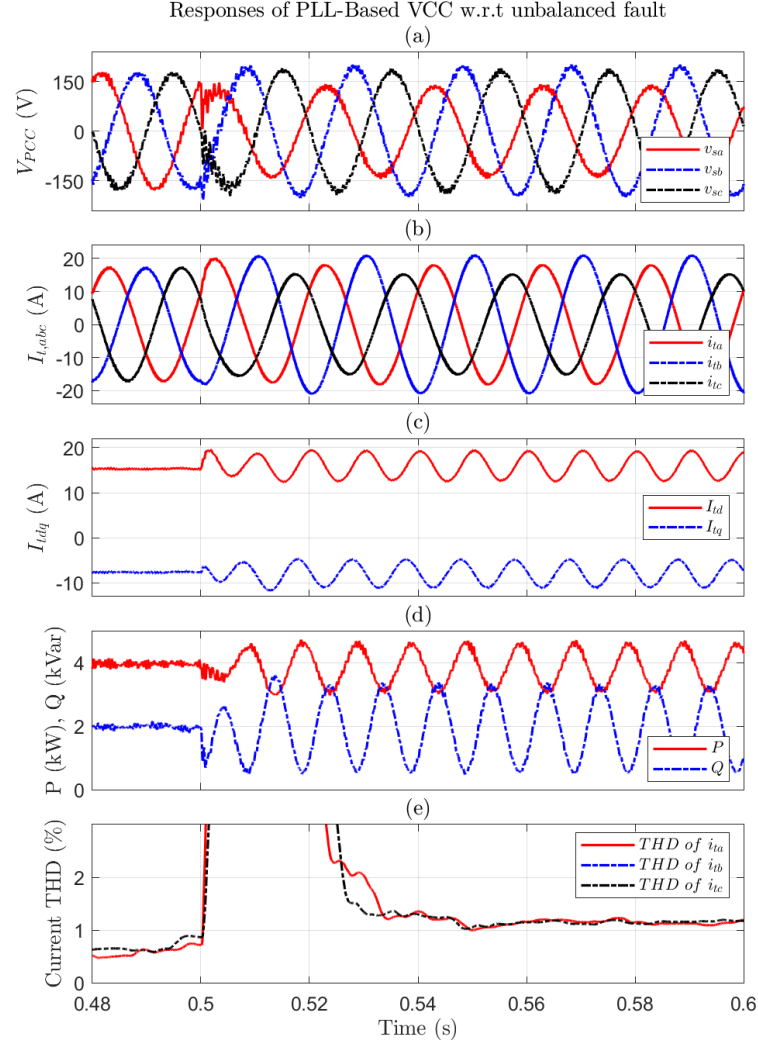


Figure 2.15 Power quality measure of the PLL-based VCC when a remote single-phase fault occurs. (a) PCC voltages, (b) VSC currents, (c) VSC dq-currents, (d) real and reactive powers, and (e) THDs of VSC currents [1]

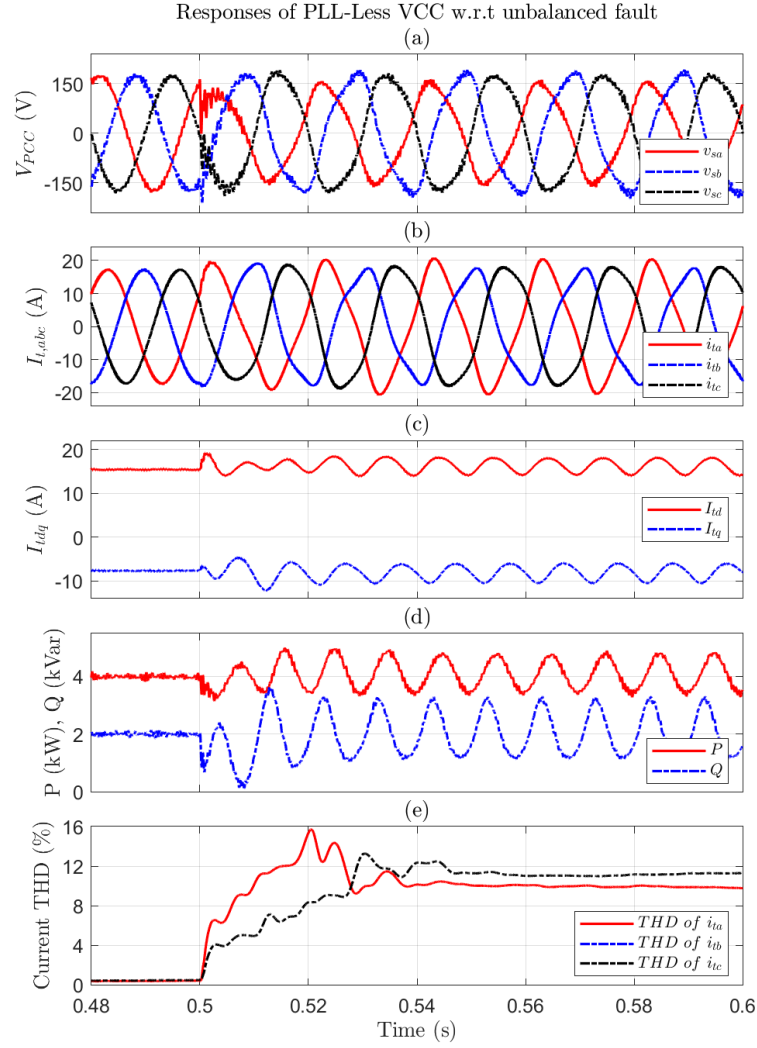


Figure 2.16 Power quality measure of the PLL-less VCC when a remote single-phase fault occurs. (a) PCC voltages, (b) VSC currents, (c) VSC dq-currents, (d) real and reactive powers, and (e) THDs of VSC currents [1]

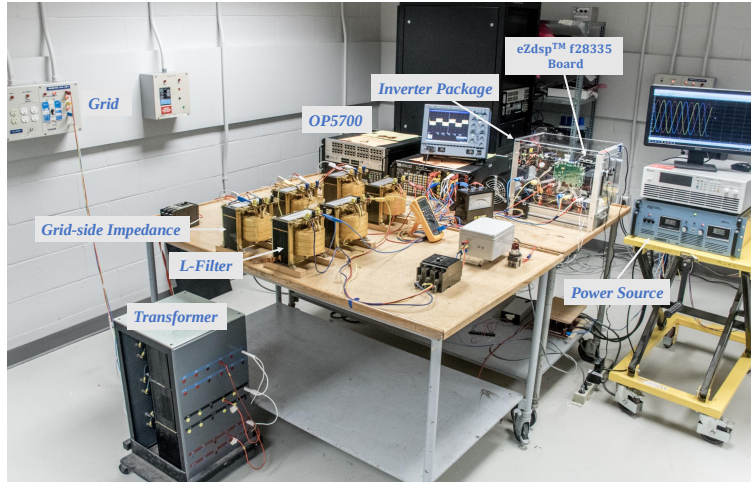


Figure 2.17 A photo of experimental setup [1]

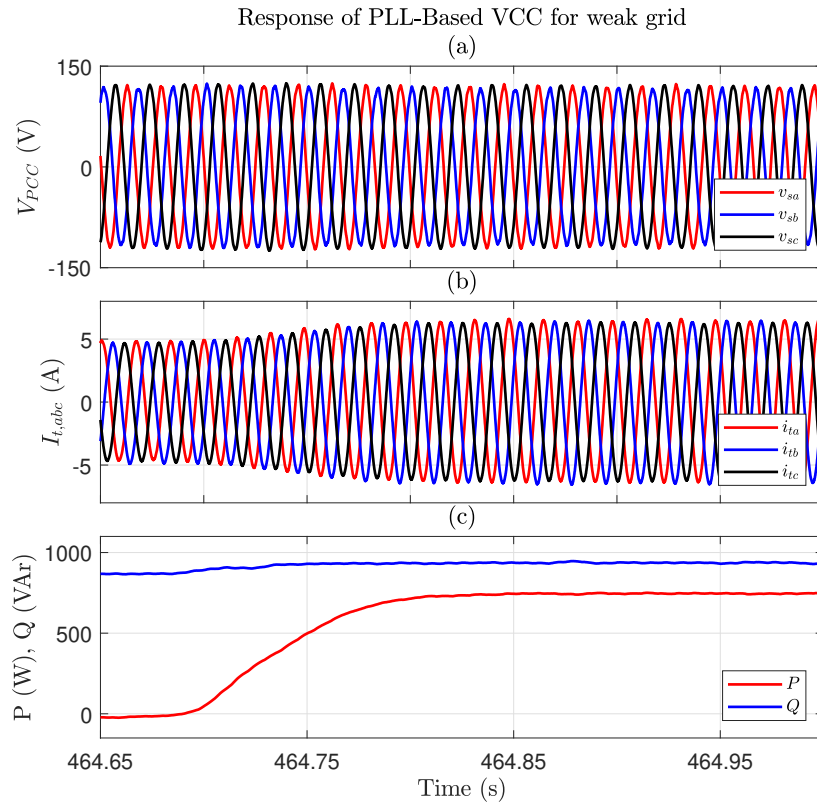


Figure 2.18 Response of the PLL-based VCC to a step change in real power: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

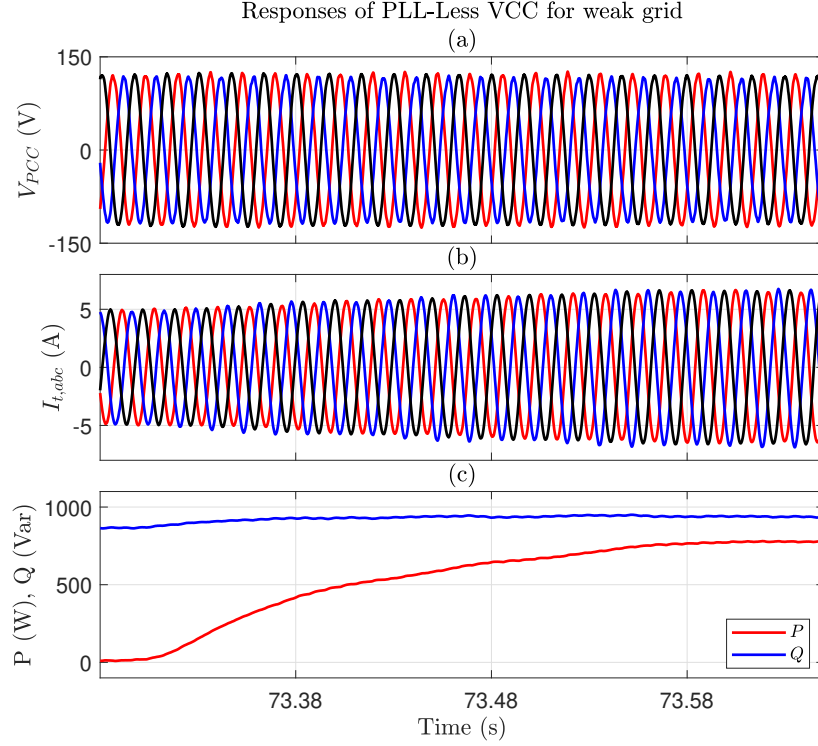


Figure 2.19 Response of the PLL-less VCC to a step change in real power: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [1]

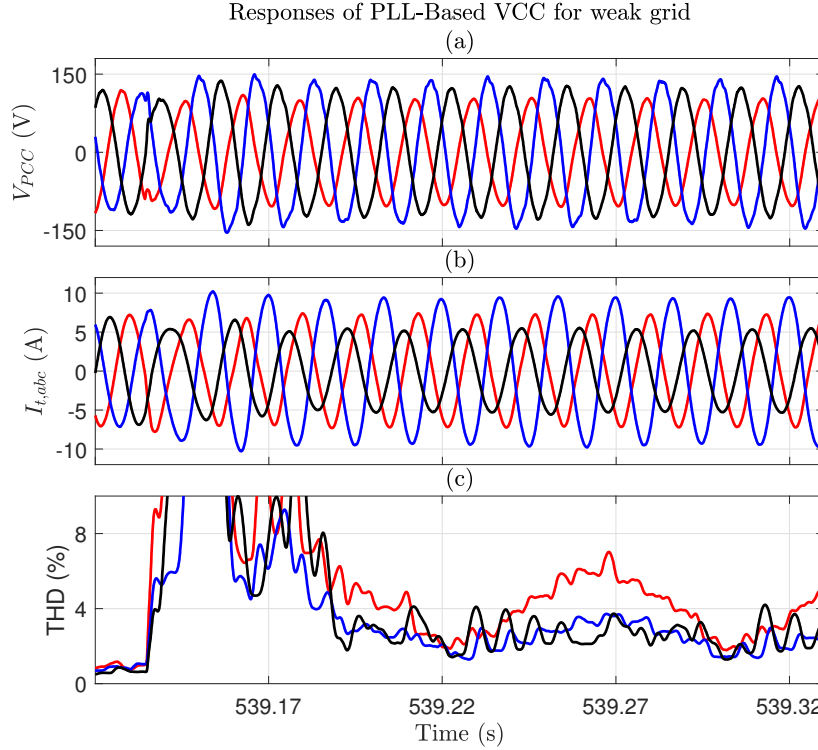


Figure 2.20 Response of the PLL-based VCC to a non-zero impedance single-phase fault: (a) PCC voltages, (b) VSC currents, (c) THDs of VSC currents [1]

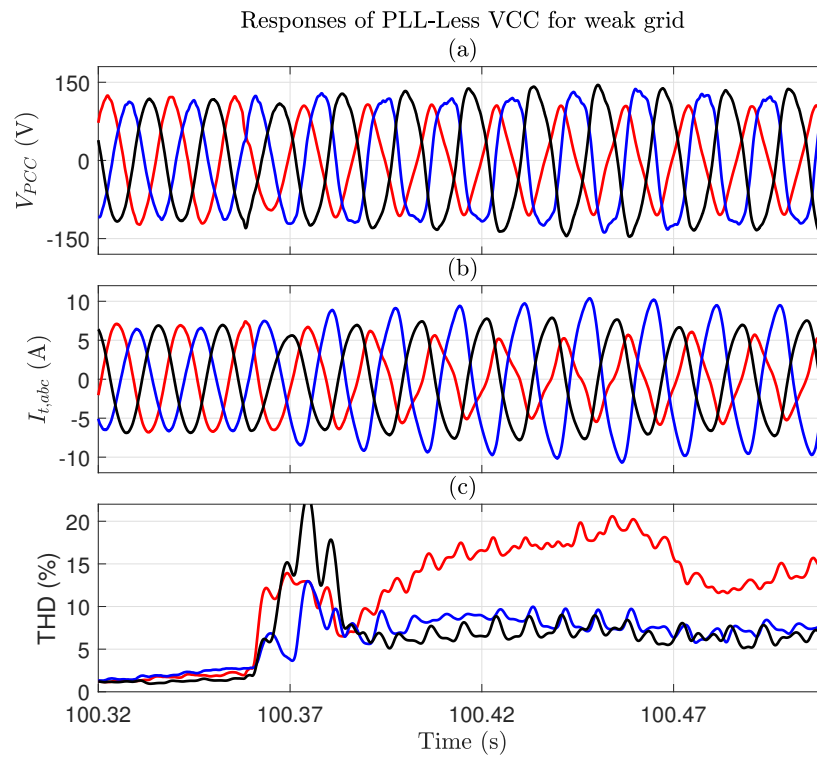


Figure 2.21 Response of the PLL-less VCC to a non-zero impedance single-phase fault: (a) PCC voltages, (b) VSC currents, (c) THDs of VSC currents [1]

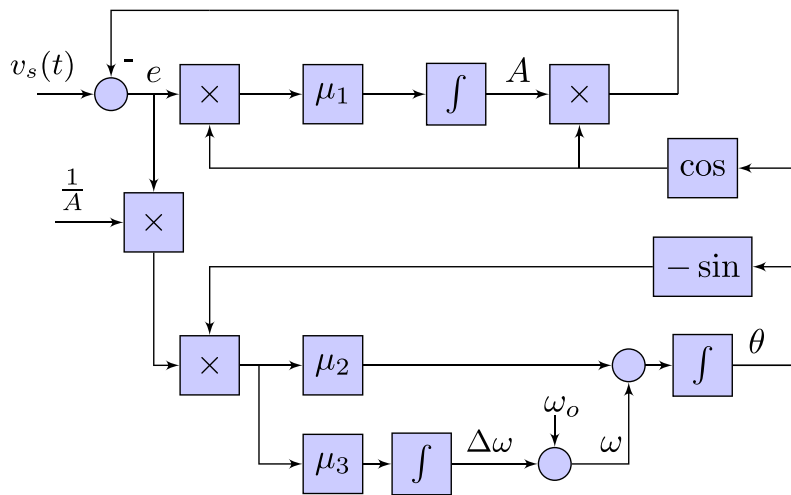


Figure 2.22 ePLL Structure [1]

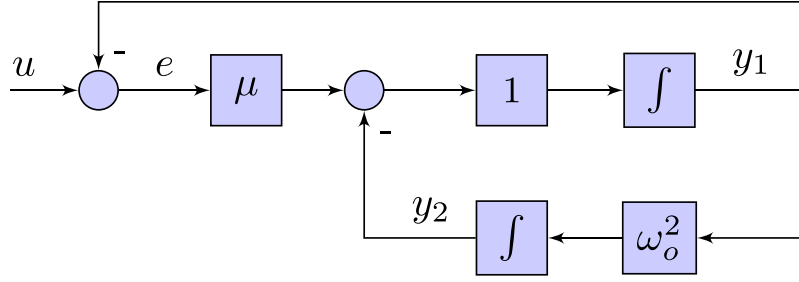


Figure 2.23 Proper implementation of the BPF [1]

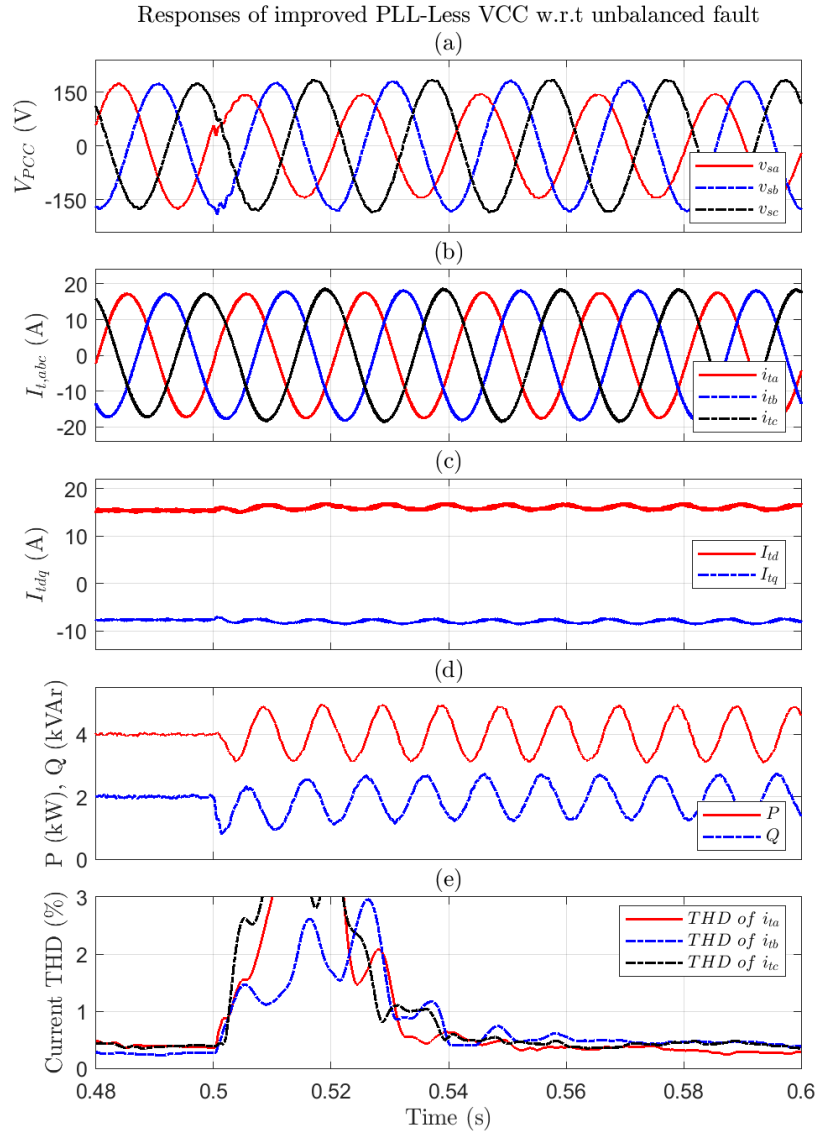


Figure 2.24 Power quality measure of the improved PLL-less VCC when a remote single-phase fault occurs. (a) PCC voltages, (b) VSC currents, (c) VSC dq-currents, (d) real and reactive powers, and (e) THDs of VSC currents [1]

CHAPTER 3 ENHANCED PERFORMANCE OF VOLTAGE SOURCE CONVERTERS UNDER WEAK GRID CONDITIONS

3.1 Background

Due to high penetration of electronically-interfaced distributed energy resource (DER) units such as renewable energy and battery energy storage systems, voltage source converters (VSCs) are increasingly utilized in the modern power system [1,65]. As mentioned in Chapter 2, in order to achieve grid synchronisation, control of the grid-connected VSC-based DERs is often carried out based on vector current control (VCC) and by using a phase-locked loop (PLL) system [66]. Under weak grid conditions, however, this control approach encounters some stability problems [67–70]. The root cause has been determined to be the interactions between PLL and other control loops [22, 23, 71, 72]. According to the research in [22], the dynamics of the ac-bus voltage are highly nonlinear and strongly coupled to those of PLL. If this problem is not appropriately addressed, the stability will suffer and eventually leads to instability under weak grid conditions, i.e., at lower short circuit ratio (SCR).

The proportional resonant (PR) based VCCs are less affected by PLL dynamics [34–36]. The effect of PLL on the output impedance of grid-tied inverters for both synchronous and stationary (i.e., dq and $\alpha\beta$) reference frames has been investigated in [36] which reveals that VCCs implemented in $\alpha\beta$ -reference frame possess less significant negative damping than those realized in dq-frame [6]. However, outer control loops, such as the PCC voltage loop, have an equivalent impact on both control strategies and lead to a decline in performance under weak grid conditions.

However, because of interconnections with extremely high impedance grid in remote areas, operation of VSCs is inevitable under weak grid conditions [73]. Reducing PLL gains is suggested in [42] in order to boost VSC's power capacity. However, the extent to which the PLL gains can be lowered is constrained by the dynamic performance criteria. An impedance-conditioning term in the PLL which synchronizes to a virtual point with a stronger voltage is proposed in [74]. Despite its advantages, this approach still has stability issues associated with PLL nonlinear dynamics, and it might not be suitable in extremely weak grid conditions. In [23], an impedance reshaping method based on double-PLLs is proposed. However, in order to study the destabilizing factors, the bandwidth of inner current control loop in [23] is assumed to be infinite, making the enhancements exclusive to the outer control loops. A frequency coupling suppression control approach is proposed in [44] in order to enhance the performance of single-phase grid-tied inverters. This resulting in stability improvement under

weak grid conditions, however, the work in [44] is only applicable for single-phase inverters. In [32], the dynamics of an enhanced PLL (ePLL) [46], are included into the single-phase inverter's extended current control loop which improves the converter stability margins in weak grid conditions. However, unless two independent single-phase PLLs are used for the α and β axes, which is not a typical practise, it cannot be readily extended to a three-phase system.

Some literature has suggested modified PLL structures to increase performance under distorted grid conditions [75–78], however they do not address the stability degradation issue when grid is extremely weak. Further, as discussed in Chapter 2, PLL-less VCCs have been proposed in [29–31], in which the synchronizing signals are generated from the grid voltage by using some band-pass filters (BPFs). However, the comprehensive analysis in Chapter 2 reveals that under weak grid conditions, these methods are susceptible to the same destabilization problems as PLL-based methods since the dynamics created by such filters are very similar to those of a PLL [1].

The objective of this chapter is to improve the stability of grid-connected VSCs controlled in stationary-frame under weak and extremely weak grid conditions. In this chapter, a new approach is proposed to systematically include the PLL dynamics in the controller. First, a detailed linear time-invariant (LTI) model of the system is derived by integrating the internal state variables of a recently introduced three-phase ePLL, also called the unified three-phase signal processor (UTSP) [3], into the dynamic equations of the VSC. It is shown that in a stationary frame, the PLL introduces coupling terms between the two control channels. Then, based on the developed model a robust multivariable controller is proposed which employs state feedbacks of the VSC variables as well as those of the ePLL [6]. The controller gains are designed using an extended version of a recent optimal control theory approach [79].

The main contributions of the research in this chapter are summarized as follows. i) The effects of the PLL on the dynamics of a VCC in stationary reference frame is formulated and studied. ii) The resulting model is not based on linearization, in contrast to other studies in the literature. iii) It is demonstrated that the PLL introduces symmetric cross-coupling terms between the controller's control channels. iv) By applying a multivariable control strategy on the augmented model of the grid-connected VSC, i.e., the converter and the PLL combined, the performance of the system has been improved under weak grid conditions. v) The problem of reactive power constraint under weak grid conditions is revisited.

Detailed theoretical analysis, software simulations, and experimental results are presented in this chapter to demonstrate the effectiveness of the proposed controller in improving the stability and robustness margins of a VSC operating in weak grid conditions.

The chapter is organized as follows. Mathematical model of the system is provided in Section 3.2. The structure of the proposed controller, control design procedure and robustness analysis are presented in Section 3.3. The problem of reactive power constraint under weak grid conditions is rigorously discussed in Section 3.4. In Section 3.5, the performance of the proposed controller is verified through MATLAB simulations, and the experimental tests for various scenarios are presented in Section 3.6. Finally, Section 3.7 concludes the study in this chapter.

3.2 Mathematical Model of the System

3.2.1 System Description

Fig. 3.1 shows the schematic diagram of the two-level three-phase VSC which is interfaced to the point of common coupling (PCC) through the series inductor L_t and resistor R_t . The grid is modeled by an ideal three-phase voltage source and a series impedance $R_g + j\omega L_g$. The dc side of the DER is modeled by an ideal voltage source with fixed voltage V_{dc} , and thus the dynamics of dc side are not considered in this study. The PCC three-phase voltages, $v_{s,abc}$, and the output currents, i_{abc} , are transformed into the stationary reference ($\alpha\beta$) frame and then are transmitted to the UTSP and the controller, respectively. Symmetrical components of voltage are estimated by the UTSP and are used by the reference generator block (i.e., Ref. Gen.) to generate the reference current signals. The output of the UTSP is also passed on to the controller where the control output U_{abc} is produced and applied to the VSC through modulation block. The gating signals of the VSC switches are generated based on sinusoidal pulse-width modulation (SPWM) technique [6].

3.2.2 Modeling of Weak Grid Condition

This section derives an equation for the PCC voltage under weak grid conditions that shows how this voltage is a function of grid impedance and converter current. This equation will be used to obtain the complete system model in the next section. Consider the set of balanced PCC voltages

$$v_{s_a} = V_m \cos(\omega t), \quad v_{s_b} = V_m \cos(\omega t - \frac{2\pi}{3}), \quad v_{s_c} = V_m \cos(\omega t + \frac{2\pi}{3}),$$

where V_m is the voltage amplitude, and ω is the angular frequency of the grid. Using the Clarke's transformation obtains the PCC voltage in $\alpha\beta$ -frame as $v_{s_\alpha} = V_m \cos(\omega t)$ and $v_{s_\beta} =$

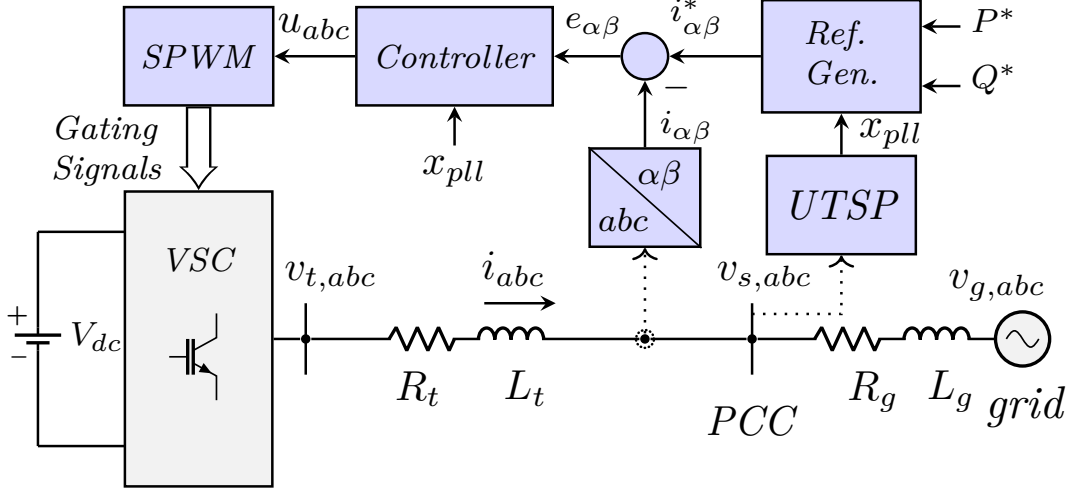


Figure 3.1 Single-line diagram of a grid-connected VSC and its controller [6]

$V_m \sin(\omega t)$. Applying the KVL to the ac-side of the VSC of Fig. 3.1, yields

$$v_{t\alpha\beta} = R_t i_{\alpha\beta} + L_t \frac{di_{\alpha\beta}}{dt} + v_{s\alpha\beta} \quad (3.1)$$

where $v_{t\alpha\beta}$, and $i_{\alpha\beta}$ are respectively VSC output voltages and currents in $\alpha\beta$ -frame.

In weak grid conditions, the coupling between PCC voltage and converter current increases and voltage becomes dependant on VSC current. This relationship may be modeled by

$$v_{s\alpha\beta} = L_g \frac{di_{\alpha\beta}}{dt} + R_g i_{\alpha\beta} + v_{g\alpha\beta}, \quad (3.2)$$

where $v_{g\alpha\beta}$ denotes the behind the impedance grid voltage in $\alpha\beta$ -frame. Substituting $\frac{di_{\alpha\beta}}{dt}$ from (3.1) into (3.2) yields

$$v_{s\alpha\beta} = \frac{L_t}{L_g + L_t} v_{g\alpha\beta} + \frac{L_g}{L_g + L_t} v_{t\alpha\beta} + \frac{(R_g L_t - L_g R_t) i_{\alpha\beta}}{L_g + L_t}, \quad (3.3)$$

which expresses the PCC voltage in terms of the VSC output voltage, the ideal grid voltage, and the converter current. Equation (3.3) indicates that the presence of R_g and L_g causes the PCC voltage to be a function of the VSC current.

3.2.3 LTI Modeling of Three-phase ePLL (UTSP)

The three-phase ePLL of [3] (which is a simplified version of the UTSP of [7]) is shown in Fig. 3.2. It is basically the combination of two interleaved PLLs for positive- and negative-

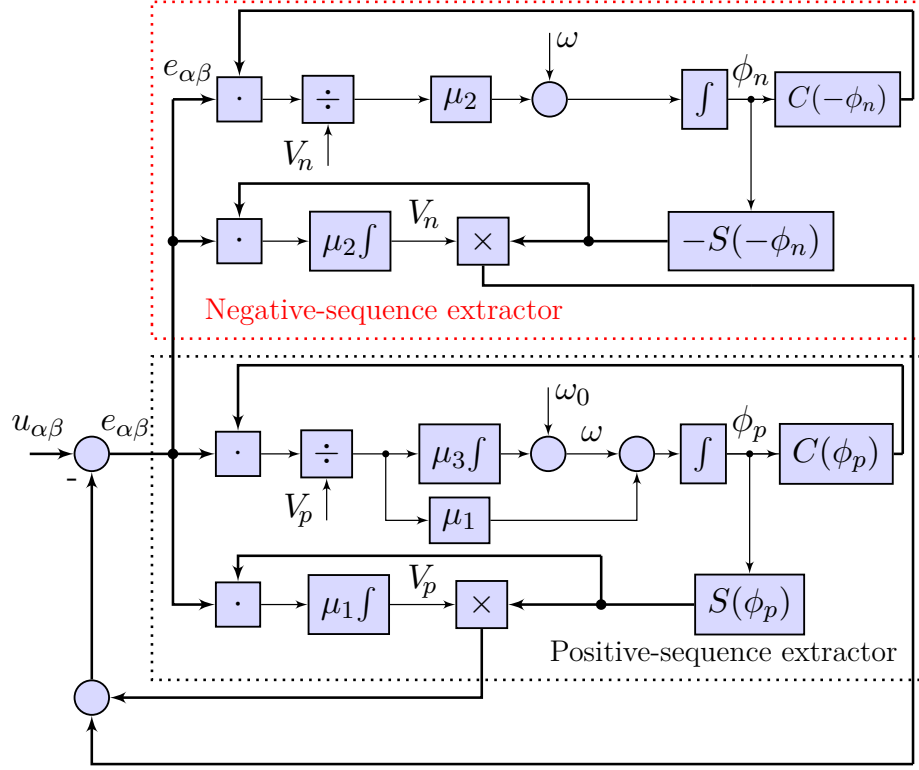


Figure 3.2 Three-phase ePLL of [3] (simplified UTSP of [7])

sequence components. As shown in [3], the LTI model of the positive-sequence extractor can be derived as

$$\begin{cases} \dot{x}_{1\text{pll}} = \mu_1 (u_\alpha - x_{1\text{pll}}) - \omega x_{2\text{pll}} \\ \dot{x}_{2\text{pll}} = \mu_1 (u_\beta - x_{2\text{pll}}) + \omega x_{1\text{pll}} \end{cases}, \quad \begin{bmatrix} x_{1\text{pll}} \\ x_{2\text{pll}} \end{bmatrix} = V_p \begin{bmatrix} \cos(\phi_p) \\ \sin(\phi_p) \end{bmatrix}. \quad (3.4)$$

In (3.4), $u_{\alpha\beta} = v_{s_{\alpha\beta}}$ is the input, and μ_1 is a setting parameter, and V_p and ϕ_p are estimations of amplitude and phase angle of the positive-sequence component, respectively. With $\xi = L_t(L_g + L_t)^{-1}$, $\sigma = L_g(L_g + L_t)^{-1}$, and $\tau = (R_g L_t - L_g R_t)(L_g + L_t)^{-1}$, substituting (3.3) in (3.4) results in

$$\begin{aligned} \dot{x}_{1\text{pll}} &= \mu_1 (\xi v_{g_\alpha} + \sigma v_{t_\alpha} + \tau i_\alpha) - \mu_1 x_{1\text{pll}} - \omega x_{2\text{pll}} \\ \dot{x}_{2\text{pll}} &= \mu_1 (\xi v_{g_\beta} + \sigma v_{t_\beta} + \tau i_\beta) - \mu_1 x_{2\text{pll}} + \omega x_{1\text{pll}}. \end{aligned} \quad (3.5)$$

3.2.4 Full LTI Modeling

Combining (3.5), (3.3) and (3.1) leads to the full model of the system, including the converter and the PLL, in weak grid conditions in state-space form

$$\dot{x}_p = A_p x_p + B_p u + F_p W, \quad y = C_p x_p, \quad (3.6)$$

where $x_p = [i_\alpha \ i_\beta \ x_{1_{pll}} \ x_{2_{pll}}]^T$, $u = [u_\alpha \ u_\beta]^T = [v_{t_\alpha} \ v_{t_\beta}]^T$, $W = [v_{g_\alpha} \ v_{g_\beta}]^T$, and

$$A_p = \begin{bmatrix} -\frac{R_t + \tau}{L_t} & 0 & 0 & 0 \\ 0 & -\frac{R_t + \tau}{L_t} & 0 & 0 \\ \mu_1 \tau & 0 & -\mu_1 & -\omega \\ 0 & \mu_1 \tau & \omega & -\mu_1 \end{bmatrix}, B_p = \begin{bmatrix} \frac{1 - \sigma}{L_t} & 0 \\ 0 & \frac{1 - \sigma}{L_t} \\ \mu_1 \sigma & 0 \\ 0 & \mu_1 \sigma \end{bmatrix}$$

$$C_p = [I_{2 \times 2} \ 0_{2 \times 2}], \ F_p = \begin{bmatrix} -\xi/L_t & 0 & \mu_1 \xi & 0 \\ 0 & -\xi/L_t & 0 & \mu_1 \xi \end{bmatrix}^T.$$

The system in (3.6) is controllable and observable for all system parameters and is minimum phase.

3.3 Proposed Controller in Stationary Frame

Fig. 3.3 shows the structure of the proposed control strategy which is realized in the $\alpha\beta$ -frame and encompasses three distinctive subsystems; i) a reference generator to produce reference currents associated with the given set-points of the real and reactive powers, ii) a servo-compensator, and iii) a stabilizing state feedback controller. The servo-compensator provides tracking of sinusoidal reference signals and rejection of disturbance signals, and is designed so that it includes the dynamics of the errors [64]. The stabilizing controller, on the other hand, guarantees robust stability and robust performance of the closed-loop system. The entire control system gains are designed in this chapter based on an extension of the newly proposed optimal control approach of [79, 80].

3.3.1 Control Objectives

The following performance specifications are desired.

1. Asymptotic stability of the resultant closed-loop system.
2. Steady-state asymptotic tracking of sinusoidal reference signal $y_{ref} = [i_\alpha^*, i_\beta^*]^T$ in the presence of disturbances.
3. Robust closed-loop stability with respect to L-filter, grid impedance parameters and power set points.

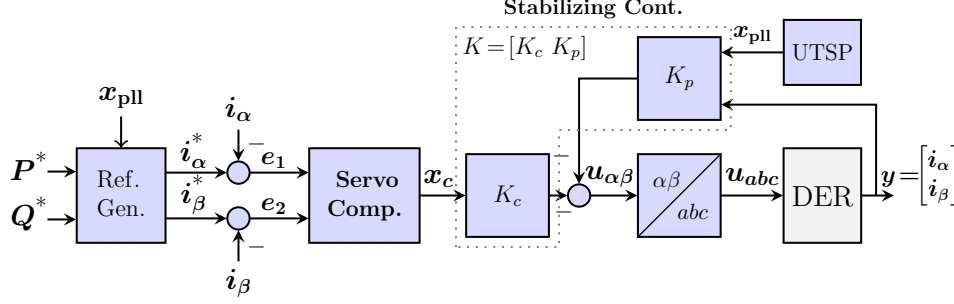


Figure 3.3 Structure of the proposed controller [6]

3.3.2 Reference Generation

The real and reactive powers at the output terminals are $P = 1.5(v_{s\alpha}i_\alpha + v_{s\beta}i_\beta)$ and $Q = 1.5(v_{s\beta}i_\alpha - v_{s\alpha}i_\beta)$. In weak grid conditions, the PCC voltage is polluted by switching noises. Thus, to have a smooth power measurement, $v_{s\alpha\beta}$ is replaced by its smooth version provided by the PLL, i.e., x_{pll} . This will lead to generation of the reference current signals as

$$i_{\alpha\beta}^* = \begin{bmatrix} i_\alpha^* \\ i_\beta^* \end{bmatrix} = \frac{2}{3} \frac{1}{V_p^2} \begin{bmatrix} x_{1pll} & x_{2pll} \\ x_{2pll} & -x_{1pll} \end{bmatrix} \begin{bmatrix} P^* \\ Q^* \end{bmatrix}, \quad (3.7)$$

where the asterisk superscript indicates the reference signals and $x_{(1,2)pll}$ are the state variables of the UTSP. Equations (3.5) and (3.7) reveal that couplings between the two control channels are introduced in the system at two points: PLL and reference generator. Therefore, adopting a multivariable control approach is recommended as used in this chapter. It is noted that if $Q^* = 0$, the reference generator does not introduce coupling terms and only PLL does.

3.3.3 Controller Design

Since the reference and disturbance signals are sinusoidal functions at frequency ω , the transformation defined by

$$p(D) = (D^2 + \omega^2), \quad (3.8)$$

(where $D = \frac{d}{dt}$ is the time derivative operator), nulls the error, $e = i_{\alpha\beta}^* - i_{\alpha\beta}$. A minimal realization of the compensator is

$$\dot{x}_c = A_c x_c + B_c e \quad (3.9)$$

where $A_c = \text{blkdiag}\{A_{c1}, A_{c2}\}$, $B_c = \text{blkdiag}\{B_{c1}, B_{c2}\}$ and

$$A_{c1} = A_{c2} = \begin{bmatrix} 0 & \omega \\ -\omega & 0 \end{bmatrix}, \quad B_{c1} = B_{c2} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}.$$

This implies that

$$\dot{x}_{c1_k} = \omega x_{c2_k}, \quad \dot{x}_{c2_k} = -\omega x_{c1_k} + \underbrace{(i_k^* - i_k)}_{e_k}, \quad (3.10)$$

for $(k = \alpha, \beta)$. Substituting (3.7) into (3.10) and (3.6) lead to

$$\begin{aligned} \dot{x}_{c1_\alpha} &= \omega x_{c2_\alpha}, \quad \dot{x}_{c1_\beta} = \omega x_{c2_\beta} \\ \dot{x}_{c2_\alpha} &= -\omega x_{c1_\alpha} + \frac{2}{3V_p^2} (P^* x_{1_{\text{pll}}} + Q^* x_{2_{\text{pll}}}) - i_\alpha \\ \dot{x}_{c2_\beta} &= -\omega x_{c1_\beta} + \frac{2}{3V_p^2} (P^* x_{2_{\text{pll}}} - Q^* x_{1_{\text{pll}}}) - i_\beta \\ \frac{di_\alpha}{dt} &= -\frac{R_t}{L_t} i_\alpha + \frac{1}{L_t} u_\alpha - \frac{1}{L_t} (\xi v_{g_\alpha} + \sigma u_\alpha + \tau i_\alpha) \\ \frac{di_\beta}{dt} &= -\frac{R_t}{L_t} i_\beta + \frac{1}{L_t} u_\beta - \frac{1}{L_t} (\xi v_{g_\beta} + \sigma u_\beta + \tau i_\beta) \\ \dot{x}_{1_{\text{pll}}} &= \mu_1 (\xi v_{g_\alpha} + \sigma u_\alpha + \tau i_\alpha) - \mu_1 x_{1_{\text{pll}}} + \omega x_{2_{\text{pll}}} \\ \dot{x}_{2_{\text{pll}}} &= \mu_1 (\xi v_{g_\beta} + \sigma u_\beta + \tau i_\beta) - \mu_1 x_{2_{\text{pll}}} - \omega x_{1_{\text{pll}}}. \end{aligned} \quad (3.11)$$

Equation set (3.11) describes the dynamics of total system with the state vector $x = [x_c^T, x_p^T]^T$, which includes the state variables of the MIMO servo-compensator, i.e., x_c , and those of the converter and PLL, i.e., x_p .

The full state feedback control law of

$$u = -Kx, \quad (3.12)$$

is employed to fully shape the modes of the closed loop system. The controller gain matrix K (2×8) comprises the two gain sub-matrices K_c (2×4) which operates on servo-compensator state, and K_p (2×4), such that $K = [K_c \ K_p]$.

The linear transformation (3.8) is applied to (3.11) to null the effect of v_g . The resultant set of equations is

$$\begin{aligned} \dot{z}_{c1_\alpha} &= \omega z_{c2_\alpha}, \quad \dot{z}_{c1_\beta} = \omega z_{c2_\beta} \\ \dot{z}_{c2_\alpha} &= -\omega z_{c1_\alpha} + \frac{2}{3V_p^2} (P^* z_{1_{\text{pll}}} + Q^* z_{2_{\text{pll}}}) - z_\alpha \\ \dot{z}_{c2_\beta} &= -\omega z_{c1_\beta} + \frac{2}{3V_p^2} (P^* z_{2_{\text{pll}}} - Q^* z_{1_{\text{pll}}}) - z_\beta \\ \frac{dz_\alpha}{dt} &= -\frac{R_t}{L_t} z_\alpha + \frac{1}{L_t} v_\alpha - \frac{1}{L_t} (\sigma v_\alpha + \tau z_\alpha) \\ \frac{dz_\beta}{dt} &= -\frac{R_t}{L_t} z_\beta + \frac{1}{L_t} v_\beta - \frac{1}{L_t} (\sigma v_\beta + \tau z_\beta) \\ \dot{z}_{1_{\text{pll}}} &= \mu_1 (\sigma v_\alpha + \tau z_\alpha) - \mu_1 z_{1_{\text{pll}}} + \omega z_{2_{\text{pll}}} \\ \dot{z}_{2_{\text{pll}}} &= \mu_1 (\sigma v_\beta + \tau z_\beta) - \mu_1 z_{2_{\text{pll}}} - \omega z_{1_{\text{pll}}}, \end{aligned} \quad (3.13)$$

where every z variable is defined as $(D^2 + \omega^2)$ operating on its corresponding x variable, e.g., $\dot{z}_{c1_\alpha} = (D^2 + \omega^2)\dot{x}_{c1_\alpha}$ and so on. As for the converter currents, they are defined as

$z_\alpha = (D^2 + \omega^2)i_\alpha$ and $z_\beta = (D^2 + \omega^2)i_\beta$. Also, the control inputs are transformed according to $v = (D^2 + \omega^2)u$ for both α and β axes. Now, (3.10) implies that $(D^2 + \omega^2)x_{c1_k} = z_{c1_k} = e_k$, $(D^2 + \omega^2)x_{c2_k} = z_{c2_k} = \frac{1}{\omega} \dot{e}_k$.

To effectively meet the design requirements, an optimal control solution is sought to minimize the cost function of

$$J = \int_0^\infty [q_1(e_\alpha^2 + e_\beta^2) + \frac{q_2}{\omega^2}(\dot{e}_\alpha^2 + \dot{e}_\beta^2) + q_5(z_\alpha^2 + z_\beta^2) + q_7 z_{1_{pll}}^2 + q_8 z_{2_{pll}}^2 + v_\alpha^2 + v_\beta^2] dt. \quad (3.14)$$

Minimization of J results in realizing a unique control law that corresponds to the weighting factors q_i . The weighting factors q_i are the elements on the main diagonal of the positive semi-definite matrix Q which are changed based on the design algorithm presented in [79].

Subsequently, the gain matrix, K , can be obtained, using Matlab *lqr* command. By using the system parameters given in Table 3.1 and changing the weighting gains q_1 through q_8 , an optimum design is achieved where

$$K = \left[\underbrace{\begin{bmatrix} -7.9e5 & -1.0e4 & -5.0e4 & -381 \\ 1.15e4 & 412.5 & -8.8e5 & -1e4 \end{bmatrix}}_{K_c} \mid \underbrace{\begin{bmatrix} 23.8 & 0.59 & 1.68 & -0.19 \\ -0.99 & 22.9 & 0.17 & 1.35 \end{bmatrix}}_{K_p} \right].$$

Fig. 3.4 shows the design procedure where q_1 has changed from 10^8 to 10^{13} , then q_2 is changed from 10 to $10^{7.3}$, then q_7 is varied from 0 to 10, and finally q_8 is changed from 0 to $10^{0.5}$. This places the closed-loop poles at $-297 \pm j632$, $-281 \pm j557$, $-683 \pm j293$ and $-505 \pm j45$ on the complex plane which yields fast and well-damped transient response.

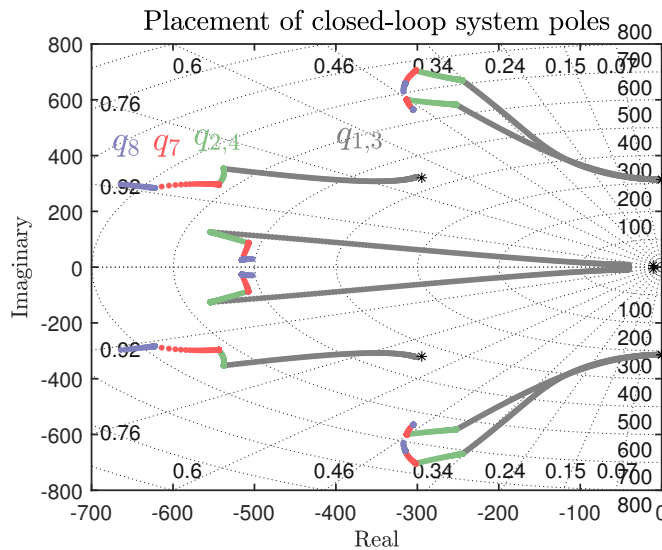


Figure 3.4 Controller design procedure using LQR method [6]

3.3.4 Robustness Analysis

Grid Impedance

The root locus of the eigenvalues of the closed-loop system $(A - BK)$ is shown in Fig. 3.5 where L_g is changed from 10 mH to 28 mH. By increasing L_g , the poles (marked in red color) move towards the imaginary axis and their damping factor decreases. For the sake of comparison, a controller which does not consider the dynamics of the PLL into its feedback system is also designed. The controller then is applied to the system in (3.11) and its closed-loop poles are calculated and plotted in Fig. 3.5(b) for different values of L_g (marked in gray color). The pairs of complex conjugate poles get very close to the imaginary axis as the grid gets weaker which indicates degradation in dynamic performance and eventually instability of the system for high values of L_g .

Real and Reactive Powers

As observed from (3.11), any change in power set-points affects the dynamic performance of the system. Therefore, it is necessary to evaluate the stability of the system against changes in operation set-points. Fig. 3.6 shows the evolution of the closed-loop poles when $P \in [0 \ 5]$ kW and $Q \in [0 \ 5]$ kVar. In Fig. 3.6, the darker shades correspond to set-point vectors close to $[0 \text{ kW}, 0 \text{ kVar}]$ and lighter shades of red indicate the set-point vectors closer to $[5 \text{ kW}, 5 \text{ kVar}]$. As it can be seen from Fig. 3.6, the poles of the closed-loop system are bounded in the left half-plane which ensures stability of the system for any given set-point vector within the specified range of P and Q .

3.4 Real/Reactive Power Constraints under Weak Grid Conditions

In weak grid conditions, there is a constraint on the minimum value of reactive power as also discussed in [31]. To understand and expand this, consider the system of Fig. 3.1 in which the complex power is $\mathbf{S} = 1.5\mathbf{V}_{\text{PCC}} \mathbf{I}^*$ where the phasors are used. By neglecting the grid resistance R_g , the VSC current satisfies $j\omega L_g \mathbf{I} = \mathbf{V}_{\text{PCC}} - \mathbf{V}_g$. Substituting this in the complex power equation yields

$$\mathbf{S} = P + jQ = \frac{3}{2} \left(\frac{|\mathbf{V}_{\text{PCC}}|^2 - \mathbf{V}_{\text{PCC}} \mathbf{V}_g}{-j\omega L_g} \right). \quad (3.15)$$

Defining $\mathbf{V}_{\text{PCC}} = V_R + jV_I$ and $K = \frac{2}{3}\omega L_g$, (3.15) results in

$$KQ + V_R V_g = V_R^2 + V_I^2, \quad -KP + V_I V_g = 0. \quad (3.16)$$

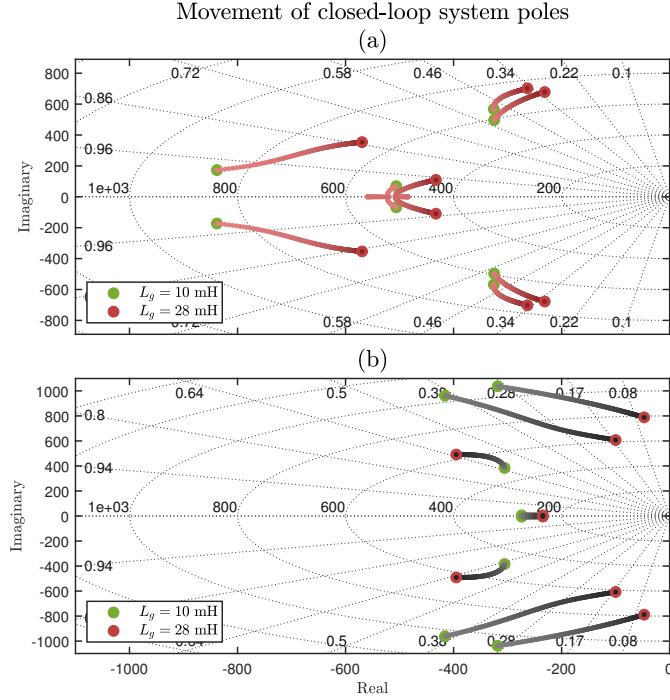


Figure 3.5 Evolution of the closed-loop systems poles with respect to changes in the grid inductance L_g when the PLL dynamics are (a) included (proposed) and (b) omitted from the feedback system (existing approach) [6]

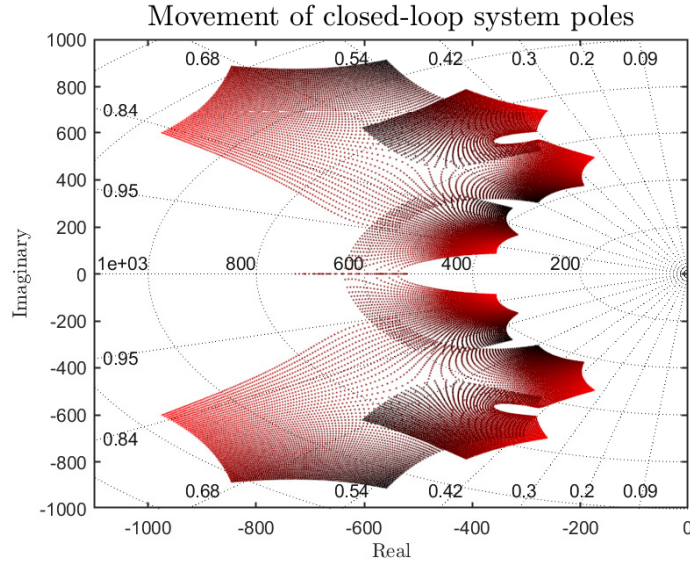


Figure 3.6 Evolution of the closed-loop poles with respect to changes in real and reactive powers ($P \in [0 \ 5]$ kW and $Q \in [0 \ 5]$ kVAr) [6]

Combining the two equations in (3.16) yields $V_R^2 - V_R V_g + (\frac{K^2 P^2}{V_g^2} - KQ) = 0$, which is a quadratic equation whose solution V_R is

$$V_R = \frac{1}{2} \left[V_g \pm \sqrt{V_g^2 - 4 \left(\frac{K^2 P^2}{V_g^2} - KQ \right)} \right]. \quad (3.17)$$

Since V_R is a real number, the term under square root must be non-negative, and therefore,

$$Q \geq \frac{K^2 P^2 - V_g^4/4}{V_g^2 K}. \quad (3.18)$$

The minimum value of reactive power is then obtained from the right side of inequality (3.18), i.e., $Q_{\min} = (K^2 P^2 - V_g^4/4)/(V_g^2 K)$, which corresponds to a unity short circuit ratio (SCR = 1). To prove this, consider the SCR definition as

$$SCR = \frac{S_{SC}}{S_{VSC}} = \frac{\frac{3}{2} V_{PCC}^2 / |Z_g|}{S_{VSC}} = \frac{1.5 V_{PCC}^2 / \omega_0 L_g}{S_{VSC}},$$

where S_{SC} , S_{VSC} , V_{PCC} and Z_g are the short-circuit apparent power of the grid, VSC apparent power, PCC voltage and the grid impedance, respectively. By considering $S_{VSC} = \sqrt{P^2 + Q_{\min}^2}$, we obtain

$$SCR = \frac{(V_R^2 + V_I^2)/K}{\sqrt{P^2 + Q_{\min}^2}}. \quad (3.19)$$

Replacing $V_I = KP/V_g$ from (3.16) and V_R from (3.17), into (3.19) results in $SCR = 1$.

The relationship between the minimum reactive power, active power and grid inductance for the study system of Table 3.1 is depicted in Fig. 3.7. It can easily be seen that for a given value of the real power, the minimum reactive power increases as the grid becomes weaker.

3.5 Performance Validation

The system parameters given in Table 3.1 are used in the simulations. Performance of the proposed controller is contrasted with a similar conventional multivariable controller which excludes the PLL state feedback terms. The conventional controller has a state feedback gain matrix K (2×6). By employing a similar LQR approach, its K is optimally designed in a way that its closed-loop poles are at $-290.5, -441.2 \pm j616$. These are close to those of the proposed controller and correspond to fast and well-damped responses.

3.5.1 Weak Grid

The system is started with both real and reactive powers set at zero. Next, the reactive power set-point is stepped up to 2.9 kVar at $t = 0.2$ s. Consequently, at $t = 0.3$ s, the real power is increased from 0 to 4 kW. This scenario is repeated for different values of SCR and the results are plotted in Fig. 3.8. The proposed controller shows fast dynamics with small overshoots across all different SCR values. This confirms the analysis carried out in

Table 3.1 Parameters of the Study System

parameters	symbol	value
dc bus voltage	V_{dc}	600 V
Grid phase voltage	v_s	120 V (rms)
System frequency	f	60 Hz
PWM switching frequency	f_{sw}	5 kHz
RL filter resistance	R_t	0.1 Ω
RL filter inductance	L_t	5 mH
Grid internal resistance	R_g	1 m Ω
Grid internal inductance	L_g	4, 10, 17, 28 mH
Short Circuit Ratio ($S \approx 5$ kVA)	SCR	7.0, 3.5, 2.0, 1.0

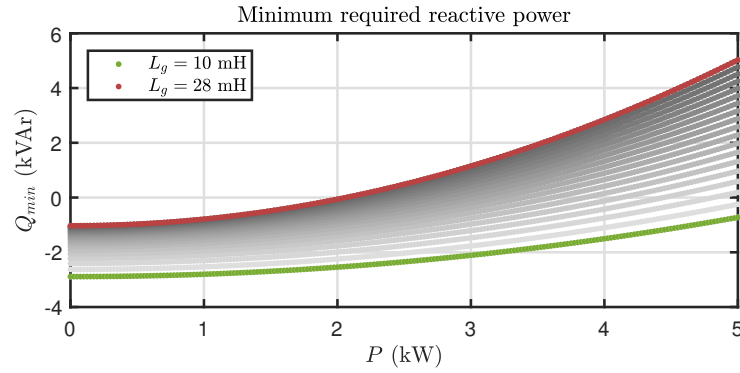


Figure 3.7 Relationship between the minimum required reactive power, real power and grid inductance [6]

Subsection 3.3.4. The responses of the conventional controller are shown in Fig. 3.9. While the response of the system for higher SCRs closely follows that of the proposed controller, overshoots and settling times significantly increase as the grid becomes weaker. The controller eventually fails to stabilize the system when the grid becomes extremely weak ($SCR=1.0$). More specifically, when SCR is decreased from 7.0 to 1.0, the proposed controller shows no overshoot and settling time increases from 15 ms to 50 ms. On the other hand, the response of the conventional controller becomes oscillatory and eventually unstable at $SCR=1.26$. Moreover, significant coupling between the two control channels is observed in the responses of the conventional controller under weaker grid conditions.

3.5.2 Unbalanced Conditions

While the VSC is injecting the real and reactive powers of 3 kW and 4 kVAr, respectively, a two-phase-to-ground fault occurs at the PCC at $t = 0.25$ s. This fault results in a voltage unbalance factor (VUF) of 80%, where VUF is the ratio of the magnitudes of negative-

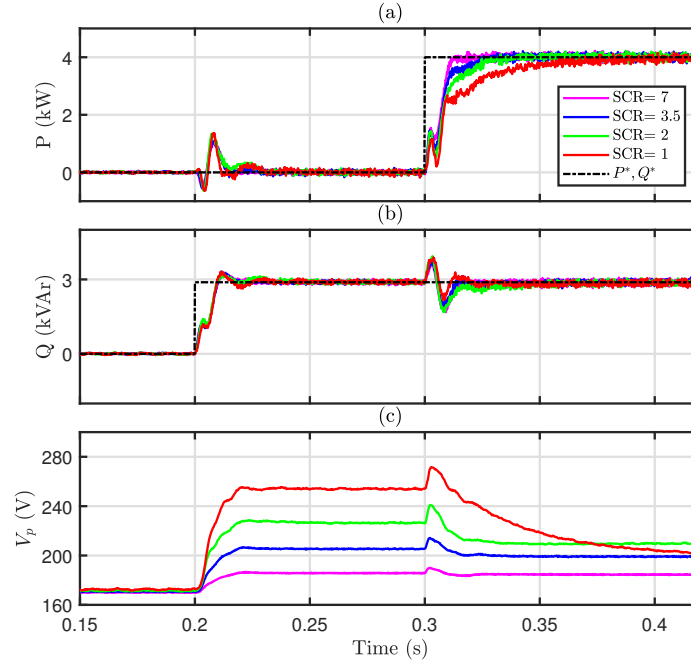


Figure 3.8 Transient response of the proposed controller to changes in real and reactive power set-points for different levels of grid stiffness. (a) real power, (b) reactive power, and (c) magnitude of PCC voltage [6]

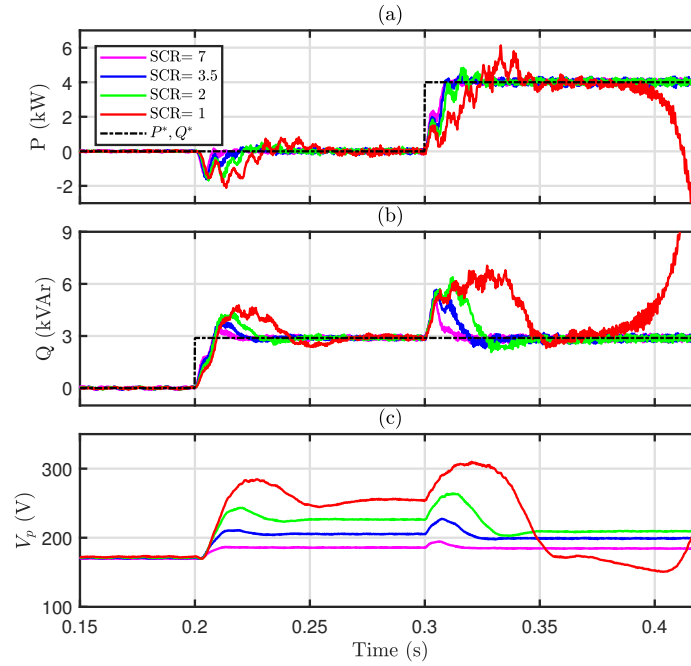


Figure 3.9 Transient response of the conventional controller to changes in real and reactive power set-points for different levels of grid stiffness. (a) real power, (b) reactive power, and (c) magnitude of PCC voltage [6]

to positive-sequence components of the PCC voltage. In this study, the grid inductance is 10 mH which corresponds to SCR=3.5. The conventional controller employs a PLL without negative-sequence extraction capability and thus, a double-frequency ripple component is

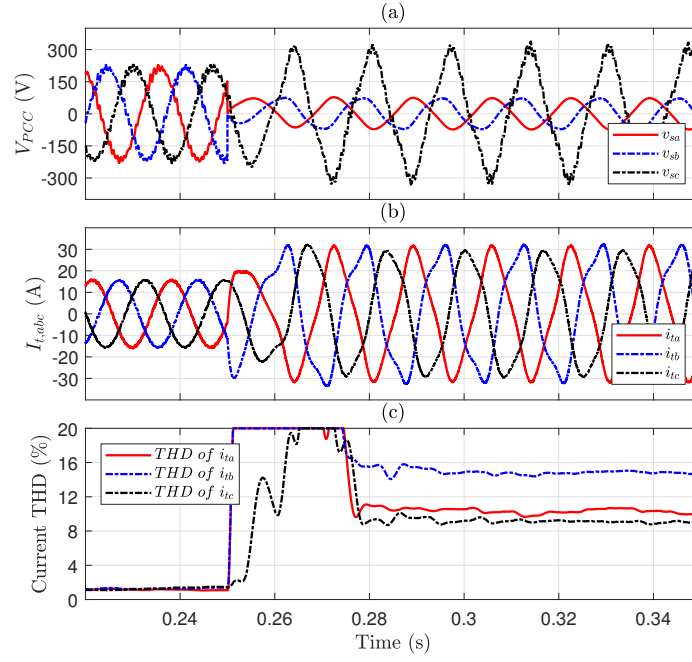


Figure 3.10 Response of conventional controller to an unbalanced fault: (a) PCC voltages, (b) VSC currents, and (c) THDs of VSC currents [6]

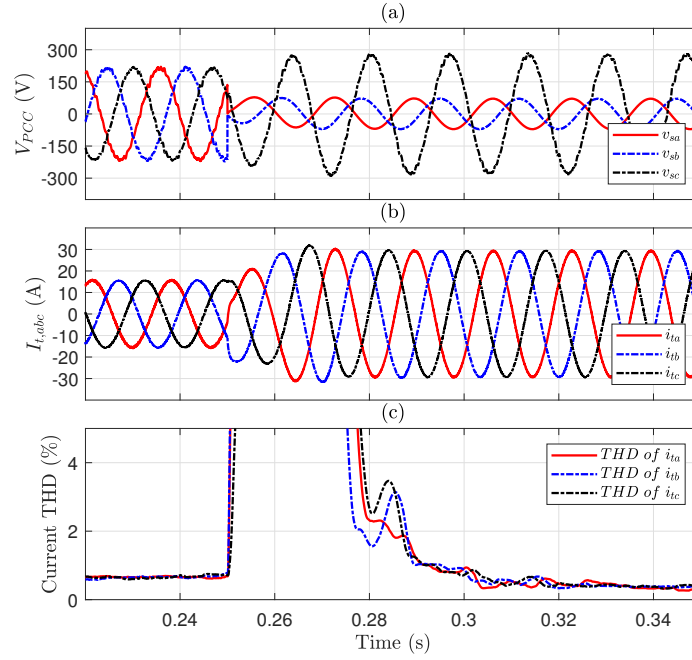


Figure 3.11 Response of proposed controller to an unbalanced fault: (a) PCC voltages, (b) VSC currents, and (c) THDs of VSC currents [6]

induced in its estimated voltage amplitude [7]. The UTSP, on the other hand, provides accurate and oscillation-free magnitudes of sequence components of the PCC voltage. Figs. 3.10 and 3.11 respectively show the responses of the conventional and the proposed controllers

with respect to the imposed asymmetric fault. It is observed that the proposed controller generates harmonic-free sinusoidal currents while the conventional controller produces distorted currents with high value of total harmonic distortion (THD), i.e., the current THD in conventional method is as high as 15%, while in the proposed controller, the current THD remains below 1%.

3.5.3 Grid Frequency Drift

In this case study, it is assumed that during normal operation, the grid frequency deviates from its nominal value within a range of ± 2 Hz, and that this event lasts for several hundredths of a second. In order to cope with such frequency drifts, the structure of the servo-compensator is modified as shown in Fig. 3.12, where the resonance controller is constantly updating with the estimated frequency provided by UTSP. Fig. 3.13 shows the response of the proposed controller when grid frequency increases by 2 Hz to reach 62 Hz and then drops to 58 Hz. As expected, the controller smoothly regulates the real and reactive powers with no steady-state offsets. For the sake of comparison, the response of the same controller with a fixed frequency compensator is also shown in Fig. 3.13 where both real and reactive powers diverge from their set-points under system frequency drifts.

3.6 Experimental Results

Parameters of the setup are summarized in Table 3.2. A prototype of the grid-connected VSC is implemented, shown in Fig. 3.14. The controller is implemented in OPAL-RT OP5700 FPGA-based real-time simulator. The voltages and currents are measured by the data acquisition unit, OP8660. The converter consists of a three-phase IGBT bridge on a Semikron MiniSkiiP with SKHI61 drivers and power rating of 40 kW. The dc-link voltage is set to 150 V provided by a PV emulator from Chroma. The regenerative grid simulator from Chroma

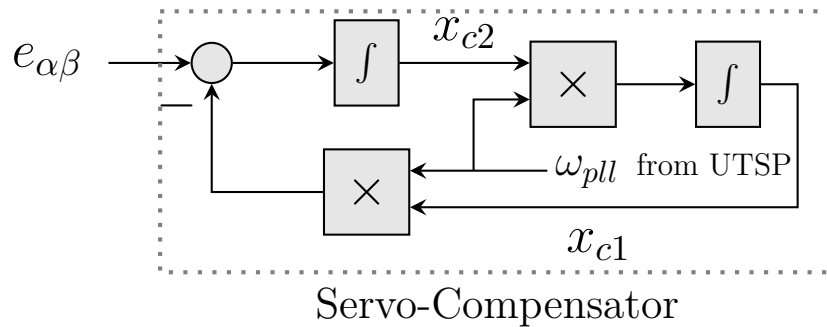


Figure 3.12 Structure of the frequency adaptive servo-compensator

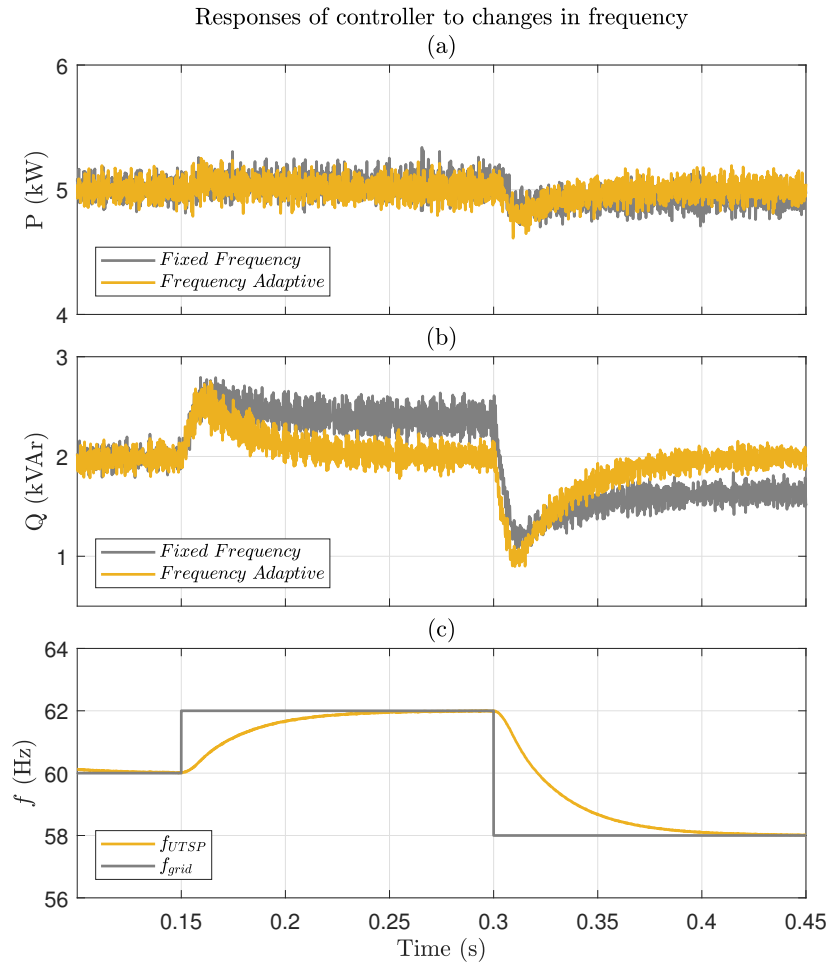


Figure 3.13 The controller responses with fixed and adaptive frequency compensators to changes in grid frequency: (a) real power, (b) reactive power, (c) grid frequency and the frequency determined by UTSP

(model 61830) is used as the grid with a line-to-line voltage of 52 V (rms) and is interfaced to the PCC via two inductances of $L_g = 5$ mH or $L_g = 20$ mH, to represent stiff or weak grid, respectively. The grid frequency is 60 Hz and the L -filter inductance is 5 mH. The data are saved on a PC and are plotted in MATLAB.

Stiff Grid

The grid inductance of $L_g = 5$ mH is considered to obtain a stiff grid (SCR=5.1, $S \approx 280$ VA). The reactive power, Q , is set at 200 VAR, and the real power set-point changes from 0 to 200 W. The dynamic responses of the proposed and conventional controllers are shown in Figs. 3.15 and 3.16, respectively, where the two controllers exhibit fast and almost identical performances.

Extremely Weak Grid

The grid is weakened by incorporating a grid inductance of $L_g = 20$ mH (SCR=1, $S \approx 350$ VA) and the following operating-point maneuvers are carried out for P and Q . First, the performance of both control systems with respect to a step change in reactive power is studied. For this purpose, Q is increased from 50 to 100 VAR while P is kept at 0. The responses of both proposed and conventional controllers are depicted in Figs. 3.17 and 3.18, respectively. It is observed that the proposed controller outperforms the conventional one in terms of performance criteria. In particular, when overshoot (M_p) and settling time (T_S) in conventional controller are 82% and 60 ms, respectively, the proposed controller exhibits a well-damped step response with $M_p = 25\%$ and $T_S = 20$ ms.

Second, while Q is set at 165 VAR, the real power is stepped up from 200 W to 300 W. The dynamic responses of the proposed and conventional controllers are respectively shown in Figs. 3.19 and 3.20. As observed, the proposed controller exhibits no noticeable overshoot and has a settling time of 25 ms whereas the conventional controller becomes unstable. It should be noted that, based on (3.18), the minimum required Q is determined as 161 VAR. Thus, the system is pushed to the verge of voltage instability, in particular, any further increase in P or decrease in Q will lead to system instability.

By comparing Figs. 3.17 and 3.19 with Figs. 3.18 and 3.20, it is seen that the proposed controller enhances dynamic performance. Furthermore, improved stability of the proposed controller becomes clear in Figs. 3.19 and 3.20, where the conventional controller becomes unstable while the proposed controller successfully regulates the power.

Table 3.2 Parameters of the Experimental Setup

Parameter	Symbol	Value
dc bus voltage	V_{dc}	150 V
Grid phase voltage	V_s	30 V (rms)
System frequency	f	60 Hz
PWM switching frequency	f_{sw}	5 kHz
RL filter inductance	L_t	5 mH
Grid internal inductance	L_g	5, 20 mH



Figure 3.14 A photograph of the experimental setup [6]

In an attempt to test the feasibility of implementing the presented controller on commercially available digital processors, the proposed controller was implemented on a TMS320F28335 digital signal processor (DSP) from Texas Instrument using *eZdspTMf28335* development card from Spectrum Digital. This DSP belongs to C2000 series which is widely used in industrial applications. To quantify the occupied processing time of the proposed controller, a general purpose input/output pin (GPIO84) is set at the beginning of the main program and is cleared at the last line. Thus, by reading the voltage of the dedicated GPIO, the running time of the algorithm can be visualized as a portion of the DSP sampling time. Figs. 3.22 and 3.23 depict the implementation of the code for setting/clearing of the GPIO and its corresponding output as visualized by the oscilloscope. As shown in Fig. 3.23, the running time of the main algorithm is around one-fourth of DSP sampling time (i.e., 200 μs). This clearly shows that the controller can be implemented on other commercially available processors even for sampling frequencies as high as 20 kHz. It also should be mentioned that the codes have not been optimized which could further reduce the computational burden.

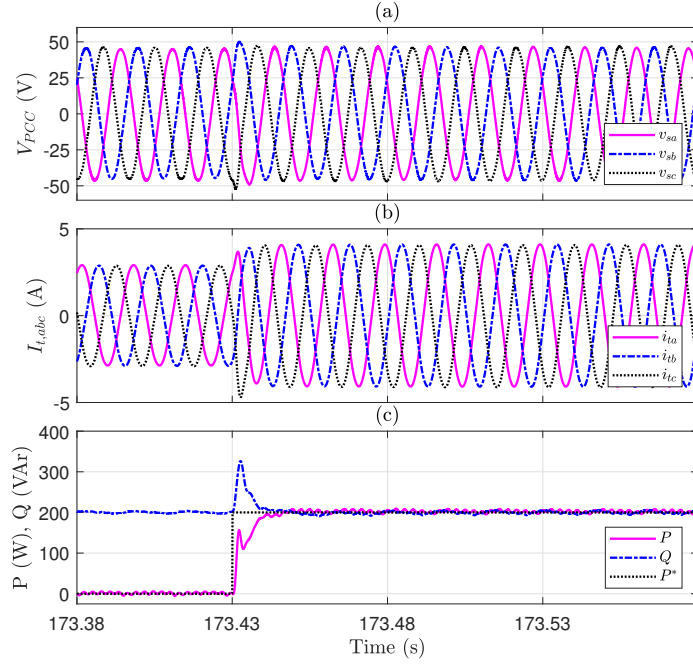


Figure 3.15 (Experimental results) Response of proposed controller to a step change in real power when the grid is stiff: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

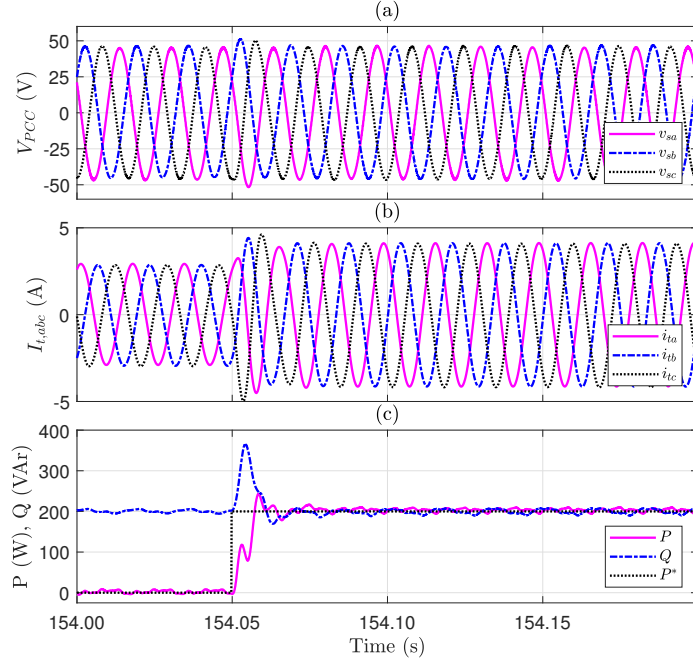


Figure 3.16 (Experimental results) Response of conventional controller to a step change in real power when the grid is stiff: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

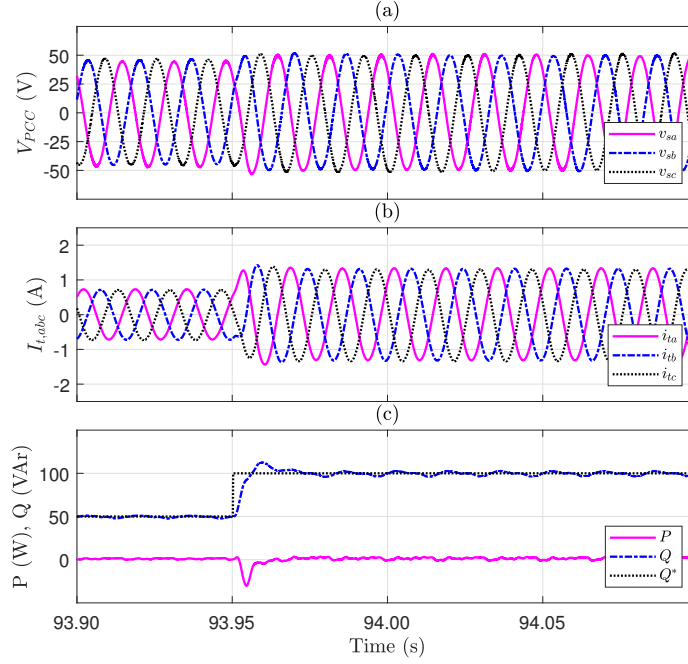


Figure 3.17 (Experimental results) Response of proposed controller to a step up in reactive power under extremely weak grid conditions: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

Current and voltage responses of the proposed and conventional controllers with respect to the same step change in real power are captured by a digital storing oscilloscope (DSO) and are depicted in Fig. 3.21. It should be mentioned that the results shown in this figure are obtained from an implementation of the controllers on an *f28335* DSP using *eZdspTM f28335* development card.

3.7 Conclusion

A new control approach in stationary reference frame is proposed for grid-connected voltage source converters (VSCs). The objective is to extend its stability margins in extremely weak grid conditions. Thus, the dynamics of the phase-locked loop (PLL) are integrated into the controller and its gains are designed using optimal control theory. The simulation results and experimental tests confirm the effectiveness of the proposed controller in improving the performance and stability of the system in extremely weak grid conditions. It is shown that the proposed controller improves the system performance. More specifically, at a given identical operating conditions, the following results are obtained.

- While the conventional controller's responses become increasingly oscillatory for SCR below 1.95 and become unstable at SCR=1.26, the proposed controller provides smooth and stable responses for SCR as low as 1.

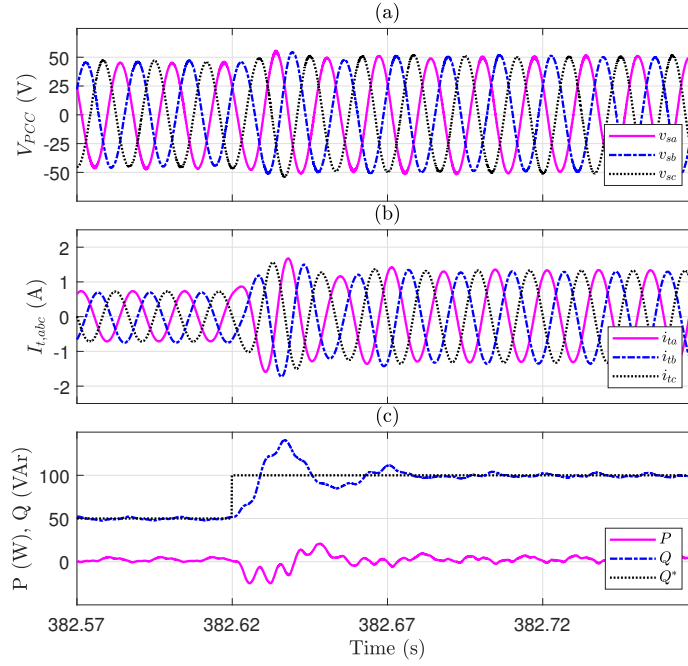


Figure 3.18 (Experimental results) Response of conventional controller to a step up in reactive power under extremely weak grid conditions: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

- The coupling between the two control channels is significantly increased in the conventional controller when SCR goes down.
- The proposed controller provides much cleaner sinusoidal currents for unbalanced and weak grid conditions.

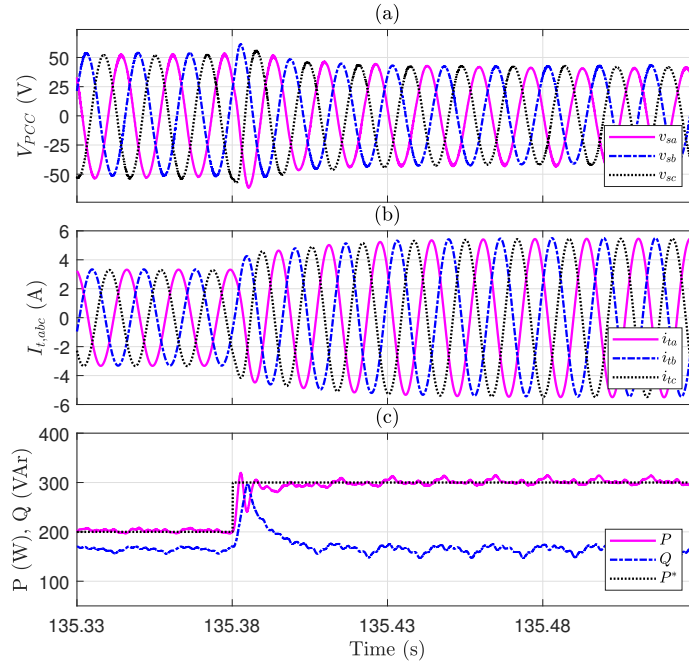


Figure 3.19 (Experimental results) Response of proposed controller to a step up in real power under extremely weak grid conditions: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

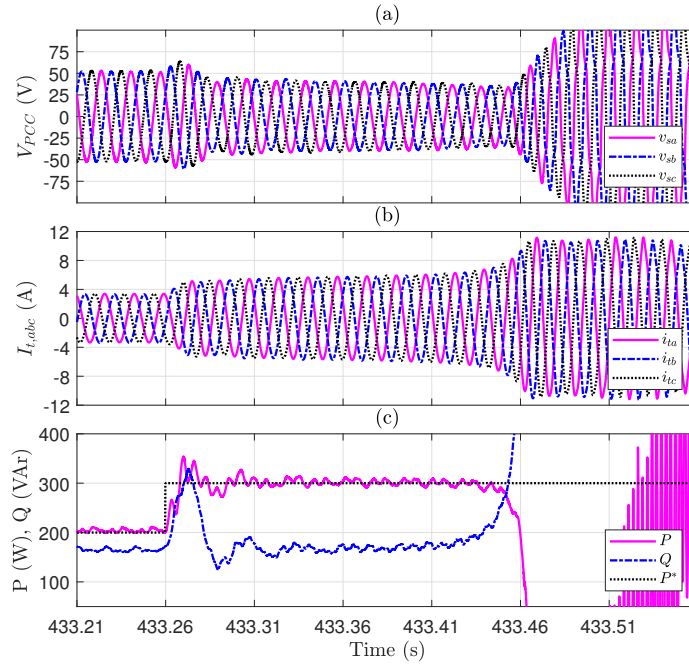


Figure 3.20 (Experimental results) Response of conventional controller to a step up in real power under extremely weak grid conditions: (a) PCC voltages, (b) VSC currents, and (c) real and reactive powers [6]

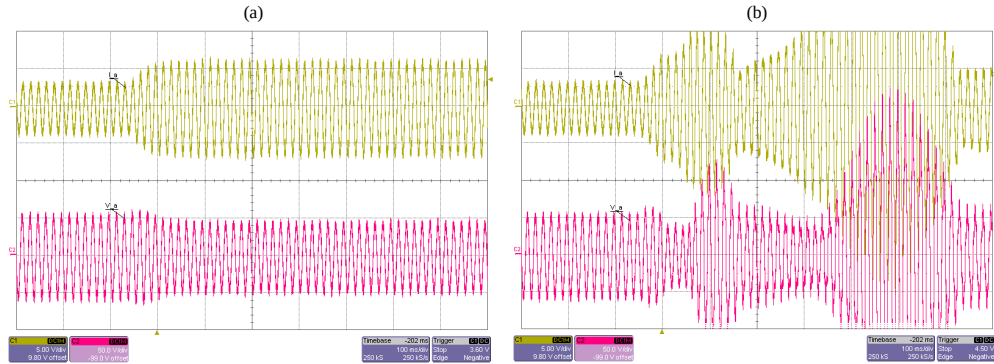


Figure 3.21 (Experimental results) Current and voltage response of controllers (shown in green and pink, respectively) to a step up in real power under extremely weak grid conditions as captured by DSO: (a) proposed, and (b) conventional controller [6]

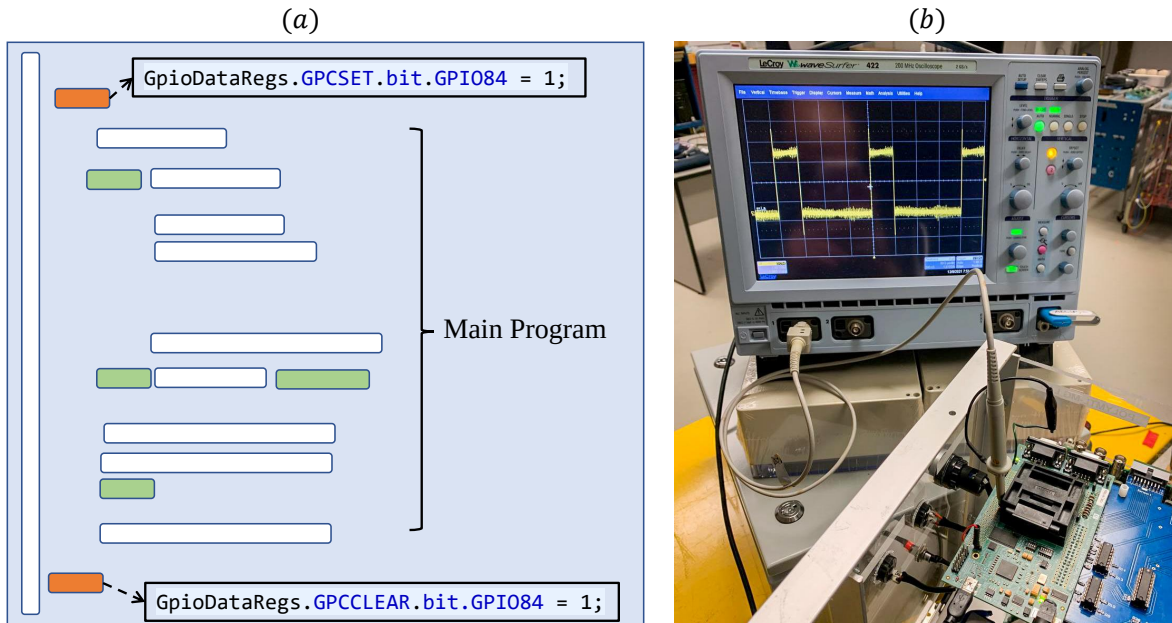


Figure 3.22 Evaluating the consumed processing capacity of the DSP by the proposed controller. (a) schematic of code implementation (b) measuring the voltage of the output pin GPIO84

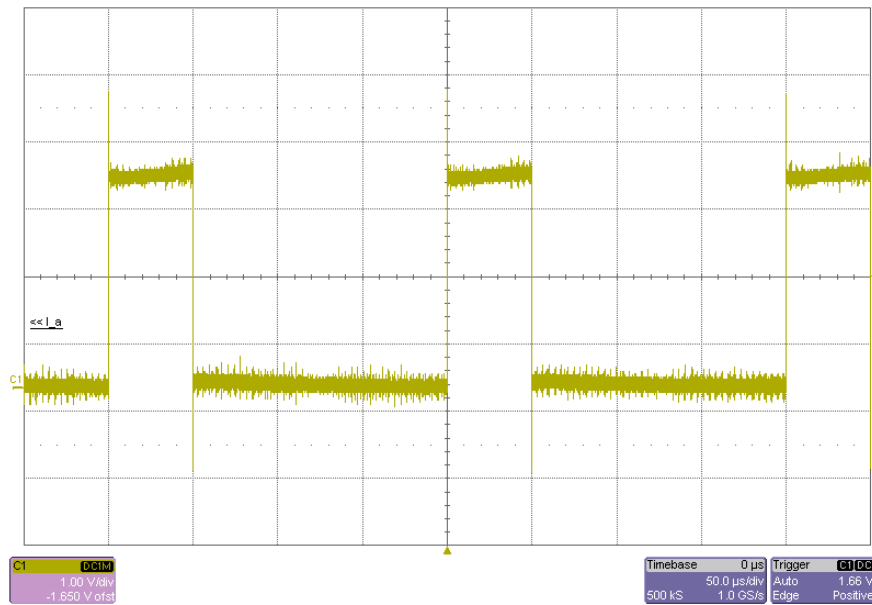


Figure 3.23 Visualized running-time signal by oscilloscope

CHAPTER 4 ENHANCED PERFORMANCE OF VOLTAGE SOURCE CONVERTERS UNDER WEAK AND UNBALANCED GRID CONDITIONS

4.1 Background

As previously mentioned in this thesis, voltage source converters (VSCs) employ vector current controllers (VCCs) implemented in dq-frame in grid following applications. In these methods a phase-locked loop (PLL) is usually used to extract the phase angle of voltage at the point of common coupling (PCC). The estimation of phase angle is then used for synchronous reference frame (SRF) transformation which facilitates the control design by transforming ac signals into constant values. However, despite the simplicity of their control structure, under weak grid conditions, VCCs suffer from stability degradation [23, 68, 71, 72, 81]. This problem is attributed to the PLL's nonlinear dynamics, coupling effects, and interactions with other control loops. To address this problem a broad range of solutions are sought in the literature. E.g., decreasing the bandwidth of PLL has been suggested in [42]. This approach, however, can violate dynamic performance criteria specially when fast dynamics are required [82]. Attempting to improve the performance of conventional PLL-based VCC, some papers have proposed direct power control (DPC) methods which use some mathematical operations and filtering stages to implement VCC [29–31]. The study in Chapter 2 however, reveals that deployment of such filters produces the same dynamics as does the PLL and thus, stability is equally affected. In a different approach, integration of PLL into the current control loop has been proposed in [32, 33] where the system performance is substantially improved under weak grid conditions. While most of the works in the literature belong to the class of SRF-based VCCs, some recent studies suggest that resonant-based controllers realized in stationary frame are less affected by PLL dynamics [34–36]. In that regard, Chapter 3 incorporated the state variables of a robust ePLL into the dynamic equations of a weak grid system to derive a linear model in stationary frame. The closed-loop system is then shaped using an optimized multivariable feedback approach which extends the stability margin of the grid-tied VSC down to the unity short circuit ratio ($SCR=1$).

While resonant-based controllers in general exhibit a more robust performance against weak grid conditions, they are less adopted in applications where dc voltage control is required under voltage imbalance. The reason for this, is that unlike SRF-based system models, realization of dc-link dynamics in stationary frame results in a nonlinear time-varying model which can not be directly used with linear control approaches. To this end, multi-loop nested control structures with voltage and current sequence decomposition are widely studied in the

literature [45, 47, 48]. These methods, however, require implementation of notch filters and sequential component extractors which on the other hand complicate the control algorithm, introduce nonlinear dynamics to the system, and jeopardize system stability under weak grid conditions. Trying to acquire unbalanced power control in single stationary frame, [49] has proposed using of vector quadrature component (i.e., variable vector delayed by 90° electrical degrees) for calculating current reference signals. The same approach has been adopted in [50, 51], where voltage quadrature components are deployed for implementing DPC methods under unbalanced conditions. However, despite their improved performance, these methods incorporate control efforts (i.e., rectifier outputs voltages) for deriving current reference signals which requires accurate estimation of system parameters and can lead to steady state errors if not precisely determined. Moreover all these methods entail rather complex structures which acquire reference currents by solving matrix equations. This particularly may lead to errors and cause stability issues when the voltages are heavily distorted under weak grid conditions. Based on the existing literature, it is obvious that there is a lack of robust dc voltage control strategies in stationary frame in which high performance current control loops can be implemented. The study in this chapter tries to fill this gap by proposing a novel control structure which systematically removes dc-link voltage oscillations in an unbalanced grid while offering enhanced performance and improved stability under weak grid conditions. To the best of our knowledge this is the first time that such controller with distinctly simple structure is proposed in stationary frame. First the work in [6] is extended to incorporate the full dynamics (including negative sequence loop) of the newly introduced ePLL (proposed in [3]) into the characteristic equations of a weak grid system and derives a linear time-invariant (LTI) model. A robust multivariable control technique with optimally designed feedback gains is then employed to improve performance and extend the stability margin of the current controller under weak grid conditions. Second, the dynamics of the dc-link are revisited and the effects of dc voltage control loops are comprehensively studied. Based on the identified dynamics of the VSC, a nested multiloop control structure is proposed consisting of two parallel sub-controllers; an integrator-based as well as a resonant-based compensator. The constraints for suppressing third-order current harmonics are derived and it is shown that by employing the proposed control strategy, steady-state decoupling of the two dc-link sub-systems is achieved. It is also revealed that absorption of reactive power adversely affects the stability of the outer control loop; a fact that has not been identified before. The end result of this study is a new controller which not only provides a ripple-free dc voltage under severe grid voltage imbalance but also improves system stability under weak grid conditions. Finally software simulations and hardware-in-the-loop tests are carried out to verify the enhanced performance of the proposed controller.

4.2 System Modeling

4.2.1 System Description

The schematic diagram of the studied system and the associated controller are shown in Fig. 4.1. The system comprises of a three-phase VSC connected to an ideal grid through the series filter impedance $L_t + j\omega L_t$ and the grid impedance $R_g + j\omega L_g$. The dc side encompasses the dc-link capacitor C_{dc} paralleled with the resistor R and the slow-time-varying current source P_{in} whose dynamic contains only constant and step-wise functions. The controller has a nested structure and is implemented in stationary frame where voltages and currents are measured at PCC and are transferred to the ePLL as well as the control loop through $abc/\alpha\beta$ transformation blocks. The dc-link voltage is measured and its squared value is deployed in the outer control loop. The controller output then is re-transformed to abc -frame where sinusoidal pulse width modulation (SPWM) is employed to produce the gating signals.

4.2.2 PLL-Integrated Model of AC Sub-System

The characteristic equation of the ac side of Fig. 4.1 can be expressed in $\alpha\beta$ -frame as

$$\frac{di_{\alpha\beta}}{dt} = -\frac{R_t}{L_t}i_{\alpha\beta} + \frac{1}{L_t}u_{t\alpha\beta} - \frac{1}{L_t}v_{s\alpha\beta} \quad (4.1)$$

where $i_{\alpha\beta}$, $v_{s\alpha\beta}$ and $u_{t\alpha\beta}$ are the VSC output current, PCC voltage and the VSC terminal voltage, respectively. Moreover, as shown in [6], the PCC voltage can be expressed based on VSC output voltage and current as well as the ideal grid voltage, i.e., $v_{g\alpha\beta}$ such that

$$v_{s\alpha\beta} = \frac{L_t}{L_g + L_t} v_{g\alpha\beta} + \frac{L_g}{L_g + L_t} v_{t\alpha\beta} + \frac{(R_g L_t - L_g R_t) i_{\alpha\beta}}{L_g + L_t}, \quad (4.2)$$

Furthermore, the linear time-invariant (LTI) model of the proposed ePLL in [3] is:

$$\begin{aligned} \dot{x}_{1pll} &= \mu_1 (u_\alpha - x_{1pll}) - \omega x_{2pll} - \mu_1 x_{3pll} \\ \dot{x}_{2pll} &= \mu_1 (u_\beta - x_{2pll}) + \omega x_{1pll} - \mu_1 x_{4pll} \\ \dot{x}_{3pll} &= \mu_2 (u_\alpha - x_{3pll}) + \omega x_{4pll} - \mu_2 x_{1pll} \\ \dot{x}_{4pll} &= \mu_2 (u_\beta - x_{4pll}) - \omega x_{3pll} - \mu_2 x_{2pll}, \end{aligned} \quad (4.3)$$

where, $u_{\alpha\beta} = v_{s,\alpha\beta}$ and $\mu_1 = \mu_2 = 300$ are some gains in the ePLL structure. Furthermore,

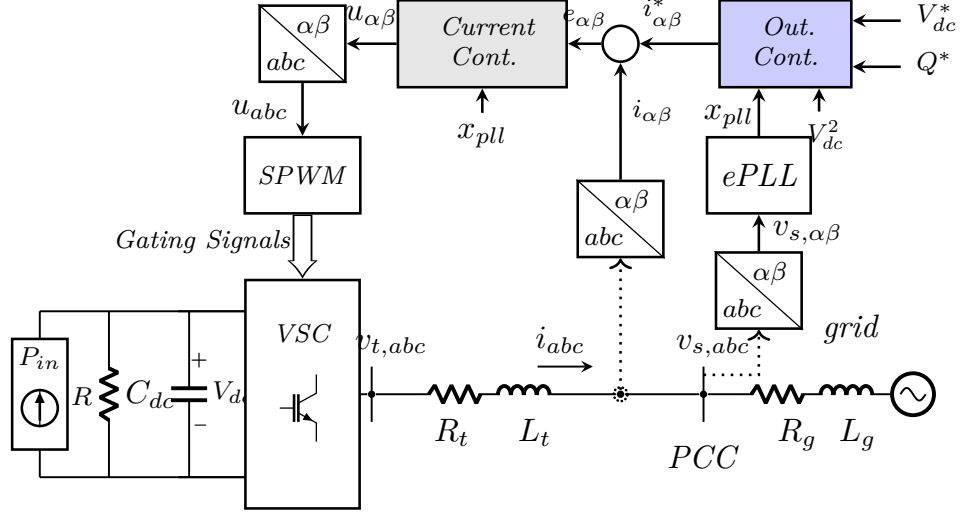


Figure 4.1 single-line diagram of the grid-connected VSC and its associated control system

in [3] the state variables of the ePLL are stated as

$$\begin{aligned} x_{1pll} &= V_p \cos(\phi_p), \quad x_{2pll} = V_p \sin(\phi_p) \\ x_{3pll} &= V_n \cos(\phi_n), \quad x_{4pll} = -V_n \sin(\phi_n) \end{aligned}$$

where $V_{p,n}$ and $\phi_{p,n}$ are estimations of amplitude and phase angle of the positive and negative sequences of voltage, respectively. Substituting (4.2) into (4.3) with $\xi = L_t(L_g + L_t)^{-1}$, $\sigma = L_g(L_g + L_t)^{-1}$, and $\tau = (R_g L_t - L_g R_t)(L_g + L_t)^{-1}$, readily results in

$$\begin{aligned} \dot{x}_{1pll} &= \mu_1 (\xi v_{grid,\alpha} + \sigma v_{t\alpha} + \tau i_\alpha) - \mu_1 x_{1pll} - \omega x_{2pll} - \mu_1 x_{3pll} \\ \dot{x}_{2pll} &= \mu_1 (\xi v_{grid,\beta} + \sigma v_{t\beta} + \tau i_\beta) - \mu_1 x_{2pll} + \omega x_{1pll} - \mu_1 x_{4pll} \\ \dot{x}_{3pll} &= \mu_2 (\xi v_{grid,\alpha} + \sigma v_{t\alpha} + \tau i_\alpha) - \mu_2 x_{3pll} + \omega x_{4pll} - \mu_2 x_{1pll} \\ \dot{x}_{4pll} &= \mu_2 (\xi v_{grid,\beta} + \sigma v_{t\beta} + \tau i_\beta) - \mu_2 x_{4pll} - \omega x_{3pll} - \mu_2 x_{2pll}. \end{aligned} \tag{4.4}$$

Thus the model of the ac sub-system by inclusion of ePLL dynamics can be represented in state-space form

$$\dot{x}_{ac} = A_{ac} x_{ac} + B_{ac} u + F_{ac} W, \quad y = C_{ac} x_{ac}, \tag{4.5}$$

where $x_{ac} = \underbrace{[i_\alpha \ i_\beta]^T}_{x_p^T} \underbrace{[x_{1pll} \ x_{2pll} \ x_{3pll} \ x_{4pll}]^T}_{x_{pll}^T}$, $u = [v_{t\alpha} \ v_{t\beta}]^T$, $W = [v_{g\alpha} \ v_{g\beta}]^T$, and

$$A_{ac} = \begin{bmatrix} -\frac{R_t + \tau}{L_t} & 0 & 0 & 0 & 0 & 0 \\ 0 & -\frac{R_t + \tau}{L_t} & 0 & 0 & 0 & 0 \\ \mu_1 \tau & 0 & -\mu_1 & -\omega & -\mu_1 & 0 \\ 0 & \mu_1 \tau & \omega & -\mu_1 & 0 & -\mu_1 \\ \mu_2 \tau & 0 & -\mu_2 & 0 & -\mu_2 & \omega \\ 0 & \mu_2 \tau & 0 & -\mu_2 & -\omega & -\mu_2 \end{bmatrix},$$

$$B_{ac} = \begin{bmatrix} \frac{1-\sigma}{L_t} & 0 & \mu_1 \sigma & 0 & \mu_2 \sigma & 0 \\ 0 & \frac{1-\sigma}{L_t} & 0 & \mu_1 \sigma & 0 & \mu_2 \sigma \end{bmatrix}^T, \quad C_{ac} = [I_{2 \times 2} \quad 0_{2 \times 4}].$$

4.2.3 DC-Link Model under Voltage Imbalance

The dynamic equation based on the power balance between the dc and the ac sub-systems can be considered as

$$P_{in} - \frac{V_{dc}^2}{R} - \frac{1}{2} C \frac{d(V_{dc}^2)}{dt} = P_t. \quad (4.6)$$

The right-hand side of (4.6) corresponds to the instantaneous converted real power at the VSC output terminals P_t which under ideal grid conditions and by neglecting the power losses in the ac-subsystem, remains constant throughout the system. Such condition, however, does not hold under voltage imbalance which will be discussed in what follows.

Consider a set of three phase and unbalanced PCC voltages

$$\begin{aligned} v_a(t) &= V_p \cos(\omega t + \delta_{vp}) + V_n \cos(\omega t + \delta_{vn}) \\ v_b(t) &= V_p \cos(\omega t + \delta_{vp} - 2\frac{\pi}{3}) + V_n \cos(\omega t + \delta_{vn} + 2\frac{\pi}{3}) \\ v_c(t) &= V_p \cos(\omega t + \delta_{vp} + 2\frac{\pi}{3}) + V_n \cos(\omega t + \delta_{vn} - 2\frac{\pi}{3}), \end{aligned}$$

where V_p , V_n , δ_{vp} and δ_{vn} are respectively the magnitudes and phase angles of positive and negative voltage components and ω is the angular frequency. Using Clarke transformation, the PCC voltages are expressed in $\alpha\beta$ frame

$$\begin{aligned} v_{s\alpha}(t) &= V_p \cos(\omega t + \delta_{vp}) + V_n \cos(\omega t + \delta_{vn}) \\ v_{s\beta}(t) &= V_p \sin(\omega t + \delta_{vp}) - V_n \sin(\omega t + \delta_{vn}). \end{aligned} \quad (4.7)$$

In (4.1), $i_{\alpha\beta}$ contains both negative and positive components such that

$$\begin{aligned} i_{\alpha}(t) &= I_p \cos(\omega t + \delta_{ip}) + I_n \cos(\omega t + \delta_{in}) \\ i_{\beta}(t) &= I_p \sin(\omega t + \delta_{ip}) - I_n \sin(\omega t + \delta_{in}), \end{aligned} \quad (4.8)$$

where δ_{ip} and δ_{in} are the positive and negative current phase angles corresponding to the

positive and negative components, respectively. Considering (4.7), (4.8) and the definition of real and reactive powers at PCC ($P = \frac{3}{2}(v_{s\alpha}i_\alpha + v_{s\beta}i_\beta)$, $Q = \frac{3}{2}(v_{s\beta}i_\alpha - v_{s\alpha}i_\beta)$), it can easily be shown that P and Q are time-varying functions consisting of a constant as well as an oscillatory part, such that

$$P(t) = \bar{P} + \tilde{P}(t), \quad Q(t) = \bar{Q} + \tilde{Q}(t),$$

where

$$\begin{aligned} \bar{P} &= \frac{3}{2} (V_p I_p \cos(\delta_{vp} - \delta_{ip}) + V_n I_n \cos(\delta_{in} - \delta_{vn})) \\ \tilde{P} &= \frac{3}{2} (V_p I_n \cos(2\omega t + \delta_{vp} + \delta_{in}) + V_n I_p \cos(2\omega t + \delta_{vn} + \delta_{ip})) \\ \bar{Q} &= \frac{3}{2} (V_p I_p \sin(\delta_{vp} - \delta_{ip}) + V_n I_n \sin(\delta_{in} - \delta_{vn})) \\ \tilde{Q} &= \frac{3}{2} (V_p I_n \sin(2\omega t + \delta_{vp} + \delta_{in}) - V_n I_p \sin(2\omega t + \delta_{vn} + \delta_{ip})) \end{aligned}$$

In order to attenuate real power oscillations at PCC, it is sufficient to adjust the current vector to satisfy the following conditions; (i) $V_n I_p = V_p I_n$ and (ii) $\delta_{vn} + \delta_{ip} = \delta_{vp} + \delta_{in} - \pi$. However, to fully mitigate the oscillations of dc-link voltage, such condition should be realized at the output terminals of the VSC, i.e., $v_{t\alpha\beta}$.

Neglecting the resistive term R_t , VSC terminal voltage can be expressed based on symmetrical components of PCC voltage and current as

$$v_{t\alpha\beta} = (V_p e^{j(\omega t + \delta_{vp})} + V_n e^{-j(\omega t + \delta_{vn})}) + (L_t \cdot j\omega) (I_p e^{j(\omega t + \delta_{ip})} - I_n e^{-j(\omega t + \delta_{in})}).$$

Thus the real power at VSC terminals can be formulated as

$$\begin{aligned} P_t &= \frac{3}{2} \Re[V_p I_p e^{j(\delta_{vp} - \delta_{ip})} + V_n I_n e^{j(\delta_{in} - \delta_{vn})} + L_t \omega I_p^2 e^{j(\frac{\pi}{2})} - L_t \omega I_n^2 e^{j(\frac{\pi}{2})}] \\ &+ \frac{3}{2} \Re[V_p I_n e^{j(2\omega t + \delta_{vp} + \delta_{in})} + V_n I_p e^{-j(2\omega t + \delta_{vn} + \delta_{ip})} \\ &+ L_t \omega I_p I_n [e^{j(2\omega t + \delta_{ip} + \delta_{in} + \frac{\pi}{2})} + e^{-j(2\omega t + \delta_{ip} + \delta_{in} + \frac{\pi}{2})}]]. \end{aligned} \quad (4.9)$$

Based on (4.9), the oscillatory part of real power at the VSC terminal is equal to PCC real power oscillation added with a double-frequency term such that

$$\tilde{P}_t = \tilde{P}_s - 3 L_t \omega I_p I_n \sin(2\omega t + \delta_{ip} + \delta_{in}). \quad (4.10)$$

Eq. (4.10) reveals strong non-linear coupling terms between symmetrical components of current. Moreover, it suggests that by oscillating the real power at PCC with the indicated double frequency term (i.e., $\tilde{P}_s = 3 L_t \omega I_p I_n \sin(2\omega t + \delta_{ip} + \delta_{in})$), one can obtain an oscillation free real power at VSC output which in turn neutralizes dc-link voltage ripples.

4.3 Proposed Controller

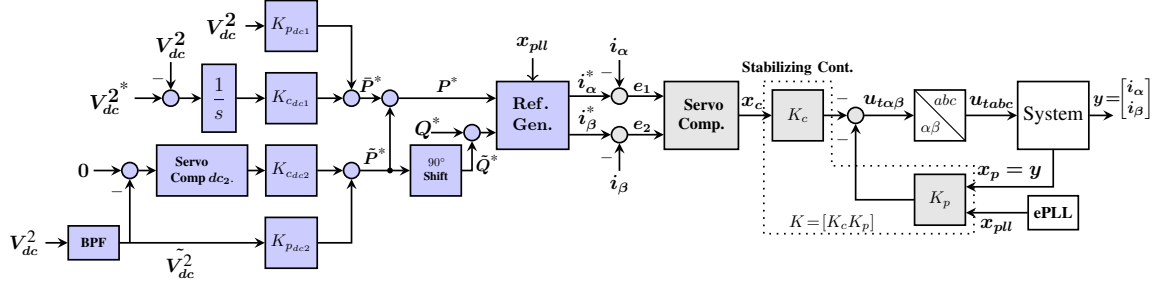


Figure 4.2 Structure of the proposed controller

Fig. 4.2 shows the schematic of the proposed multi-loop control strategy and the three sub-systems it entails; an outer V_{dc} control loop, a reference generator as well as an inner current controller realized in $\alpha\beta$ -frame. The output of dc-voltage controller is applied to the reference generator (Ref. Gen.) block to produce the sinusoidal reference signals of the current control loop. Properties and structures of inner and outer control loops as well as the reference generator are discussed in this section.

4.3.1 Control Objectives

The closed-loop system should be designed so as to meet the following performance specifications.

1. Asymptotic stability of the resultant closed-loop systems.
2. Asymptotic tracking with zero steady-state errors and disturbance rejection for the class of reference signals belonging to constant and double-frequency sinusoidal waveforms for the outer loop and fundamental-frequency sinusoidal signals for the inner current controller.
3. Robust closed-loop stability for the widest possible range of grid impedance parameters and changes in power set-points.
4. Steady-state decoupling between control channels with minimal transient effects.

4.3.2 Current Controller

The structure of the proposed inner current controller is shown in Fig. 4.2. The controller employs the multivariable control approach presented in [6], which entails the ePLL dynamics in its structure but as discussed in 4.2.3, modifies the model to also include the dynamics of the ePLL negative sequence loop. In this regard, the controller incorporates two distinct mechanism; i.e., a servo-compensator and a stabilizing controller. The servo-compensator includes the dynamics of the error signal (i.e., $e_{\alpha,\beta}$) [64] and provides tracking of reference signals and rejection of all disturbance signals belonging to the set $[\sin(\omega t), \cos(\omega t)]$. The stabilizing controller on the other hand, is a full-state feedback gain matrix ensuring robust stability and robust performance of the closed-loop system whose gains are designed based on a new approach to linear quadratic regulator (LQR) theory [80].

As discussed in [6], the reference current signals are generated based on real and reactive power set-points (i.e., P^*, Q^*) and ePLL positive loop state variables (i.e., $x_{(1,2)pll}$).

$$\begin{bmatrix} i_{\alpha}^* \\ i_{\beta}^* \end{bmatrix} = \frac{2}{3} \frac{1}{V_p^2} \begin{bmatrix} x_{1pll} & x_{2pll} \\ x_{2pll} & -x_{1pll} \end{bmatrix} \begin{bmatrix} P^* \\ Q^* \end{bmatrix}. \quad (4.11)$$

The error signals are then produced based on (4.11) and are deployed by the servo-compensator whose minimal realization is

$$\dot{x}_c = A_c x_c + B_c e,$$

where $A_c = \text{blkdiag}\{A_{c_1}, A_{c_2}\}$, $B_c = \text{blkdiag}\{B_{c_1}, B_{c_2}\}$ and

$$A_{c_1} = A_{c_2} = \begin{bmatrix} 0 & \omega \\ -\omega & 0 \end{bmatrix}, \quad B_{c_1} = B_{c_2} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}.$$

This implies

$$\dot{x}_{c1_k} = \omega x_{c2_k}, \quad \dot{x}_{c2_k} = -\omega x_{c1_k} + \underbrace{(i_k^* - i_k)}_{e_k}, \quad (4.12)$$

for $(k = \alpha, \beta)$. Substituting (4.11) into (4.12) and (4.5) results in

$$\begin{aligned}
\dot{x}_{c1\alpha} &= \omega x_{c2\alpha}, \quad \dot{x}_{c1\beta} = \omega x_{c2\beta} \\
\dot{x}_{c2\alpha} &= -\omega x_{c1\alpha} + \frac{2}{3V_p^2} (P^* x_{1\text{pll}} + Q^* x_{2\text{pll}}) - i_\alpha \\
\dot{x}_{c2\beta} &= -\omega x_{c1\beta} + \frac{2}{3V_p^2} (P^* x_{2\text{pll}} - Q^* x_{1\text{pll}}) - i_\beta \\
\frac{di_\alpha}{dt} &= -\frac{R_t}{L_t} i_\alpha + \frac{1}{L_t} u_\alpha - \frac{1}{L_t} (\xi v_{g\alpha} + \sigma u_\alpha + \tau i_\alpha) \\
\frac{di_\beta}{dt} &= -\frac{R_t}{L_t} i_\beta + \frac{1}{L_t} u_\beta - \frac{1}{L_t} (\xi v_{g\beta} + \sigma u_\beta + \tau i_\beta) \\
\dot{x}_{1\text{pll}} &= \mu_1 (\xi v_{g\alpha} + \sigma u_\alpha + \tau i_\alpha) - \mu_1 x_{1\text{pll}} - \omega x_{2\text{pll}} - \mu_1 x_{3\text{pll}} \\
\dot{x}_{2\text{pll}} &= \mu_1 (\xi v_{g\beta} + \sigma u_\beta + \tau i_\beta) - \mu_1 x_{2\text{pll}} + \omega x_{1\text{pll}} - \mu_1 x_{4\text{pll}} \\
\dot{x}_{3\text{pll}} &= \mu_2 (\xi v_{g\alpha} + \sigma u_\alpha + \tau i_\alpha) - \mu_2 x_{3\text{pll}} + \omega x_{4\text{pll}} - \mu_2 x_{1\text{pll}} \\
\dot{x}_{4\text{pll}} &= \mu_2 (\xi v_{g\beta} + \sigma u_\beta + \tau i_\beta) - \mu_2 x_{4\text{pll}} - \omega x_{3\text{pll}} - \mu_2 x_{2\text{pll}}.
\end{aligned} \tag{4.13}$$

Equation set (4.13) describes the dynamics of the augmented system with the state vector $x = [x_c^T, x_p^T]^T$.

Subsequently, the linear state feedback law $u = -Kx$, can be optimally designed to effectively meet the performance specifications of the closed loop system. To this end, the linear transformation $p(D)$ ($= D^2 + \omega^2$, where $D = \frac{d}{dt}$ is the time derivative operator) is applied to (4.13), resulting in

$$\begin{aligned}
\dot{z}_{c1\alpha} &= \omega z_{c2\alpha}, \quad \dot{z}_{c1\beta} = \omega z_{c2\beta} \\
\dot{z}_{c2\alpha} &= -\omega z_{c1\alpha} + \frac{2}{3V_p^2} (P^* z_{1\text{pll}} + Q^* z_{2\text{pll}}) - z_\alpha \\
\dot{z}_{c2\beta} &= -\omega z_{c1\beta} + \frac{2}{3V_p^2} (P^* z_{2\text{pll}} - Q^* z_{1\text{pll}}) - z_\beta \\
\frac{dz_\alpha}{dt} &= -\frac{R_t}{L_t} z_\alpha + \frac{1}{L_t} v_\alpha - \frac{1}{L_t} (\sigma v_\alpha + \tau z_\alpha) \\
\frac{dz_\beta}{dt} &= -\frac{R_t}{L_t} z_\beta + \frac{1}{L_t} v_\beta - \frac{1}{L_t} (\sigma v_\beta + \tau z_\beta) \\
\dot{z}_{1\text{pll}} &= \mu_1 (\sigma v_\alpha + \tau z_\alpha) - \mu_1 z_{1\text{pll}} - \omega z_{2\text{pll}} - \mu_1 z_{3\text{pll}} \\
\dot{z}_{2\text{pll}} &= \mu_1 (\sigma v_\beta + \tau z_\beta) - \mu_1 z_{2\text{pll}} + \omega z_{1\text{pll}} - \mu_1 z_{4\text{pll}} \\
\dot{z}_{3\text{pll}} &= \mu_2 (\sigma v_\alpha + \tau z_\alpha) - \mu_2 z_{3\text{pll}} + \omega z_{4\text{pll}} - \mu_2 z_{1\text{pll}} \\
\dot{z}_{4\text{pll}} &= \mu_2 (\sigma v_\beta + \tau z_\beta) - \mu_2 z_{4\text{pll}} - \omega z_{3\text{pll}} - \mu_2 z_{2\text{pll}}.
\end{aligned} \tag{4.14}$$

Based on the transformed system model of (4.14), the quadratic cost function

$$\begin{aligned}
J &= \int_0^\infty [q_1(e_\alpha^2 + e_\beta^2) + \frac{q_2}{\omega^2}(\dot{e}_\alpha^2 + \dot{e}_\beta^2) + q_5(z_\alpha^2 + z_\beta^2) \\
&\quad + q_7 z_{1\text{pll}}^2 + q_8 z_{2\text{pll}}^2 + q_9 z_{3\text{pll}}^2 + q_{10} z_{4\text{pll}}^2 + v_\alpha^2 + v_\beta^2] dt.
\end{aligned} \tag{4.15}$$

is formed where its minimization leads to an optimal feedback gain K that corresponds

to the weighting factors q_i . The closed loop poles are then shaped based on the design algorithm provided in [79]. By using the system parameters given in Table 4.1 and changing the weighting gains q_1 through q_{10} , an optimum design is achieved where

$$K = \left[\underbrace{\begin{bmatrix} -7.9e5 & -1.0e4 & -5.0e4 & -381 \\ 1.15e4 & 412.5 & -8.8e5 & -1e4 \end{bmatrix}}_{K_c} \mid \underbrace{\begin{bmatrix} 23.8 & 0.59 & 1.68 & -0.19 & 0.35 & 0.12 \\ -0.99 & 22.9 & 0.17 & 0.35 & 0.52 & 1.35 \end{bmatrix}}_{K_p} \right].$$

Fig. 4.3 shows the design procedure where $q_{1,3}$ are changed from 10^8 to $10^{14.5}$, $q_{2,4}$ are increased from 10 to $10^{8.7}$, $q_{5,6}$ are varied from 10^{-1} to $10^{2.6}$, $q_{7,8}$ are shifted from 10^{-2} to $10^{0.5}$ and finally $q_{9,10}$ are changed from 10^{-1} to $10^{0.65}$. This places the closed-loop poles at $-280 \pm j200$, $-360 \pm j300$, $-610 \pm j560$, $-670 \pm j680$, and -1000 on the complex plane which yields fast and well-damped transient response.

4.3.3 DC-Link Controller

Third-order Harmonic Currents Suppression

Under voltage imbalance, (4.11) produces reference currents which contain third-order harmonics. This can be shown by assuming that real and reactive power reference signals have

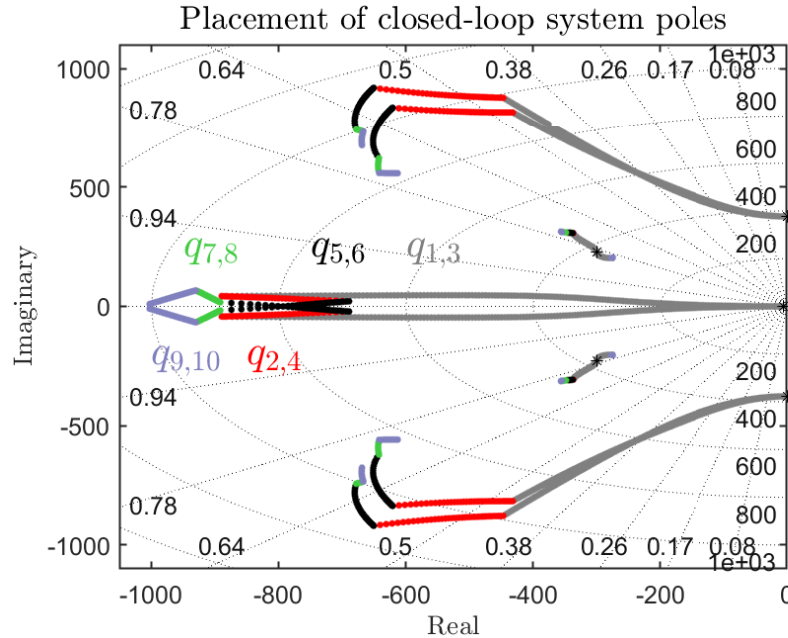


Figure 4.3 Controller design procedure using LQR method

Table 4.1 Parameters of the Study System

parameters	symbol	value
dc bus voltage	V_{dc}	600 V
Grid phase voltage	v_s	120 V (rms)
System frequency	f	60 Hz
PWM switching frequency	f_{sw}	5 kHz
RL filter resistance	R_t	0.1 Ω
RL filter inductance	L_t	5 mH
Grid internal resistance	R_g	1 m Ω
Grid internal inductance	L_g	5, 10, 17, 28 mH
Short Circuit Ratio ($S \approx 5$ kVA)	SCR	6.5, 2.5, 1.9, 1.0

the general form of

$$P^* = A + B \cos(2\omega t + \delta), \quad Q^* = C + D \cos(2\omega t + \zeta), \quad (4.16)$$

where A, B, C and D are constants values corresponding to the constant and oscillatory parts of the control effort while δ and ζ are the oscillation phase angles. Using (4.11), i_α can be calculated

$$\begin{aligned} i_\alpha = & \frac{2}{3} \frac{1}{V_p^2} [A (v_p \cos(\omega t + \delta_p)) + C (v_p \sin(\omega t + \delta_p)) \\ & + B v_p \left\{ \frac{\cos(\omega t + \delta - \delta_p) + \cos(3\omega t + \delta_p + \delta)}{2} \right\} \\ & + D v_p \left\{ \frac{\sin(3\omega t + \delta_p + \zeta) - \sin(\omega t + \zeta - \delta_p)}{2} \right\}]. \end{aligned} \quad (4.17)$$

According to (4.17) (where i_β can be used alternatively), in order to eliminate third-order harmonics, the following conditions should hold

$$B = D, \quad \delta = \zeta + 90^\circ. \quad (4.18)$$

As shown in Fig. 4.2, the above conditions are met by applying a 90° delay to the oscillatory component of the real power and adding it to the constant component of reactive power reference signal.

Control Structure

As discussed in Section 4.2.3, real power at VSC terminal contains nonlinear terms consists of positive and negative components of current. More specifically, calculating real power at VSC terminals using (4.17) by considering the constraints in (4.18) and assuming infinit

bandpass of current control loop, yields

$$\begin{aligned}
\bar{P}_t &= A + B \frac{V_n}{V_p} \cos(\delta - \delta_{vp} - \delta_{vn}) \\
\tilde{P}_t &= A \frac{V_n}{V_p} \cos(2\omega t + \delta_{vn} + \delta_{vp}) + C \frac{V_n}{V_p} \sin(2\omega t + \delta_{vn} + \delta_{vp}) \\
&+ B \cos(2\omega t + \delta) + \frac{4\omega L_t}{3 V_p^2} B [-A \sin(2\omega t + \delta) + C \cos(2\omega t + \delta)]
\end{aligned} \tag{4.19}$$

Equation (4.19), reveals nonlinear couplings between constant and oscillatory components of dc-voltage control loop. However, given that the system has a stable operating point in which the oscillatory part of (4.19) becomes zero, steady-state decoupling could be achieved by ensuring proper disturbance rejection features of the loop. In other words, if properly designed, the resonant-based mechanism of the control loop can treat the oscillatory coupling effects as disturbances. One way of doing this, is to widen the band gap between the resonant and integral characteristics of the closed-loop system. To this end, equation (4.6), can be stated as the sum of the differential equations

$$\begin{aligned}
\dot{x}_{dc1} &= A_{dc1} x_{dc1} + B_{dc1} u_{dc1} + F_{dc1} W_{dc1}, \quad y_1 = x_{dc1} \\
\dot{x}_{dc2} &= A_{dc2} x_{dc2} + B_{dc2} u_{dc2} + F_{dc2} W_{dc2}, \quad y_2 = x_{dc2}.
\end{aligned} \tag{4.20}$$

where, $x_{dc1} = \bar{V}_{dc}^2$, $x_{dc2} = \tilde{V}_{dc}^2$, $u_{dc1} = \bar{P}^*$, $u_{dc2} = \tilde{P}^*$, $A_{dc1,2} = -\frac{2}{RC_{dc}}$, $B_{dc1,2} = -\frac{2}{C_{dc}}$, $F_{dc1,2} = -\frac{2}{C_{dc}}$, and $W_{dc1,2}$ contain the coupling terms in (4.19). The dynamics of x_{dc1} and x_{dc2} are characterized based on their constant and oscillatory natures.

Based on (4.20), the dc voltage control loop comprises of two parallel sub-controllers; i.e., an integrator-based control system, as well as a resonant-based compensator. While the constant part of the voltage is adjusted by the dc control loop, the resonant-based compensator is responsible for eliminating the double-frequency terms and providing a ripple-free dc voltage. The structure of the proposed controller is shown in Fig. 4.2 where the two parallel compensators are developed based on the subsets of (4.20) and are individually designed by deploying the robust control technique in [79].

Controller Design

Given that reference and disturbances are step-wise functions in the first subset of (4.20), the transfer function of the dc-loop servo-compensator essentially becomes an integrator with the minimal realization of

$$\dot{x}_{cdc1} = A_{cdc1} x_{cdc1} + B_{cdc1} e_{dc1} \tag{4.21}$$

where $A_{cdc1}=0$, $B_{cdc1}=I$ and $e_{dc1} = \bar{V}_{dc}^2 - \bar{V}_{dc}^{2*}$ are scalar values.

As for the resonant-based compensator, with frequency tuned at 2ω , the structure of the servocompensator becomes

$$\dot{x}_{cdc2} = A_{cdc2}x_{cdc2} + B_{cdc2}e_{dc2} \quad (4.22)$$

where

$$A_{cdc2} = \begin{bmatrix} 0 & 2\omega \\ -2\omega & 0 \end{bmatrix}, \quad B_{cdc2} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad e_{dc2} = 0 - \widetilde{V_{dc}^2}.$$

Subsequently, the augmented systems including the servo compensators as well as the dc sub-systems in (4.20), are represented by

$$\begin{aligned} \dot{x}_k &= A_{augk}x_k + B_{augk}u_{dc_k} + B_{refk}y_{refk} + F_{augk}W_{dc_k} \\ y_k &= C_{augk}x_k, \quad A_{augk} = \begin{bmatrix} A_{dc_k} & -B_{dc_k} \\ 0 & A_{dc_k} \end{bmatrix}, \quad B_{augk} = \begin{bmatrix} 0_{(k \times 1)} \\ B_{dc_k} \end{bmatrix} \\ B_{refk} &= \begin{bmatrix} B_{dc_k} \\ 0 \end{bmatrix}, \quad F_{augk} = \begin{bmatrix} 0_{(k \times 1)} \\ F_{dc_k} \end{bmatrix}, \quad C_{augk} = [0_{(1 \times k)} \quad 1], \end{aligned} \quad (4.23)$$

where $k = 1, 2$ corresponds to each sub-system.

The gain matrices K_{dc_k} in the state vector $u_{dc_k} = -K_{dc_k}x_k$ are then designed by deploying the same optimal approach in Section 4.3.2. The derivative transformations $p_1(D) = D$ and $p_2(D) = D^2 + 4\omega$ are respectively applied on the integrator and resonant-based sub-systems of (4.23) resulting in

$$\dot{z}_{dc_k}(t) = A_k z_{dc_k} + B_k v_{dc_k}, \quad v_{dc_k} = -K_{dc_k} z_{dc_k}$$

where every $z = p(D)x$ and $v = p(D)u$.

Finally, the cost functions J_{dc_k} are defined as

$$\begin{aligned} J_{dc1} &= \int_0^\infty [q'_1 e_{dc1}^2 + q'_2 z_{dc1}^2 + r' v_{dc1}^2] dt, \\ J_{dc2} &= \int_0^\infty [q''_1 e_{dc2}^2 + \frac{q''_2}{4\omega^2} \dot{e}_{dc2}^2 + q''_3 z_{dc2}^2 + r'' v_{dc2}^2] dt, \end{aligned}$$

and are minimized through changing weighting factors q'_i and q''_i . Fig. 4.4 visualizes the design procedure of both dc sub-control loops where the gain matrices are obtained as

$$K_{dc1(1 \times 2)} = \left[\underbrace{2.37}_{K_{cdc1}} \mid \underbrace{-6.8e-2}_{K_{pdc1}} \right], \quad K_{dc2(1 \times 3)} = \left[\underbrace{-2.4e4 \quad 4.6}_{K_{cdc2}} \mid \underbrace{-7.4e-2}_{K_{pdc2}} \right].$$

From Fig. 4.4, it is evident that the feedback gain matrices are designed so that the resonant-

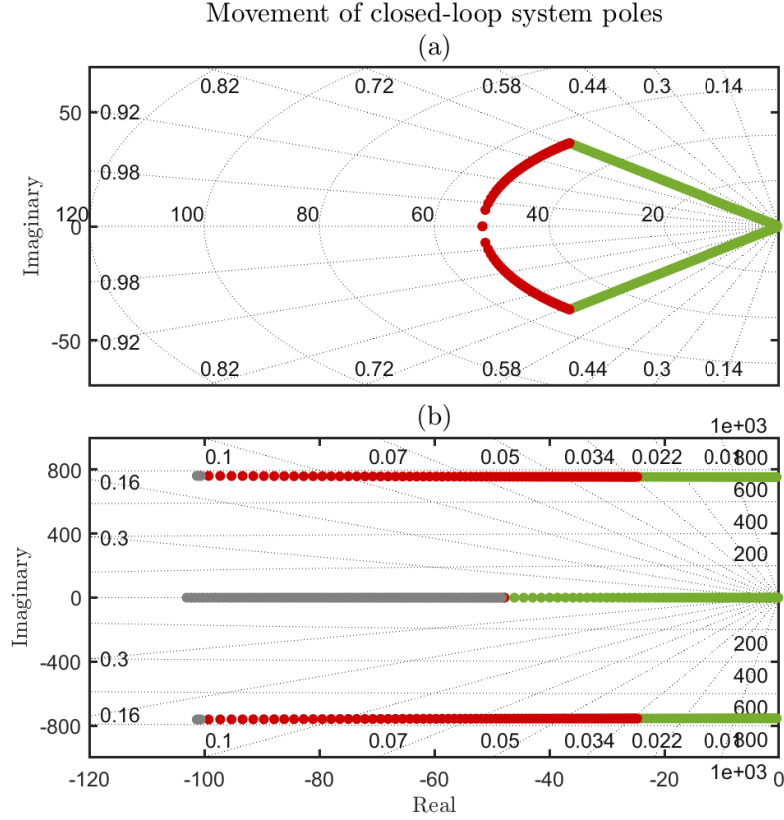


Figure 4.4 Outer controller design procedure using LQR method

based compensator exhibits a much faster dynamic response and the bandwidth of the two adjacent control loops are separated by a relatively large factor.

4.3.4 Controller Decoupling Features

Resonant-Based Compensator

The constant part of reactive power in (4.19) is denoted by C which is a constant value with no dynamics. Furthermore, regarding faster transient response of the resonant-based control loop, the output of integrator-based compensator, u_{dc1} ($=A$), can also be treated as a constant term in W_{dc2} . Thus, defining $M\cos(2\omega t + \varphi) = A\frac{V_n}{V_p}\cos(2\omega t + \delta_{vn} + \delta_{vp}) + C\frac{V_n}{V_p}\sin(2\omega t + \delta_{vn} + \delta_{vp})$, $E = \frac{4\omega L_t}{3 V_p^2}$, and $h = -2\omega B\sin(2\omega t + \delta)$, oscillatory dynamics in (4.23) can be stated as

$$\dot{x}_2 = A_{aug2}x_2 + B'_{aug2}u_{dc2} + B''_{aug2}h + B_{ref2}y_{ref2} + F_{aug2}W'_{dc2}, \quad (4.24)$$

where

$$B'_{aug2} = [0, 0, -\frac{2(1+CE)}{C_{dc}}]^T, B''_{aug2} = [0, 0, -\frac{EA}{\omega C_{dc}}]^T, W'_{dc2} = M \cos(2\omega t + \varphi).$$

Given that $h = d(u_{dc2})/dt$ and $u_{dc2} = -K_{dc2}x_2$, h can be written based on the augmented system state variables such that

$$h = -K_{dc2}(1)2\omega x_{c2dc2} - K_{dc2}(2)(-2\omega x_{c1dc2} - x_{dc2}) - K_{dc2}(3)\dot{x}_{dc2}.$$

Thus, the system model considering the coupling effects is achieved where the dc components of real and reactive power set-points (i.e., A and C), exist as structured uncertainties. The root locus analysis of the closed-loop system in (4.24) is provided in Fig. 4.5, where $C \in [-5e3, 5e3]$ and $A \in [-15e3, 15e3]$. The lighter shades of red correspond to the values of C closer to the upper limit of its closed interval while darker shades indicate its values closer to its lower boundary. From Fig. 4.5, it is evident that the closed-loop system is robustly stable for a wide range of changes in operating set-points however negative values of C (i.e., absorption of reactive power) degrade the stability by moving the closed loop poles towards the origin.

Integrator-Based Compensator

Assuming that the resonant-based control loop achieves reference tracking with zero steady-state error, (4.19) yields

$$\begin{aligned} EBA \cos\left(2\omega t + \delta - \frac{\pi}{2}\right) + EBC \cos(2\omega t + \delta + \pi) = \\ aA \cos(2\omega t + \Delta) + aC \cos\left(2\omega t + \Delta - \frac{\pi}{2}\right) + B \cos(2\omega t + \delta), \end{aligned} \quad (4.25)$$

where $\Delta = \delta_{vp} + \delta_{vn}$ and $a = \frac{V_n}{V_p}$. The phasor representation of (4.25) considering the angular frequency 2ω and initial phase angle $-\Delta$ is depicted in Fig. 4.6 where the following set of equations can be derived

$$\begin{aligned} EBA \cos\left(\delta - \Delta - \frac{\pi}{2}\right) - EBC \cos(\delta - \Delta) &= aA + B \cos(\delta - \Delta), \\ EBA \sin\left(\delta - \Delta - \frac{\pi}{2}\right) - EBC \sin(\delta - \Delta) &= B \sin(\delta - \Delta) - aC. \end{aligned} \quad (4.26)$$

Solving (4.26) for $B \cos(\delta - \Delta)$, results in

$$B \cos(\delta - \Delta) = -aA / ([1 + EC]^2 + E^2 A^2).$$

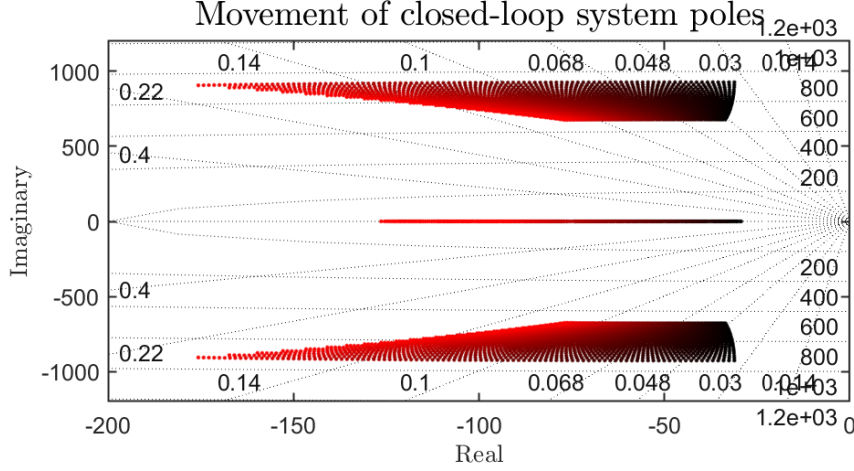


Figure 4.5 Evolution of closed-loop poles of the resonant-based compensator due to coupling effects

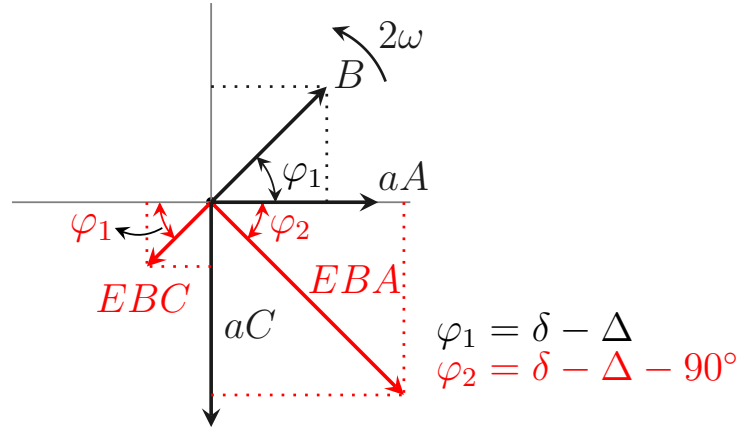


Figure 4.6 Phasor representation of oscillatory terms of the real power

Thus, the coupling dynamics in (4.20) nonlinearly affect the dc sub-system input such that

$$\begin{aligned} \dot{x}_{dc1} &= A_{dc1}x_{dc1} + B_{dc1}u'_{dc1}, \quad y_1 = x_{dc1} \\ u'_{dc1} &= u_{dc1} - \frac{au_{dc1}}{[1 + EC]^2 + E^2u_{dc1}^2} \end{aligned} \quad (4.27)$$

Fig. 4.7 depicts the dc sub-system input u'_{dc1} when $u_{dc1} \in [-5e3, 5e3]$, $C \in [-2e3, 5e3]$, $a = 1$ (i.e., VUF=100%) and for system parameters of Table 4.1.

It is evident from Fig. 4.7, that for $C \in [0, 5e3]$ (depicted in blue) the effect of nonlinear term is negligible and the system input can be approximated by the linear function $u'_{dc1} = u_{dc1}$ (solid black line). For $C < 0$ (shown in red color) however, the approximation error becomes significant around origin which can destabilize the system.

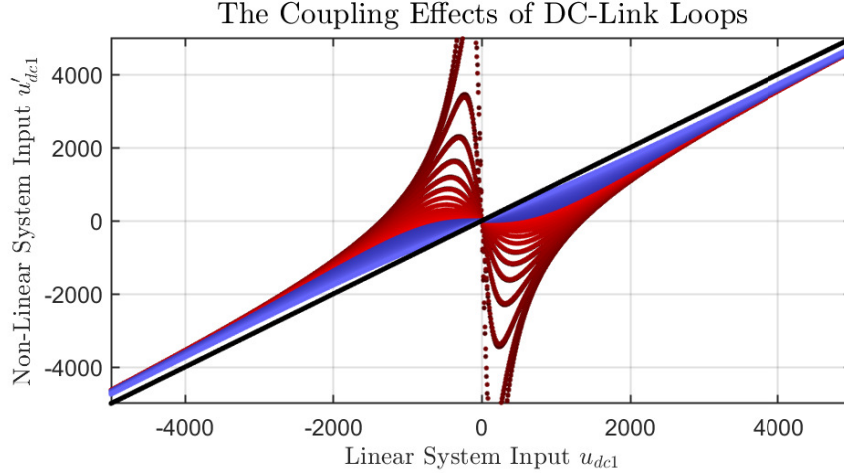


Figure 4.7 The coupling effects of resonant-based compensator on the integrator-based control loop

4.4 Simulation Results

4.4.1 Unbalanced Grid

In order to validate the performance of the proposed controller under voltage imbalance, the following scenario is considered and carried out using MATLAB software and the system parameters of Table 4.1: the system is operating in steady-state conditions when a single-phase-to-ground fault occurs at $t = 0.3\text{s}$ causing voltage imbalance at PCC ($VUF = 50\%$). The VSC is injecting 3 kW of real power with unity power factor, dc-voltage is set on 600 V and grid is stiff, i.e., $L_g=1\text{ mH}$. The dc-link voltage is increased by 100 V and is returned to 600 V at $t=0.4\text{s}$ and $t=0.6\text{s}$, respectively. Then the set-point of reactive power is changed from 0 kVar to 3 kVar at $t=0.8\text{s}$ followed by a step-change in direction of real power from 3 kW to -2 kW at $t=0.9\text{s}$. The response of the controller is shown in Fig. 4.8, where the proposed controller successfully damps the dc-link oscillations while providing a smooth response to the 100 V step-changes. The proposed controller also successfully regulates both real and reactive powers even when the VSC is confronted with a sudden change in real power that alters the direction of power.

To better show the effect of the resonant compensator and its integration in the proposed method, under same unbalanced conditions, the shifting operator is disabled. As shown in Fig. 4.9, by removing the 90° shift, third-order harmonics rise at the output currents of the VSC. Next, the resonant-based compensator is completely withdrawn from the outer-loop. As evident in Fig. 4.9, the double-frequency oscillations re-emerge on the dc-link voltage.

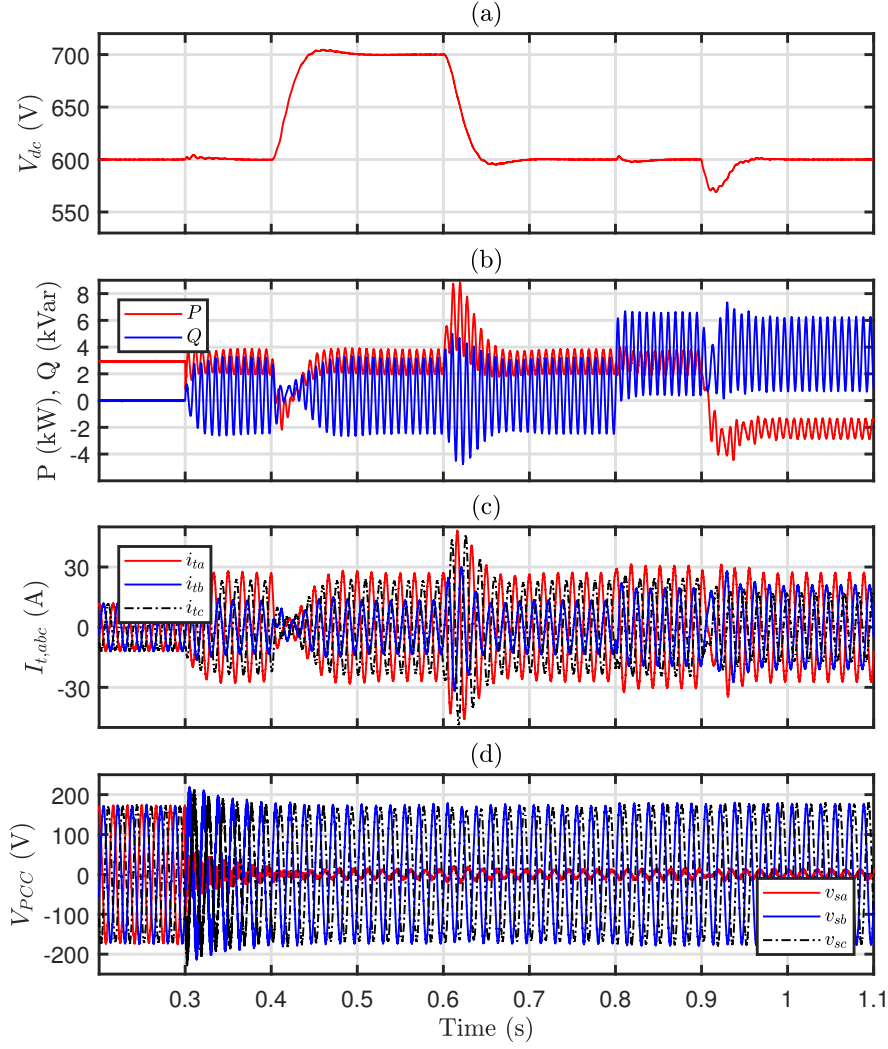


Figure 4.8 Response of the proposed controller to a single-phase-to-ground fault: (a) DC voltage, (b) real and reactive powers, (c) three-phase currents and (d) PCC voltages

4.4.2 Weak Grid

In this case study the performance of the proposed controller is studied with different levels of grid stiffness. Based on Table 4.1, grid inductance L_g ranges from 5 to 28 mH. As formulated in [6], under the operating points $P=2$ kW and $Q=1.97$ kVar, these grid inductances correspond to the short circuit ratios (SCRs): 6.5, 2.5, 1.8 and 1.0. While SCRs higher than 3.0 indicate stiff grids, the unity SCR is identified as extremely weak grid condition which puts the system on the verge of voltage instability. While the VSC is injecting 1.8 kW into the grid, at $t=0.5$ s, the set-point of reactive power is changed from 0 to 1.97 kVar. Further, at $t=0.6$ s, the injected real power increases to 2 kW through a step-change. The response of the proposed controller for different levels of SCR is shown in Fig. 4.10 where it can be seen

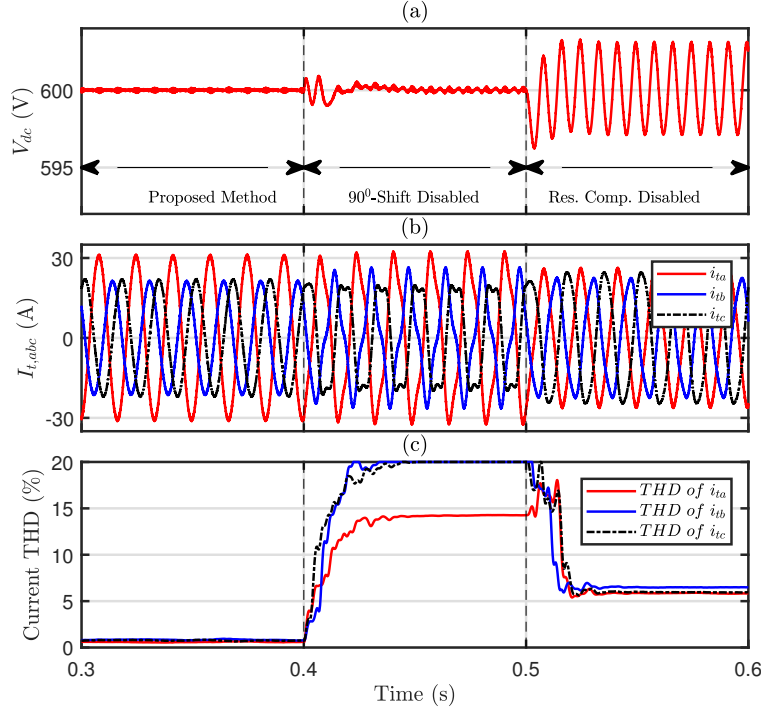


Figure 4.9 Performance of the proposed controller with and without 90° shift operator and resonant compensator under voltage imbalance: (a) DC voltage, (b) three-phase currents and (c) VSC currents THD

that the controller has a very fast response in adjusting the reactive power while exhibiting smooth dynamics for changes in real power. There are some overshoots and coupling as the system becomes weaker, however the system remains stable even under extremely weak grid conditions (SCR = 1).

To highlight the enhanced performance of the proposed controller under weak grid conditions, the same scenario is repeated with a conventional controller. This conventional controller benefits from the same optimally designed multivariable structure as the proposed controller but the dynamics of the ePLL are not considered in its inner current control loop. The performance of the conventional controller is depicted in Fig. 4.11 where it exhibits similar dynamics as the proposed controller for higher SCRs. The controller performance, however, degrades drastically as the grid becomes weaker. The overshoots and oscillations worsen and system eventually becomes unstable at SCR = 1.9.

4.5 Experimental Results

To verify the enhanced performance of the proposed controller, hardware-in-the-loop tests are carried out. The HIL test setup is shown in Fig. 4.12 where the controller is implemented using OPAL-RT OP5700 FPGA-based real-time simulator. The voltages and currents are

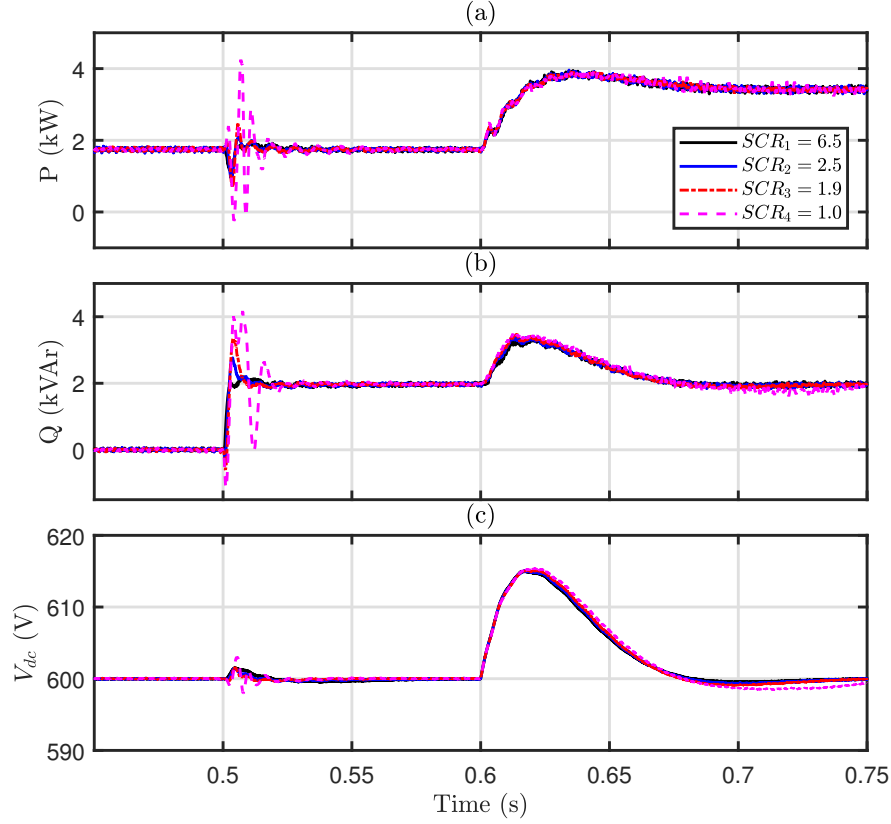


Figure 4.10 Response of the proposed controller under extremely weak grid conditions: (a) DC voltage, (b) real and reactive powers, (c) three-phase currents and (d) PCC voltages

measured using the data acquisition unit, OP8660 and are transferred to the OP5700 to be deployed by the controller. The regenerative grid simulator from Chroma (model 61830) is used as the ideal grid and the programmable dc power supply 62000H from Chroma is employed for generating the dc power. The data is stored on the host PC and then is plotted using MATLAB software. The system parameters used in the experimental tests are shown in Table 4.2.

4.5.1 Severe Unbalanced Fault

In this case study, the grid is stiff, the VSC is injecting 250 W of real power into the grid and reactive power is set on 200 VAr when a single-phase-to-ground fault occurs at the grid side resulting voltage imbalance at PCC with VUF=45%. After 5 cycles, a second unbalanced fault occurs causing 85% of VUF. The transient response of the proposed controller is shown in Fig. 4.13 where it can be seen that the controller successfully eliminates dc voltage oscillations while providing purely sinusoidal currents. Further, to show the effectiveness of the proposed resonant-based dc voltage control loop, this loop is deactivated when $L_g=5$ mH and under

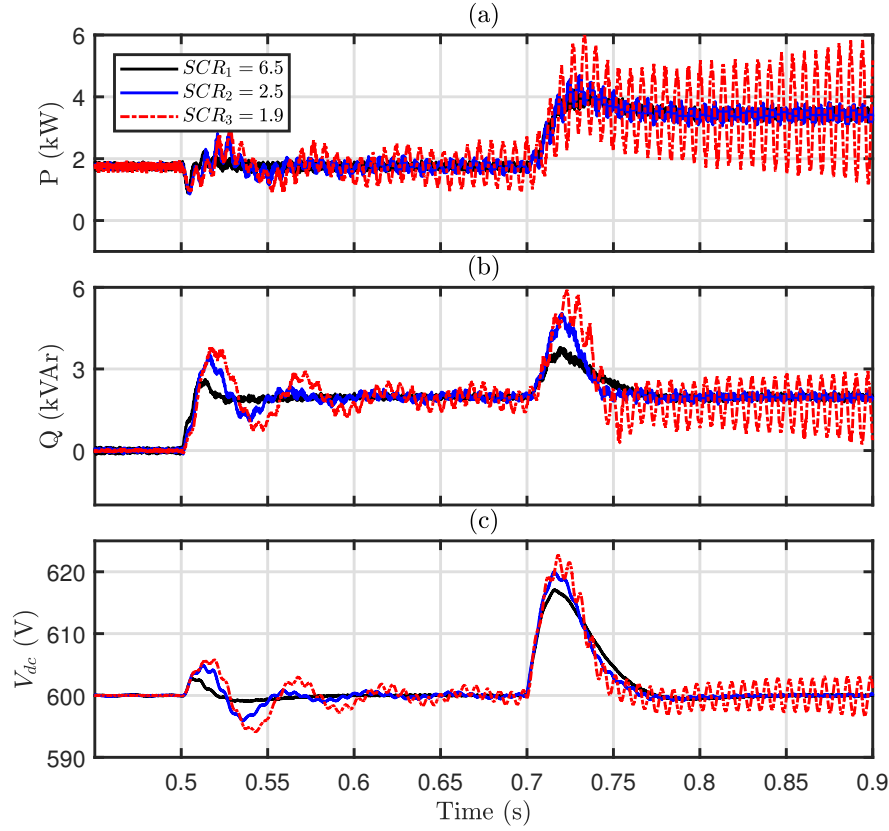


Figure 4.11 Responses of the conventional controller under extremely weak grid conditions: (a) DC voltage, (b) real and reactive powers, (c) three-phase currents and (d) PCC voltages

single-phase unbalanced fault, where the transient response is shown in Fig. 4.14. It could readily be observed that as soon as the resonant-based loop is disabled, double-frequency oscillations emerge on the dc voltage.

4.5.2 Weak and Unbalanced Grid Conditions

For this test, the grid is weakened with L_g of 10 mH and the system is working under 15% of voltage imbalance. The constant part of the real and reactive powers injected by the VSC is 60 W and 100 VAr, respectively. Then the real power increases to 250 W due to a step-change in dc power-source output current. Given that the voltage amplitude in the faulty phase of PCC is less than the other two phases, it works closer to its power capacity limit. This situation corresponds to weak grid conditions with equivalent short circuit ratio of 1.5 ($SCR=1.5$). The response of the proposed controller is shown in Fig. 4.15 where the controller exhibits smooth transients and successfully regulates the powers. For sake of comparison and to highlight the improved performance of the system under weak grid conditions, the test is repeated for the conventional controller under identical conditions. The dynamic response



Figure 4.12 A photograph of the experimental setup

of the controller is depicted in Fig. 4.16 where it is seen that after the step-up in the dc power-source output, the system becomes unstable.

Furthermore, the response of the proposed controller to a step-change in dc-link voltage under weak and unbalanced conditions, is shown in Fig. 4.17, where it can be seen that the controller smoothly tracks the reference voltage set-point from 150 to 160 V.

4.5.3 Extremely weak Grid Conditions

To show the capability of the proposed controller under extremely weak grid conditions, the grid is weakened by incorporating $L_g=20$ mH (SCR=1, $S\approx 350$ VA). Reactive power is set at 165 VAR and real power increases from 250 W to 300 W by a step-up in the generated dc power. The response of the proposed controller is depicted in Fig. 4.18 where the controller succeeds to adjust the powers.

4.6 Conclusion

In this chapter, a new robust controller is proposed for enhanced performance of grid following VSCs under extremely weak and unbalanced grid conditions. The controller has a nested multi-loop structure consisting of a dc-link voltage regulator cascaded with an inner current

Table 4.2 Parameters of the Experimental Setup

Parameter	Symbol	Value
DC-link voltage	V_{dc}	150 V
DC-Link Capacitor	C_{dc}	1500 μ f
Grid phase voltage	V_s	30 V (rms)
System frequency	f	60 Hz
PWM switching frequency	f_{sw}	5 kHz
RL filter inductance	L_t	5 mH
Grid internal inductance	L_g	10, 20 mH

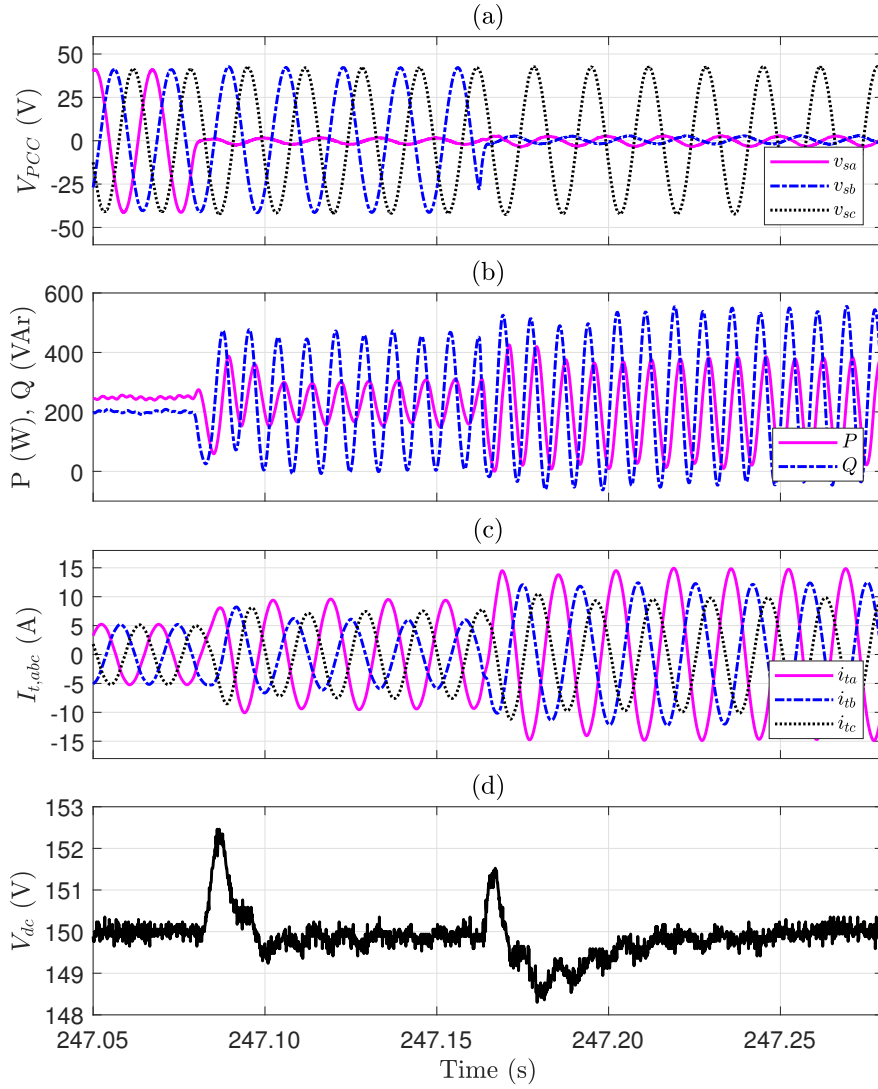


Figure 4.13 Responses of the proposed controller to a severe unbalanced fault with 85% of VUF: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

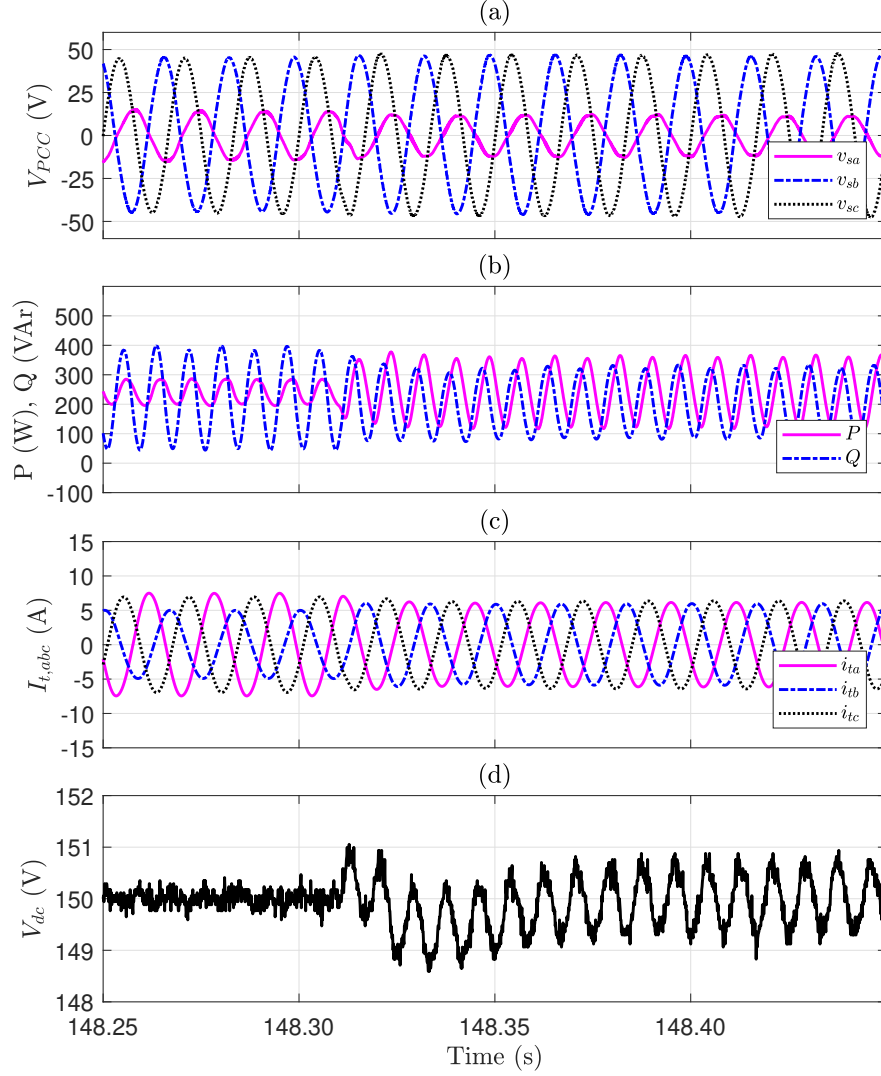


Figure 4.14 Transients of the proposed controller when the dc-link resonant-based compensator is disabled: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

controller. The current controller is realized in stationary frame and benefits from a MIMO structure which incorporates the dynamics of a newly introduced ePLL into its state-feedback mechanism and improves system performance and stability under weak grid conditions. The outer control loop on the other hand, comprises of two parallel sub-systems; i.e., an integrator-based as well as a resonant-based compensator. The dynamics of the dc-link and the impacts of the control loops are thoroughly studied and both dc-link sub-controllers are rigorously designed so that steady-state decoupling between them is achieved. Finally, the improved performance of the controller is verified through software simulations as well as experimental tests.

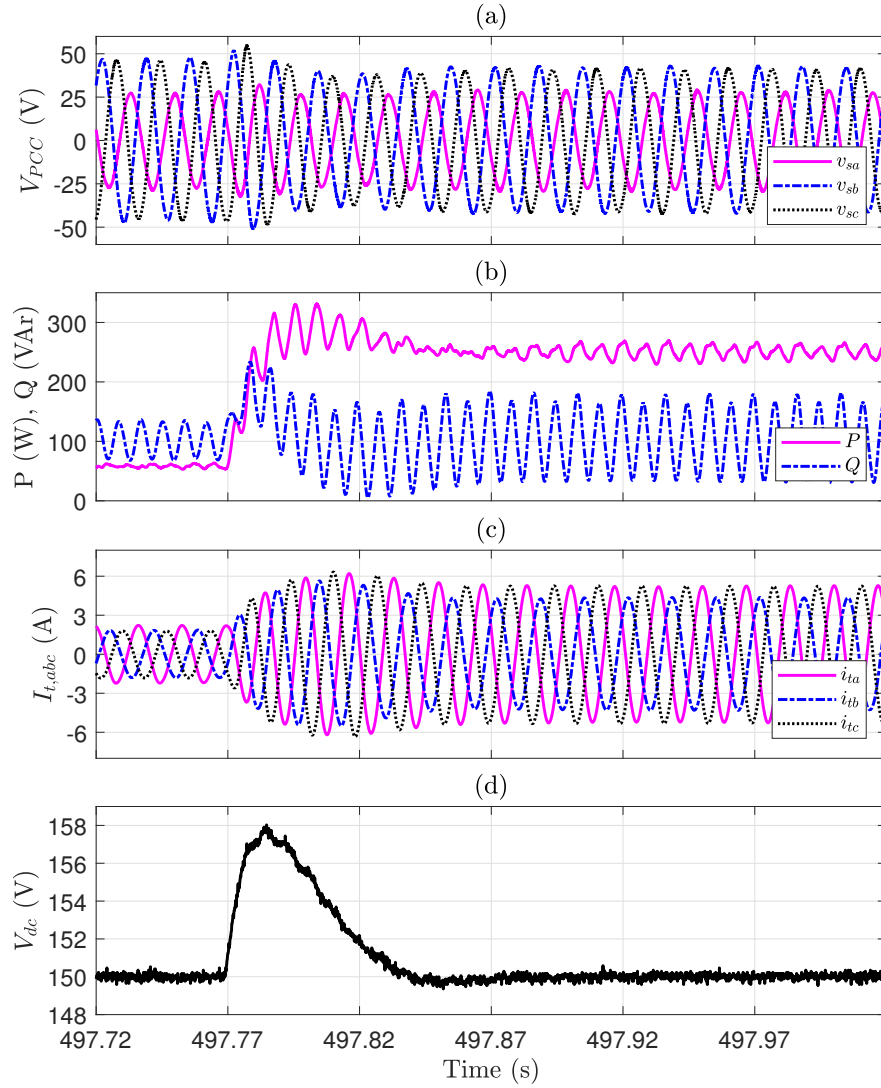


Figure 4.15 Responses of the proposed controller under weak and unbalanced grid conditions: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

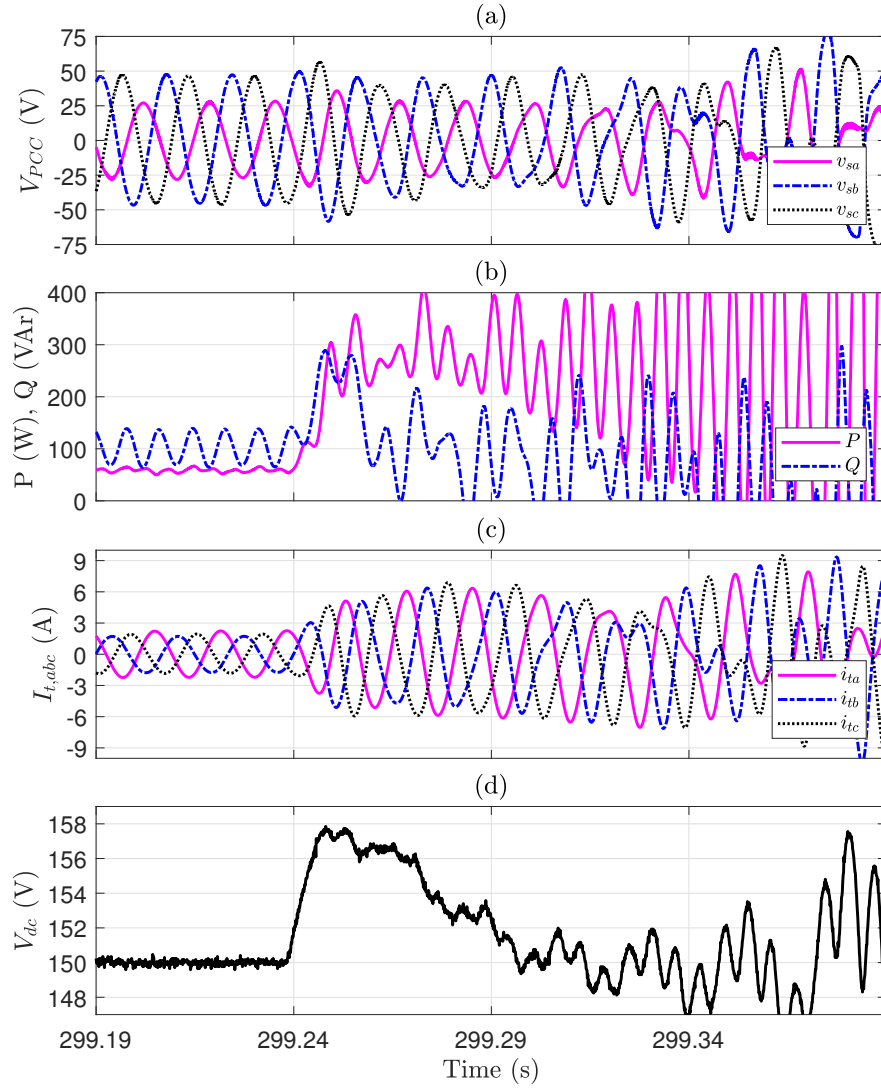


Figure 4.16 Responses of the conventional controller under weak and unbalanced grid conditions: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

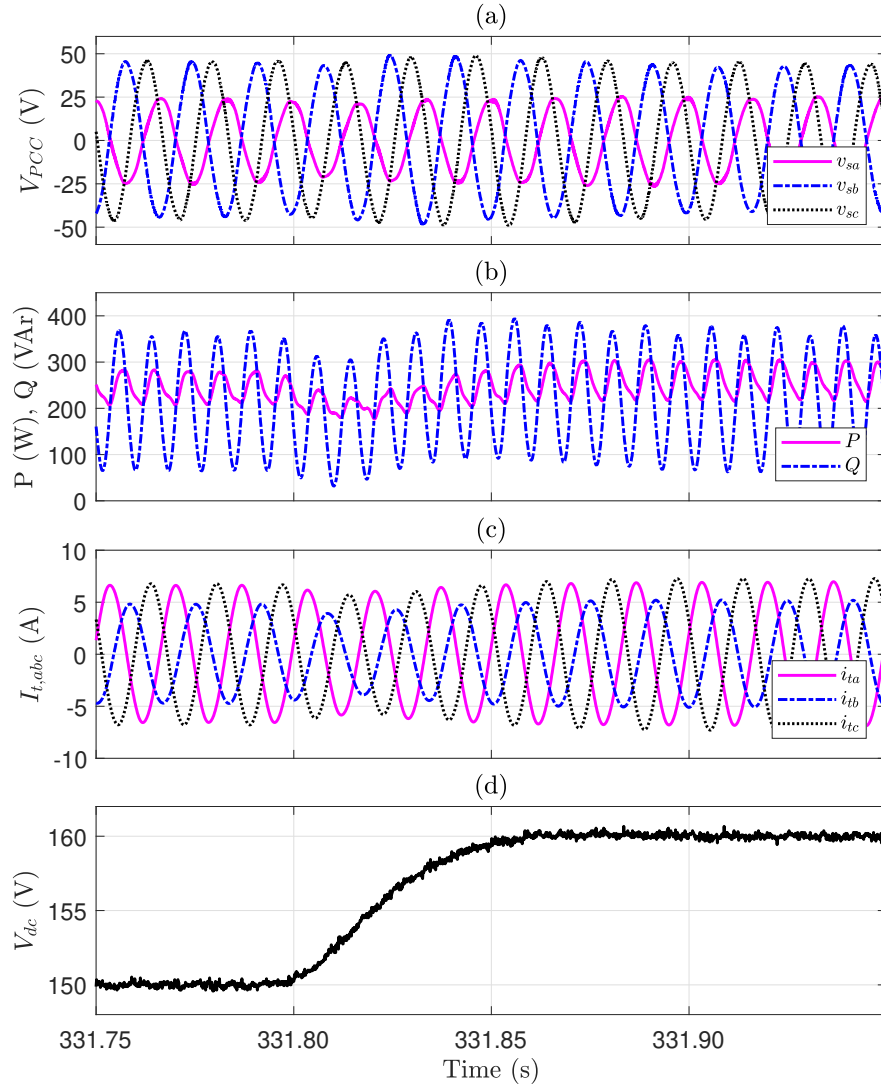


Figure 4.17 Responses of the proposed controller to dc voltage step-change under weak and unbalanced grid conditions: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

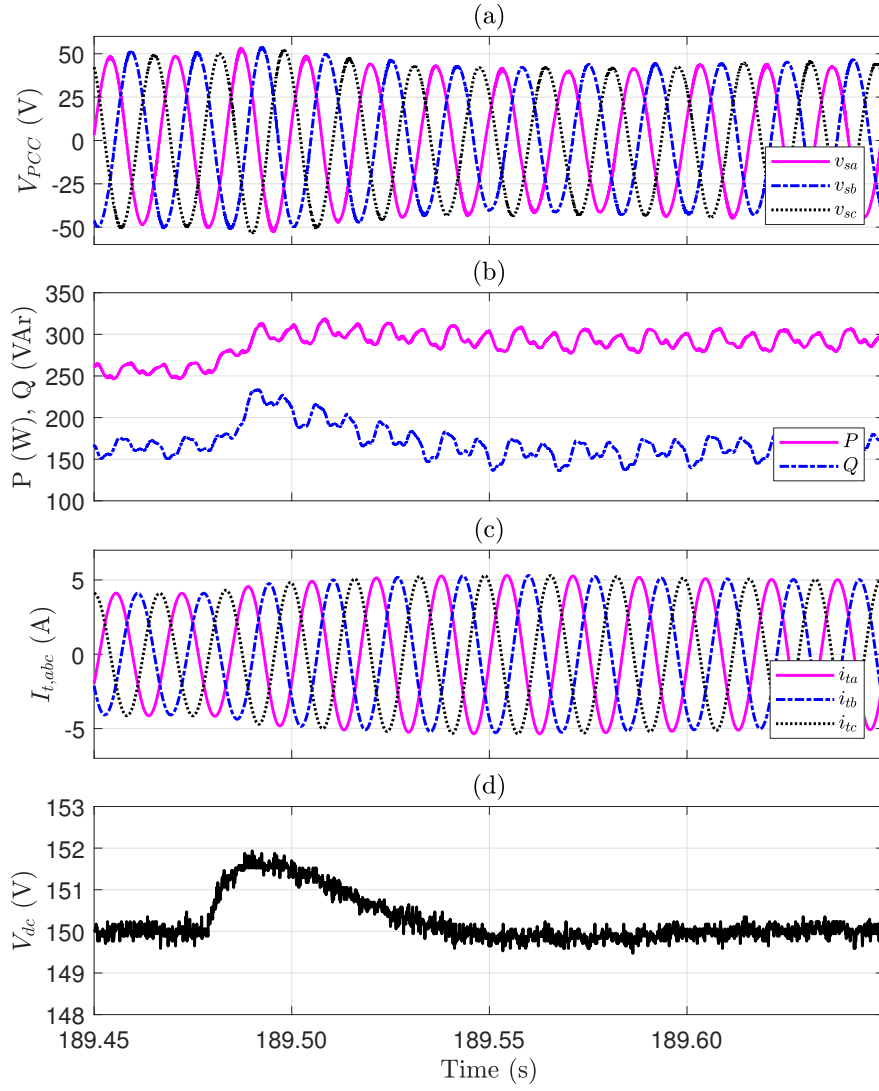


Figure 4.18 Responses of the proposed controller under extremely weak grid conditions: (a) PCC voltages, (b) real and reactive powers, (c) three-phase currents and (d) DC voltage

CHAPTER 5 ROBUST CONTROL STRATEGY IN STATIONARY REFERENCE FRAME FOR GRID-FORMING VOLTAGE SOURCE CONVERTERS

5.1 Background

The increased integration of distributed energy resources (DERs) has changed the structure of the modern power grid which was previously dominated by centralized synchronous generators (SGs) [83–86]. Such advancement is acquired due to the inherent flexibility of voltage source converters (VSCs) which are increasingly employed as an interface between renewable energy sources and the power grid [87–89]. As discussed in previous chapters, grid-following (GFL) control is the most common approach used for VSC-based DERs, where the VSC is controlled as a current source to regulate the real and reactive power exchanged with the grid [90]. In GFL mode, the voltage and frequency are dictated by the host grid [91]. The VSCs can also be controlled as a voltage source to regulate the voltage at the point of common coupling (PCC). Such mode of operation is called grid-forming (GFM) control mode, where the VSC improves the grid voltage profile and supplies the demanded power to the grid. The GFM converter can also operate as an oscillation damping source and enhance the stability margin of the system [92].

The performance of both the GFL and GFM VSCs, however, degrades under weak grid conditions since the phase-locked loop (PLL) dynamics adversely affect the robustness of the overall closed-loop system [69, 93–95]. In [95] it is shown that PLL interactions with dc-link voltage control loop can be modeled as a partial positive feedback which attributes to system instability under weak grid conditions. Based on the analysis of a classic control structure under weak grid, an improved outer loop control scheme is provided in [25] which is based on decoupling and gain-scheduling control techniques and considers the system nonlinearities. The controller in [25], however, becomes overly complicated to design. PLL-less controllers have recently been introduced (e.g., [29–31]) with the claim to enhance controller performance under weak grid conditions. The study in Chapter 2 however, revealed that such methods are prone to the same stability degradation issues as PLL-based methods and thus, can not provide substantially improved performance.

Generally, when designing outer-loop controllers, the present literature ignores the PLL's influence while the dominant poles affecting the converter stability are significantly correlated with the states of the PLL and the ac voltage at the PCC [69, 95, 96].

To address the issues with the existing ac-voltage controllers under weak and very weak-grid conditions, an alternative strategy is suggested in this chapter where a new robust control strategy is proposed for GFM VSCs. The objectives of the controller are defined as; 1) tracking the set-point of real power and, 2) compensating the impact of the weak grid on PCC voltage by adjusting its voltage magnitude. To reach these goals, the multivariable PLL-integrated controller developed in Chapter 3 is employed within a cascaded multi-loop control structure in stationary reference frame. Adoption of the ePLL, presented in [3], enables tracking of PCC voltage amplitude and adds frequency-adaptive feature to the proposed controller while incorporation of the PLL dynamics into the control structure significantly improves the performance of GFM converter against weak grid conditions and system uncertainties.

This chapter is organized as follows. Section 5.2 describes the studied system. The mathematical model of the system is provided in section 5.3. The structures of the controllers are presented in Section 5.4. To validate the controller performance simulation results for various case studies are presented in Section 5.5. Finally, Section 5.6 concludes this chapter.

5.2 Studied System

The single line diagram of the grid-connected VSC and its associated controller is shown in Fig. 5.1. For simplicity, the dynamics of the dc-side are ignored and thus, the dc sub-system is modeled with a constant dc voltage source. Power conversion between the dc and ac sub-systems is carried out using a three-phase VSC consisting of six ideal switches. The line impedance $R_t + j\omega L_t$ is used to interface the VSC to the PCC. The grid impedance $R_s + j\omega L_s$ is considered between the PCC and an ideal three-phase voltage source to represent a weak power grid. The abc to $\alpha\beta$ transformation blocks are applied to output currents and PCC voltages. The PLL measures the amplitude of PCC voltage and then passes its value to the outer control loop. The output of the controller is transformed back to abc -frame on which, the gating signals of the VSC are produced based on the sinusoidal pulse-width modulation (SPWM) method.

5.3 System Modeling

This section, presents the descriptive equations on dynamics of the grid-connected VSC and derives the state-space model which later is used to design the voltage controller.

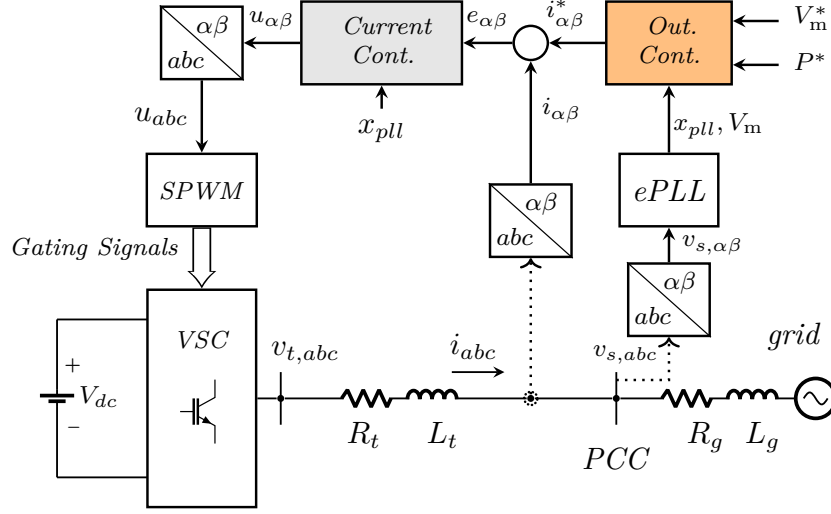


Figure 5.1 Single-line diagram of the system with the controllers

5.3.1 System Modeling

The characteristic equation of the system in Fig. 5.1 in $\alpha\beta$ -frame is

$$v_{s_{\alpha\beta}} = L_g \frac{di_{\alpha\beta}}{dt} + R_g i_{\alpha\beta} + v_{g_{\alpha\beta}}. \quad (5.1)$$

where $i_{\alpha\beta}$, $v_{s_{\alpha\beta}}$ and $u_{t_{\alpha\beta}}$ are the VSC output current, PCC voltage and the VSC terminal voltage, respectively.

Following the method explained in Chapter 3, the PCC voltage can be stated based on the VSC output voltage and current as well as the ideal grid voltage, i.e., $v_{g_{\alpha\beta}}$ such that

$$v_{s_{\alpha\beta}} = \frac{L_t}{L_g + L_t} v_{g_{\alpha\beta}} + \frac{L_g}{L_g + L_t} v_{t_{\alpha\beta}} + \frac{(R_g L_t - L_g R_t) i_{\alpha\beta}}{L_g + L_t}. \quad (5.2)$$

By considering $\xi = L_t(L_g + L_t)^{-1}$, $\sigma = L_g(L_g + L_t)^{-1}$, and $\tau = (R_g L_t - L_g R_t)(L_g + L_t)^{-1}$, substitution of (5.2) into the linear time-invariant (LTI) model of the proposed ePLL in [3], results in the ePLL-integrated model of the ac sub-system which can be expressed in state-space form of

$$\begin{aligned} \dot{x}_p &= A_p x_p + B_p u + F_p W, \\ y &= C_p x_p, \end{aligned} \quad (5.3)$$

where $x_p = [i_\alpha \ i_\beta \ x_{1_{pll}} \ x_{2_{pll}}]^T$, $u = [u_\alpha \ u_\beta]^T = [v_{t_\alpha} \ v_{t_\beta}]^T$, $W = [v_{g_\alpha} \ v_{g_\beta}]^T$, and

$$A_p = \begin{bmatrix} -\frac{R_t + \tau}{L_t} & 0 & 0 & 0 \\ 0 & -\frac{R_t + \tau}{L_t} & 0 & 0 \\ \mu_1 \tau & 0 & -\mu_1 & -\omega \\ 0 & \mu_1 \tau & \omega & -\mu_1 \end{bmatrix}, B_p = \begin{bmatrix} \frac{1 - \sigma}{L_t} & 0 \\ 0 & \frac{1 - \sigma}{L_t} \\ \mu_1 \sigma & 0 \\ 0 & \mu_1 \sigma \end{bmatrix}$$

$$C_p = [I_{2 \times 2} \ 0_{2 \times 2}], \ F_p = \begin{bmatrix} -\xi/L_t & 0 & \mu_1 \xi & 0 \\ 0 & -\xi/L_t & 0 & \mu_1 \xi \end{bmatrix}^T.$$

It could be easily shown that the system in (5.3) is controllable and observable for all system parameters and is minimum phase.

5.3.2 Relationship Between Reactive Power and PCC Voltage

As shown in Chapter 3, under weak grid conditions, the PCC voltage becomes dependent on system parameters and set-points; i.e., injected real and reactive powers. The relationship between reactive power and PCC voltage can be represented by considering the complex power formulation

$$\mathbf{S} = 1.5 \mathbf{V}_{\text{PCC}} \mathbf{I}^*$$

where $\mathbf{V}$ and $\mathbf{I}$ are the voltage and current phasors. Furthermore, the voltage drop between PCC and the ideal grid by neglecting the grid resistance R_g is

$$j\omega L_g \mathbf{I} = \mathbf{V}_{\text{PCC}} - \mathbf{V}_g$$

Thus, the complex power equation could be expressed based on system parameters and PCC voltage such that

$$\mathbf{S} = P + jQ = \frac{3}{2} \left(\frac{|\mathbf{V}_{\text{PCC}}|^2 - \mathbf{V}_{\text{PCC}} \mathbf{V}_g}{-j\omega L_g} \right). \quad (5.4)$$

Given that PCC voltage in its phasor form consists of real and imaginary parts, it could be expressed as

$$\mathbf{V}_{\text{PCC}} = V_R + jV_I$$

Thus, by defining $K = \frac{2}{3}\omega L_g$, (5.4) yields

$$\begin{aligned} KQ + V_R V_g &= V_R^2 + V_I^2 = V_{\text{PCC}}^2, \\ -KP + V_I V_g &= 0. \end{aligned} \quad (5.5)$$

The second part of equation set (5.5) yields:

$$V_I = \frac{KP}{V_g} \quad (5.6)$$

Combining the two equations in (5.6) and the first part of (5.5) results in

$$V_R^2 - V_R V_g + \left(\frac{K^2 P^2}{V_g^2} - KQ \right) = 0,$$

which is a quadratic equation with a solution for V_R in the form of

$$V_R = \frac{1}{2} \left[V_g \pm \sqrt{V_g^2 - 4 \left(\frac{K^2 P^2}{V_g^2} - KQ \right)} \right]. \quad (5.7)$$

In (5.7), given that for $P = Q = 0$, V_R could not be zero, only the positive sign is acceptable and thus

$$V_{PCC} = \sqrt{KQ + \frac{V_g}{2} \left[V_g + \sqrt{V_g^2 - 4 \left(\frac{K^2 P^2}{V_g^2} - KQ \right)} \right]}. \quad (5.8)$$

Based on (5.8), for a given real power set-point, there is a direct relationship between reactive power and amplitude of PCC voltage. Furthermore from (5.8) it is evident that increasing of grid inductance L_g (and thus K), increases the dependency of voltage on system set-points.

Using the system parameters given in Table 5.1, the relationship between PCC voltage and the reactive power is shown in Fig 5.2 where grid inductance is changed from 2 to 12 mH and $Q \in [0 \ 3]$ kVAr. In Fig 5.2, weaker grid (higher values of L_g) is indicated with the red color.

In general, increasing of active power results in a drop in voltage amplitude while injection of more reactive power elevates the voltage. As evident in Fig. 5.2, the $Q - V_{PCC}$ relationship could be approximated with a linear function, which in particular, enables deployment of droop-based compensators.

5.4 Proposed Control Strategy

In this section, the general characteristics and features of the proposed controller are discussed. The structure of the controller is shown in Fig. 5.3 where it employs a nested control structure with two cascaded loop; i) an inner current controller which is a multivariable PLL-integrated current controller developed in Chapter 3, ii) an outer controller consisting of a

Table 5.1 Parameters of the Study System

parameters	symbol	value
dc bus voltage	V_{dc}	600 V
Grid phase voltage	v_s	120 V (rms)
System frequency	f	60 Hz
PWM switching frequency	f_{sw}	5 kHz
RL filter resistance	R_t	0.1 Ω
RL filter inductance	L_t	5 mH
Grid internal resistance	R_g	1 m Ω
Grid internal inductance	L_g	[2:12] mH

PI-based compensator which is responsible for adjusting the PCC voltage and, a reference generator block which produces the current reference signals based on the provided power set-points P_{ref} and Q_{ref} .

5.4.1 Voltage Controller

The voltage controller is developed based on (5.8) and is basically a droop-based PI controller whose output (Q_{ref}) is used as the control input for the inner current controller. The structure of the PCC voltage controller is shown in Fig.5.4 where $K=10000$.

5.4.2 Current Controller

The current controller is implemented in stationary reference frame and it follows the PLL-integrated multivariable design proposed in Chapter 3. The structure of the controller is shown in Fig. 5.3 where it could be seen that current references are generated based on the arbitrary real power set-point P_{ref} as well as the reactive power set-point which is produced by the outer voltage control loop. The controller has a MIMO structure whose design follows the procedure in [37]. Under certain conditions, the controller should (i) result in an asymptotically stable closed-loop system with a fast transient response and (ii) track the reference signals and reject the perturbation while causing minor coupling between control channels [97]. The current controller consists of two parts; a servo-compensator and a stabilizing compensator. The servo-compensator accounts for accurate tracking of the reference signals and disturbance rejection. The stabilizing compensator guarantees closed-loop stability and fast transient response.

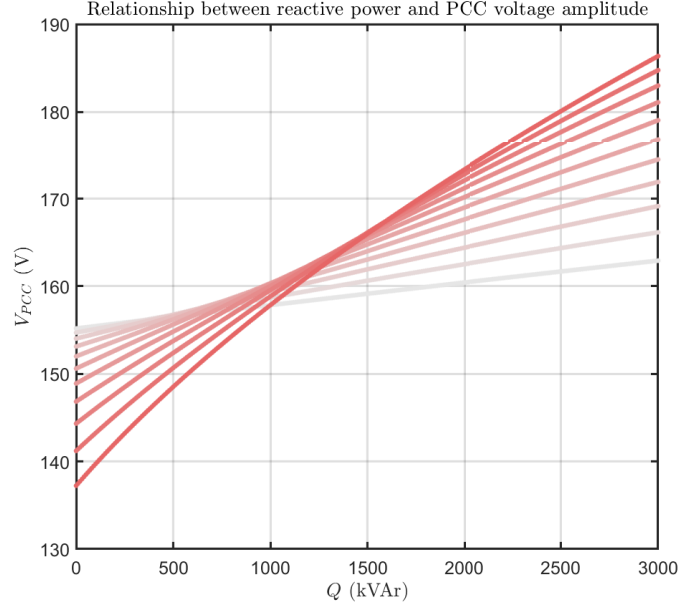


Figure 5.2 The relationship between PCC voltage and reactive power when $L_g \in [2 \ 12]$ mH and $Q \in [0 \ 3]$ kVAr

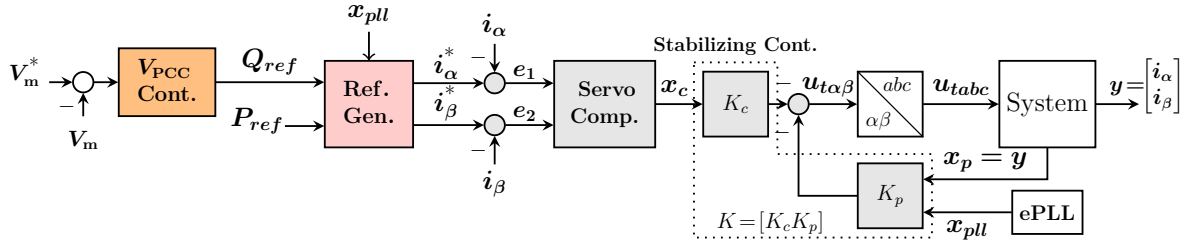


Figure 5.3 Structure of the proposed multi-loop controller

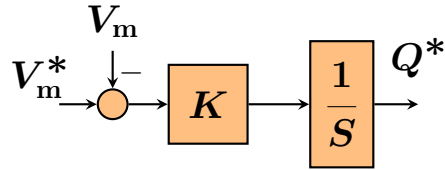


Figure 5.4 The structure of the droop control

The general structure of the proposed controller and its integration into the system is demonstrated in Fig. 5.3 where the reference current signals are generated based on

$$\begin{bmatrix} i_{\alpha}^* \\ i_{\beta}^* \end{bmatrix} = \frac{2}{3} \frac{1}{V_p^2} \begin{bmatrix} x_{1pll} & x_{2pll} \\ x_{2pll} & -x_{1pll} \end{bmatrix} \begin{bmatrix} P^* \\ Q^* \end{bmatrix}. \quad (5.9)$$

The control design directly follows the method presented in Chapter 3 where the dynamics of the errors are added to the system as the new set of control variables x_c . The errors are applied to the servo-compensator whose dynamics can be represented in the state-space form

$$\dot{x}_c = A_c x_c + B_c e \quad (5.10)$$

where $B_c = [0, 1]^T$ and A_c is any (2×2) matrix with the characteristic polynomial in the form of a resonance controller determined by

$$f(\lambda) = \lambda^2 + \omega^2$$

One possible implementation of $A_{c_{2 \times 2}}$ can be $\begin{bmatrix} 0 & \omega \\ -\omega & 0 \end{bmatrix}$, in which (5.10), transforms into:

$$\begin{aligned} \dot{x}_{c1} &= \omega x_{c2} \\ \dot{x}_{c2} &= -\omega x_{c1} + \underbrace{(i_{\alpha\beta}^* - i_{\alpha\beta})}_e. \end{aligned} \quad (5.11)$$

Substituting (5.9) into (5.11) readily yields

$$\dot{x}_{c2} = -\omega x_{c1} + ([\frac{2}{3V_p^2}(x_{1pll}P + x_{2pll}Q)] - i_\alpha) \quad (5.12)$$

and thus the augmented system by integration of UTSP states is:

$$\begin{aligned} \dot{x}_{c1\alpha} &= \omega x_{c2\alpha} \\ \dot{x}_{c2\alpha} &= -\omega x_{c1\alpha} + ([\frac{2}{3V_p^2}(x_{1pll}P + x_{2pll}Q)] - i_\alpha) \\ \dot{x}_{c1\beta} &= \omega x_{c2\beta} \\ \dot{x}_{c2\beta} &= -\omega x_{c1\beta} + ([\frac{2}{3V_p^2}(x_{2pll}P - x_{1pll}Q)] - i_\beta) \\ \frac{di_\alpha}{dt} &= -\frac{R_t}{L_t}i_\alpha + \frac{1}{L_t}u_\alpha - \frac{1}{L_t}[Av_{grid,\alpha} + Bu_\alpha + Ci_\alpha] \\ \frac{di_\beta}{dt} &= -\frac{R_t}{L_t}i_\beta + \frac{1}{L_t}u_\beta - \frac{1}{L_t}[Av_{grid,\beta} + Bu_\beta + Ci_\beta] \\ \dot{x}_{1,pll} &= \mu_1 (Av_{grid,\alpha} + B(u_\alpha) + Ci_\alpha) - \mu_1 x_{1,pll} - \omega x_{2,pll} \\ \dot{x}_{2,pll} &= \mu_1 (Av_{grid,\beta} + B(u_\beta) + Ci_\beta) - \mu_1 x_{2,pll} + \omega x_{1,pll}. \end{aligned} \quad (5.13)$$

Eq. (5.13) describes an augmented system with the state variable vector $x = [x_c^T, x_p^T, x_{pll}^T]^T$, which includes the plant's state variables as well as dynamics of the error vector and PLL's states.

Thus, a state feedback principle:

$$u = -Kx \quad (5.14)$$

can be employed to assure the robustness of the closed-loop system. As discussed in Chapter 3 and shown in Fig. 5.3, the controller gain $K_{2 \times 8}$ is a combination of the gain matrix $K_{c2 \times 4}$ which operates on the added control states (x_c), the plant's state feedback gain, $K_{p2 \times 2}$, and the PLL's state feedback gain, $K_{pll2 \times 2}$ such that $K = [K_c K_p K_{pll}]$. To meet the design requirements, the same LQR method is employed to minimize the cost function of J

$$J = \int_0^\infty (q_{1,3} x_{c1,\alpha\beta}^2 + q_{2,4} x_{c2,\alpha\beta}^2 + q_{5,6} x_{p\alpha\beta}^2 + q_7 x_{pll,1}^2 + q_8 x_{pll,2}^2 + r_1 v_1^2 + r_2 v_2^2) dt. \quad (5.15)$$

It could be studied that the cost function in (5.15), is the same one utilized in Chapter 3 and thus, applying the same pole placement technique results in the feedback gain matrix

$$K = \left[\underbrace{\begin{bmatrix} -7.9e5 & -1.0e4 & -5.0e4 & -381 \\ 1.15e4 & 412.5 & -8.8e5 & -1e4 \end{bmatrix}}_{K_c} \mid \underbrace{\begin{bmatrix} 23.8 & 0.59 & 1.68 & -0.19 \\ -0.99 & 22.9 & 0.17 & 1.35 \end{bmatrix}}_{K_p} \right].$$

5.5 Performance Validation

In this section, the performance of the proposed controller is evaluated when applied to the DER system of Fig. 5.1 with parameters given in Table 5.1. The case studies are carried out using numerical time-domain simulations in MATLAB software. Like the previous chapters, to highlight the effect of PLL dynamics inclusion into the control structure, two control systems are considered; i) A controller which only employs state feedbacks from plant (without considering the effect of PLL), which is called the conventional controller, and ii) The proposed controller which incorporates feedbacks from PLL state variables. To verify the performance criteria of the proposed controller, various case studies are studied. In the following case studies, the transient behaviour of the conventional and the proposed controllers with respect to changes in voltage and power set-points are investigated under stiff, weak and extremely grid conditions.

5.5.1 Stiff Grid Conditions

In this case-study, the dynamic response of the conventional and proposed systems are studied under stiff grid conditions. For this purpose, the grid inductance L_g is selected to be 2 mH and thus the system is operating under stiff grid conditions. the set-points for the amplitude of PCC voltage, v_m^* , and real power, P^* , are initially set at 180 V, and 4 kW, respectively. In $t=0.3$ s, real power set-point is stepped up to 8 kW through a step change, where based on the short circuit ratio definition

$$SCR = \frac{1.5 V_{PCC}^2}{\omega_0 L_g S_{VSC}}, \quad (5.16)$$

results in: $SCR \approx 7.2$. Subsequently, in $t=0.4$ s, v_m is dropped to 145 V ($SCR \approx 4.0$). The dynamic response of the two systems are shown in Figs. 5.5 and 5.6, respectively. It can be seen that the transient behavior of the two systems in terms of PCC voltages, currents and real power is smooth, rapid and with no overshoots. Although, the power quality of the proposed controller is slightly better. However, it should be noticed that under stiff grid conditions, PCC voltage regulation becomes challenging given that much more control effort (in terms of reactive power) is required; a fact that has been shown in Fig. 5.2.

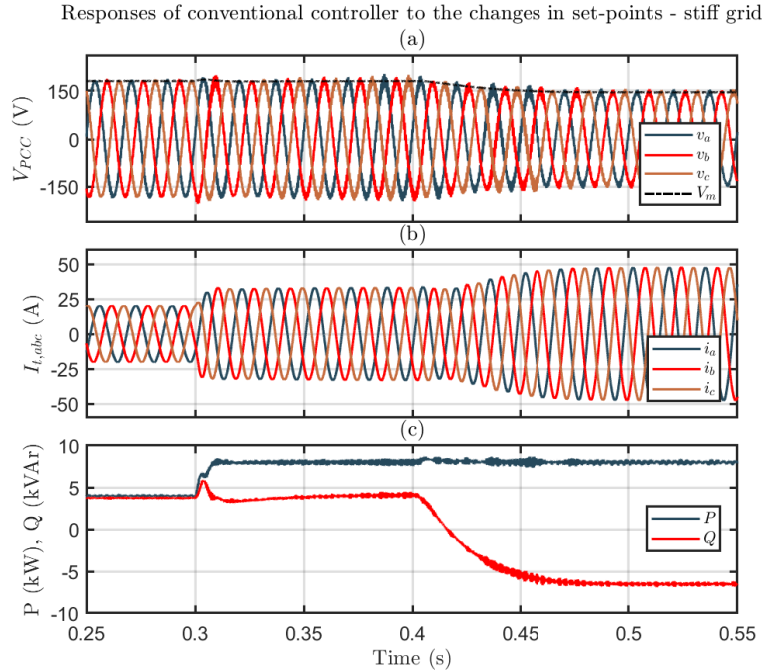


Figure 5.5 Dynamic response of the conventional controller to changes in set-points of real power and PCC voltage. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

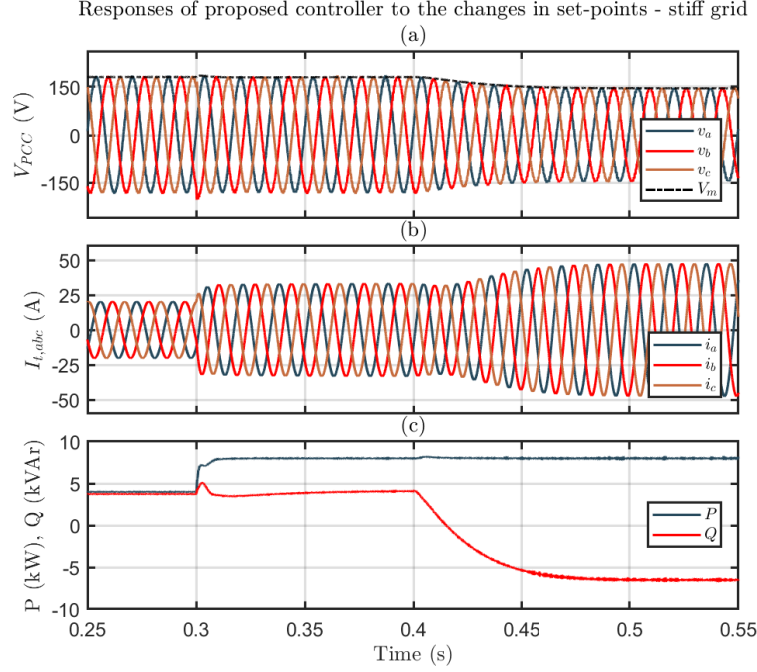


Figure 5.6 Dynamic response of the proposed controller to changes in set-points of real power and PCC voltage. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

5.5.2 Weak Grid Conditions

The grid is weakened by selecting a larger grid inductance, i.e., $L_g=8$ mH. The same scenario as the previous case-study is then repeated for both control systems where real power and PCC voltage references are changed through step-wise functions, eventually resulting in $SCR \approx 1.3$. Figs. 5.7 and 5.8 show the responses of the conventional and the proposed controllers, respectively. From Figs. 5.7 and 5.8 it is evident that the proposed controller provides a better dynamic response in terms of settling time and overshoot.

5.5.3 Extremely Weak Grid Conditions

To show the capability of the proposed controller to cope with extremely weak grid conditions ($SCR=1$), the same scenario as in the previous case-studies are repeated with grid inductance L_g of 10 mH. Given the steady-state conditions of $P=8$ kW and $V_m=145$ V, and considering the SCR definition in (5.16), this yields unity SCR ($SCR=1$). Under such conditions, the maximum power capacity of the ac system is utilized; i.e., any further increase in real power or decrease in line voltage, results in system instability. The responses of the conventional and proposed controllers are shown in Figs. 5.9 and 5.10, respectively. It is evident that the conventional controller struggles to adjust the real power at 8 kW and eventually becomes

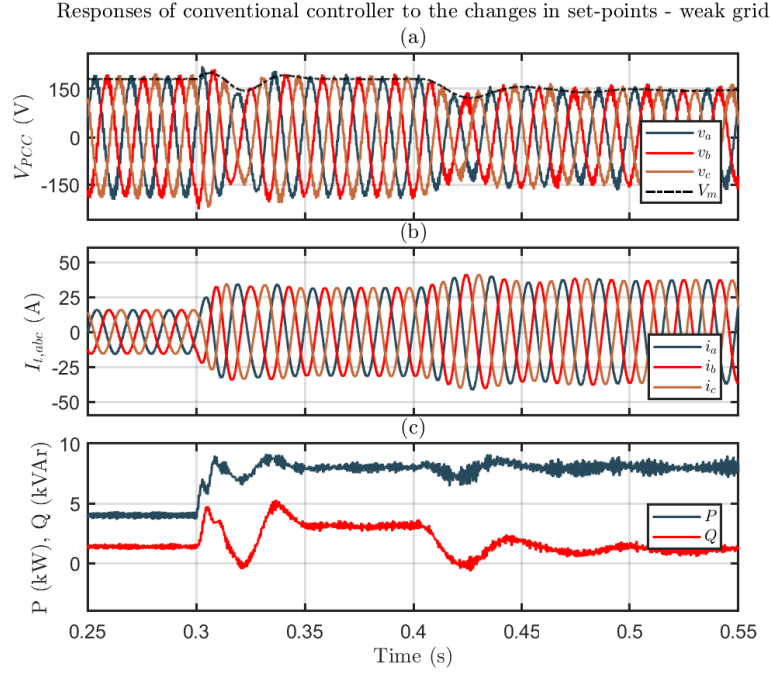


Figure 5.7 Dynamic response of the proposed controller to changes in set-points of real power and PCC voltage. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

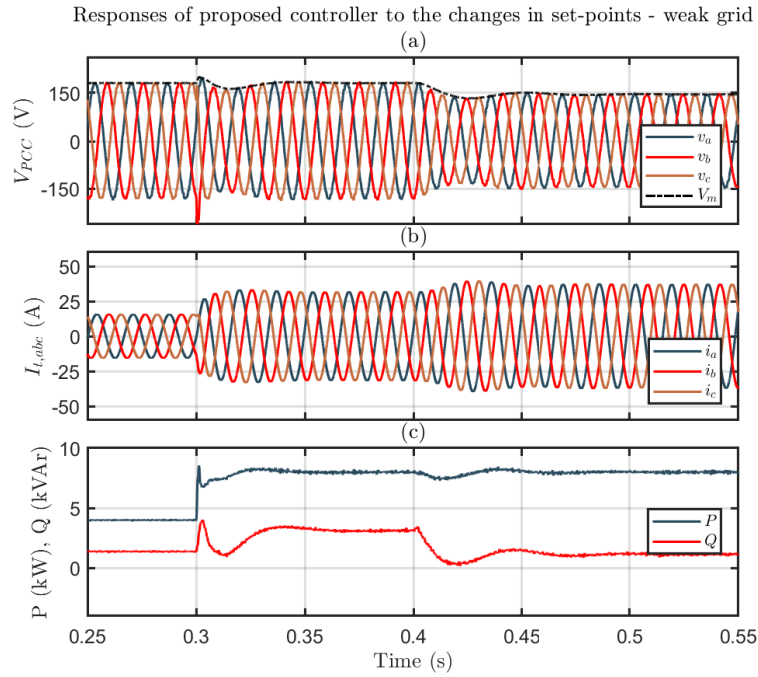


Figure 5.8 Dynamic response of the proposed controller to changes in set-points of real power and PCC voltage. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

unstable at $t=0.45$ s, when V_m^* drops to 145 V. The proposed controller however, successfully follows the reference signals smoothly and without any significant oscillations.

5.6 Conclusion

In this chapter, an application of the the proposed current controller (developed in Chapter 3) for the GFM voltage source converter is provided.

The objectives of the controller is to i) adjust the real power on its provided set-point value, ii) regulate the amplitude of PCC voltage (and hence GFM), and iii) improve system performance under weak grid conditions. The controller utilizes a cascaded multi-loop structure which employs a simple droop-based PI compensator as the outer control loop, while incorporating the advanced current controller of Chapter 3 in its inner loop. The advanced current controller deploys feedbacks from PLL state variables and applies them on the gain matrix whose elements are optimally designed based on the LQR approach.

In order to show the effectiveness of the proposed control scheme, its performance is contrasted with a conventional controller which lacks PLL state feedbacks in its structure. Through different software simulated case-studies and under different SCR levels, it is shown that incorporation of PLL dynamics within the inner current controller, substantially enhances system performance and results in improved stability margin of the system under weak and extremely weak grid conditions.

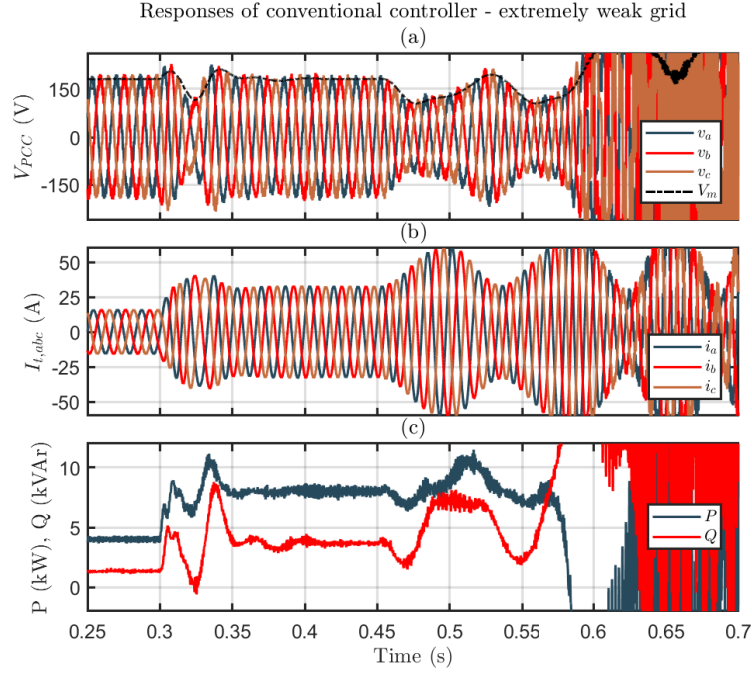


Figure 5.9 Dynamic response of the conventional controller to changes in set-points of real power and PCC voltage under extremely weak grid conditions. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

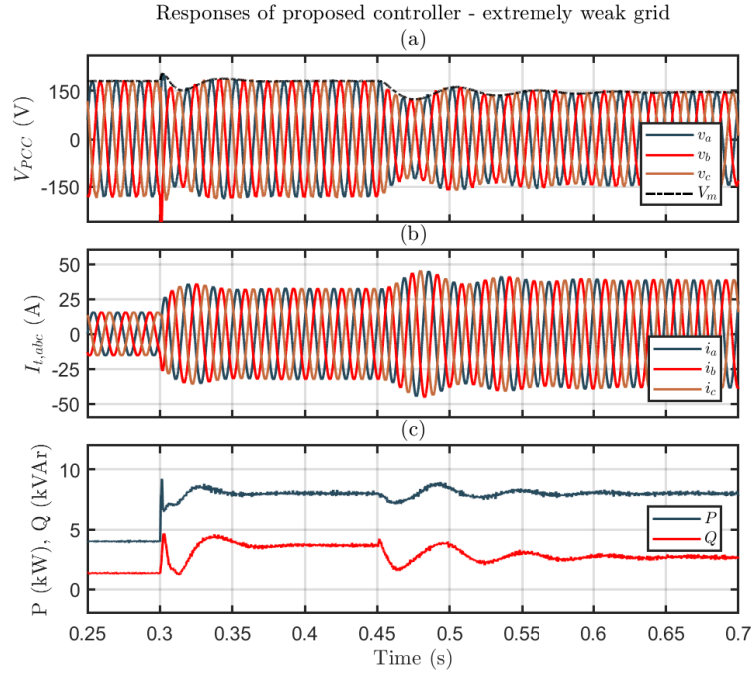


Figure 5.10 Dynamic response of the proposed controller to changes in set-points of real power and PCC voltage under extremely weak grid conditions. (a) PCC voltages, (b) VSC currents, and (c) injected real and reactive powers

CHAPTER 6 CONCLUSION

6.1 Summary

In this research, robust control of grid connected voltage source converters (VSCs) under weak and unbalanced grid conditions is comprehensively studied and a new multivariable control approach is proposed to enhance performance and improve stability of these grid-connected VSC-based systems.

New studies have revealed that the phase-locked loop (PLL) is the culprit of system stability degradation under weak grid conditions. Thus, recently, several publications have tried to address this problem by presenting various control strategies which are discussed and categorized in the first chapter of this thesis. In Chapter 2, the newly promoted PLL-less method is rigorously investigated. This method which is also known as the direct power control (DPC) in the literature, aims to improve the performance of PLL-based VSCs under weak grid conditions but without using a PLL and just by employing two second-order band-pass filters. In this thesis, however, it is shown that under weak grid conditions, despite its claim, PLL-less method is prone to the same stability issues as the PLL-based VSCs. Further, it is also shown that under certain operating conditions, PLL-less method suffers from performance limitations such as steady-state tracking errors under grid frequency drifts and distorted output currents under voltage imbalance. The thesis further proposes a method to overcome the later problem.

Chapter 3 of this thesis proposes a new robust multivariable control scheme in stationary reference frame which results in improved performance of VSCs under weak grid conditions. First, the dynamics of an enhanced PLL (ePLL) are incorporated in the characteristic equations of a grid-connected VSC in stationary reference frame to derive a linear-time-invariant (LTI) model. A robust multivariable controller with optimally designed gains is then applied on the system model. The proposed controller considers the PLL as part of the plant and employs the PLL state variables into its feedback structure. It is shown that the proposed approach substantially improves system performance under weak and extremely weak grid conditions. Further, the formulation of short circuit ratio (SCR) is also revisited in Chapter 3 and the constraints on minimum reactive power injection for voltage stability under weak grid conditions is revealed. Moreover, it is shown that the proposed controller extends the stability margins of the VSC to the unity short circuit ratio (i.e., $SCR=1$).

Extending the work of Chapter 3, Chapter 4 of this thesis presents a new controller in stationary frame for enhanced performance of VSCs under weak and unbalanced grid conditions. The controller incorporates a nested multi-loop structure consisting of an inner PLL-integrated current controller as well as an outer dc-link voltage control loop based on integral-resonant control mechanism. The controller neutralizes second-order oscillations on the dc-link voltage as well as third-order harmonics on the output currents under voltage imbalance while improving system performance and stability under weak grid conditions. First, the state-variables of an ePLL are fully integrated into the system's model. A robust multivariable control technique is then employed to improve the weak grid performance of the system. Furthermore, the dc-link dynamics under unbalanced conditions are revisited. It is shown that in the power balance equation, there are nonlinear coupling terms between different mechanisms of the dc-link control loop. Further, an approach is proposed to surpass these terms by separating dc voltage dynamics.

Finally, Chapter 5 presents an application of the proposed controller in grid forming (GFM) VSCs where the amplitude of the ac voltage at point of common coupling is adjusted in a nested multi-loop control structure. In this chapter it is shown that deploying the proposed current controller results in major performance improvement under weak grid conditions to the extent that the VSC is able to injects real power to the grid with the full power capacity of the line (i.e., SCR=1).

6.2 Future Work

In continuation of this work and to further expand the methods and ideas discussed in this thesis, the following topics are suggested for the future research:

1. **Extending the control approach to applications with LC or LCL filter:** In this thesis, the dynamics of an ePLL is integrated into the characteristic equations of an ac-subsystem which employs L-filters. However, utilization of higher order power filters such as LC (e.g., GFM application) or LCL filters (e.g., harmonic suppression purposes) can further improve system performance. To this end and in order to integrate the state variables of the ePLL into the plant model, the system equations should be revisited.
2. **Studying the impact of weak grid on the outer control loops:** As discussed in this thesis, the proposed current controller substantially improves system performance under weak grid conditions. The effects of the PLL under weak grid conditions however, has not been studied on the outer control loops (including dc-link voltage and PCC

voltage controllers). Analyzing and compensating such effects, nonetheless, can elevate system performance and stability to a further extend.

3. **Integration of dc-link voltage controller within the multivariable control scheme:** Given that the proposed current controller in this thesis is implemented in stationary frame, control of dc-link voltage is realized using a cascaded multi-loop structure. Despite the effectiveness of this approach, however, the dynamic of dc voltage control is constrained by the bandwidth of the inner current control loop. This constraint could be resolved by developing a LTI model of the plant which includes the dynamics of the dc-link in a linear form.
4. **Current limiting features under voltage imbalance:** One of the benefits of the nested multi-loop control structure in this thesis is its ability to limit the current reference signal during faults and voltage sags. Such feature however has not been studied during the voltage imbalance which could make a very interesting research topic with practical applications in industry.

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