

From State-of-the-Art Grid-following Converters to Grid-forming Converters

by

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Abstract

This thesis investigates the performance of the state-of-the-art decoupled current controller widely used in grid-following (GFL) inverter-based resources (IBRs). Small-signal analysis is used to evaluate the sensitivity of key control parameters and the stability limits of the controller under ultra-low short-circuit ratio (SCR) conditions, while large-signal simulations are conducted to examine its limitations during islanded operation. The results show that although the decoupled current controller demonstrates strong small-signal performance under weak-grid conditions, it has a fundamental stability limit at extremely low SCR and cannot maintain stable operation in islanded mode.

Motivated by these limitations, this thesis further develops and investigates simple, yet effective, control topologies that can enhance the grid-firming capability of a conventional grid-following inverter or transform it to a grid-forming (GFM) inverter. A minimalistic grid-forming control topology is proposed and evaluated through both small-signal analysis and large-signal simulations. To further improve stability and dynamic performance, a simplified virtual admittance loop is incorporated, which enhances the performance of the proposed GFM controller.

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List of Abbreviations

GFM	grid-forming inverter
GFL	grid-following inverter
EMT	electromagnetic transients
IBRs	inverter-based resources
VSC	voltage source converter
PLL	phase-locked loop
SCR	short-circuit ratio
IGBTs	insulated-gate bipolar transistor
MOSFETs	metal-oxide-semiconductor field-effect transistor
PWM	pulse-width modulation
STATCOM	static var compensators
SMIB	single machine infinite bus
CCM	component connection method
VA	virtual admittance
PPC	power plant controller
POI	point of interconnection
DER	distributed energy resources
PSCAD	power system computer-aided design
HVDC	high-voltage direct current
BESS	battery energy storage system
RLC	resistive inductive capacitive
P,Q	active, reactive power
X/R	reactive impedance over resistive impedance

Chapter 1

Introduction

1.1 Background

As part of green initiatives for sustainable energy systems, many countries have chosen renewable energy resources in fulfilling new energy demand and replacing fossil fuel-based generation. These renewable generations are primarily based upon solar and wind power and their integration to the utility grid is generally through the voltage source converter (VSC) technology. Despite the many environmental benefits of renewable energy generation, the increasing levels of VSC penetration put a strain on the grid's stability and reliability. These renewable resources, also referred to as inverter-based resources (IBRs), are mostly controlled as grid-following (GFL) inverters. Unlike the traditional gas or coal generators, which typically use synchronous machines, the GFL IBRs are sensitive to the grid's strength and heavily rely on the nearby synchronous machines to provide inertia in voltage angles and magnitudes. In a grid that has diminishing synchronous machine-based generation and is low in short-circuit levels, the GFL IBRs are prone to have challenges such as voltage and angle stability issues, unwanted interaction between IBRs, and, under worst case conditions, blackouts.

Facing these challenges, grid-forming (GFM) IBRs are gaining popularity in recent years. GFM has promised to synthesize inertia similar to a synchronous machine and to provide stability and damping to the grid. There have been many types of GFM control topologies under research in the past few years. Some topologies have been developed and moved into implementation. The few key topologies that have been considered by major inverter manufacturers include the droop-based control and virtual synchronous

machine-based control. Although the exact implementation details are not fully disclosed due to vendor restrictions, the basic concepts for these types of control can be found in publicly available literature as listed below:

- The droop control is based on the traditional coupling characteristics of a power system that has a mostly inductive network and rotating machines (voltage vs. Q ; angle vs. P). The GFM under droop control follow these characteristics and help to maintain the grid frequency and voltage by supplying power and reactive power respectively. The droop control is further developed in a few categories such as frequency-based droop control [1] and power synchronization control [2].
- Virtual synchronous machine-based control mimics the operation of a synchronous machine and emulates the inertial response similar to a mechanical machine [3][4].

There are various other types of control topologies, such as virtual oscillator control [5] or virtual synchronous control utilizing DC-link capacitor dynamics [6]. Although these emerging control topologies appear to be promising they are still under research and development.

To further improve the current-limiting capabilities of the GFM, additional inner control loops must be considered. The typical methods for GFM current limiting are the current limiter method [7] and virtual impedance method [8]. The current limiter method can continue to leverage the state-of-the-art decoupled current control methodology [9]. This makes the converter behaves as a current source and is extremely good at current limiting. However, during a fault when the inner current control loop is limited by the maximum current threshold, it can cause potential wind up for the outer GFM control loops [10][11]. The virtual impedance method allows the inverter to behave more similarly to a voltage source, which improves the fault recovery and mitigates the controller wind up. However, it has lower current-limiting performance in comparison to the decoupled current controller method and the designer must balance current-limiting ability and control stability objectives [12].

Design and analysis of the multiple layers of the inner-loop and outer-loop controls for GFM can also be challenging especially under various operating and grid conditions. The complexity in GFM control topologies can reduce the reliability of the GFM converter and hinders the public's confidence in its wider adoption. Furthermore, for the existing GFL inverters to adapt to these multi-layer controls can be nearly impossible due to the incompatibility in control topologies.

1.2 Problem Definition

Most existing GFL inverters utilize the state-of-the-art decoupled current control method and a phase-locked loop (PLL) as the inner control loop because of its superior current-limiting capability. This design topology has been widely implemented and matured over the years. The decoupled current controller method has been designed for stiff grid conditions, and much effort has been made to improve its performance at low short-circuit ratio (SCR) levels of around 1.2 – 3. However, there has been a void in research for the decoupled current controller operating under ultra-low SCR values between 0 to 1. For the GFM topologies that still rely on the decoupled current controller, the operating SCR can often be in the ultra-low range. This thesis addresses this void in research and provides analysis and insight into GFL control stability under ultra-low SCR operating conditions.

Furthermore, there have been many challenges to develop the GFM control around the existing GFL control topologies. The simplest method for adaptation would be the droop-based control, but it has problems of control windup and lack for inertial support [10][11]. A VSM control can provide inertia responses similar to a real synchronous machine but the outer loop design and the integration to the decoupled current controller can often be extremely sophisticated [12].

Based on the initial in-depth analysis of GFL control stability, this thesis proposes and investigates control strategies for both GFM and GFL operation. The approaches emphasize minimalistic design and straightforward implementation in existing PLL-based GFL inverters with state-of-the-art decoupled current control.

1.3 Thesis Motivation

The GFM algorithm is still undergoing extensive research and the actual GFM performance from some of the major inverter manufacturers is still to be tested in the real world. Two of the major motivators for this thesis to accelerate the adaptation of GFM are to (i) start from well-established and tested foundations, and (ii) develop further with the simplest path.

From an engineering design perspective for power systems where reliability and public safety are the utmost objectives, every step of innovation requires extensive study, analysis, and testing. Therefore, starting from a well tested and reliable control layer, e.g., the decoupled current controller, is highly beneficial.

Furthermore, the saying “*the simplest solution can often be the best*” from Occam’s Razor could be an ideal, but it does suggest clarity over sophistication. A simple or minimalistic solution can often be easier to understand, assess, test, and implement. Therefore, the other motivator for this thesis is to develop and investigate control strategies with a minimalistic design while maintaining key performance requirements.

1.4 Organization of the Thesis

The remainder of the thesis is organized into the following chapters.

- Chapter 2 introduces common inverter topologies and discusses the state-of-the-art GFL decoupled current controller and its small-signal analysis.
- Chapter 3 discusses the limitations of GFL under islanded conditions. A GFM topology is proposed to overcome the limitations.
- Chapter 4 shows the performance and limitations of the proposed GFM topology. A simplified virtual admittance loop is implemented to further improve the stability of the proposed GFM topology.

- Chapter 5 examines the impact of the simplified virtual admittance loop on the stability of GFL controllers.
- Chapter 6 concludes the thesis and suggests avenues for future work.

1.5 Thesis Scope and Assumptions

In this thesis, the DC-side representation of an inverter is approximated and not considered in detail. For time-domain studies, the DC side is modeled as an ideal voltage source behind an impedance coupled with the converter's internal capacitance. For small-signal analysis, the DC side is treated as a stiff voltage source. Based on industry experiences, such an approximation is reasonable for battery or PV systems with the energy resources directly coupled with the DC-link capacitor. Under such system conditions, a well-designed DC-side voltage controller and sufficiently large DC-link capacitors can reasonably decouple the DC-side dynamic from the AC side. However, for more sophisticated IBRs with buck-boost converters or ones with a poorly designed DC circuit, the DC-side dynamics do affect the AC side, and a more nuanced modeling approach is required [14]. However, these were not the focus of this thesis.

Chapter 2

Inverter and GFL Control Methods

2.1 VSC Technology, PWM, and Filtering

Upon the new innovations in high power semiconductor devices such as IGBTs or MOSFETs, fast switching in power-system applications has become possible and has led to a revolution in power electronics. The crown jewels of the modern power electronics are the two-level [15] and three-level [16] VSCs that power most of the renewable energy resources and modular multilevel converters (MMCs) that empower most recent high voltage DC transmission projects [17].

The key technique behind all the modern power electronic devices is pulse-width modulation (PWM). This technique creates an AC waveform from a fixed DC source by modulating the duty cycle, or the ratio of the “on” time to the total period of switching. The duty cycle is derived from the comparison of a low-frequency reference signal with a high-frequency carrier signal. Figure 2.1 depicts the working principles of the PWM. The reference signal’s frequency is 5 Hz, and the carrier signal frequency is 100 Hz. The ratio of the reference’s amplitude to the carrier’s amplitude in this example is 0.8, which is also called the modulation index. The modulation index directly controls the fundamental component of the output voltage. The PWM operation produces the desired voltage at the reference frequency; however, there are unwanted harmonics at the multiples of the reference signal’s frequency (or switching frequency) such as 100 Hz, 200 Hz, etc.

To reduce the harmonic contents of the inverter’s PWM output voltage, a physical filter is required at the inverter terminal. For inverter control, it is crucial to further remove the unwanted harmonics from the measured voltage and current, and digital filters are typically implemented. Based on industry experiences,

second-order LC filters or third-order LCL filters are popular due to their low cost and acceptable performance. For more sophisticated filters designs, notch filters are used and typically tuned at the first and second multiples of the inverter switching frequency. Additional low-pass filters further remove higher order harmonics.

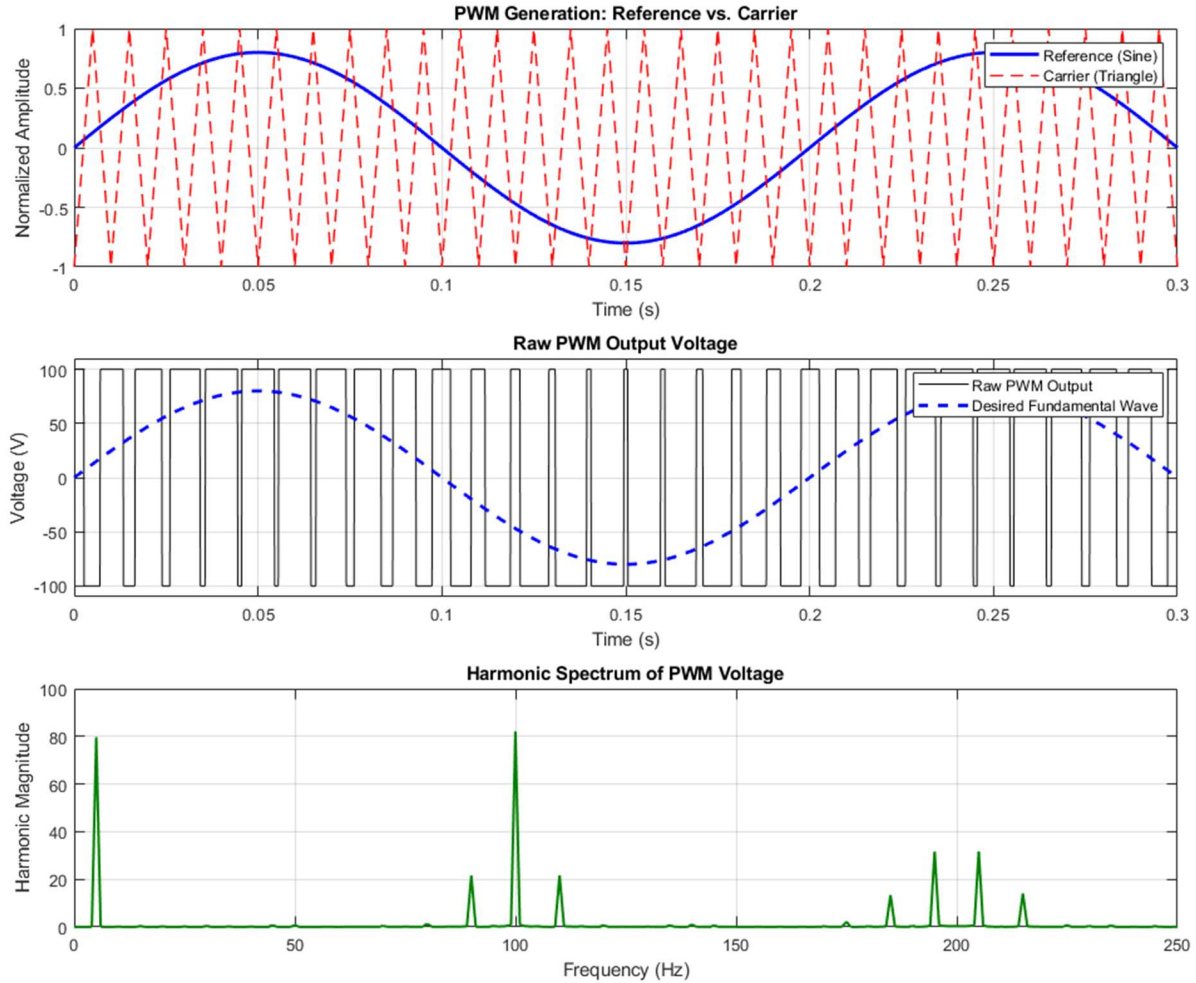


Figure 2.1: Working principles of the PWM under an ideal condition

In addition to the physical filter, a layer of digital filters for the measured voltage and current is implemented to further remove the unwanted noise. The digitally filtered voltage and current will be used by the controller. For the digital filter design, the design principle can be similar to the ones from the physical filter design. A notch filter is typically arranged around the switching frequency to further reject the

harmonic emission from the inverter. A low-pass filter can be further added with a cut-off frequency around the switching frequency to remove any higher order harmonics. The digital filter does come with a few extra degrees of freedom to allow the gain around the fundamental frequencies to be as close to unity as possible, but additional consideration for anti-aliasing is also required. The inverter used in this thesis has a fundamental frequency of 50 Hz (reference frequency) and a carrier frequency of 3000 Hz. An example digital filter design bode plot is shown in Figure 2.2 and the key features of this digital filter are listed below.

- The gain remains close to unity at low frequency
- A notch at 3000 Hz rejects the harmonic noise from the switching
- A steady decline in gains beyond 3000 Hz rejects higher order harmonics

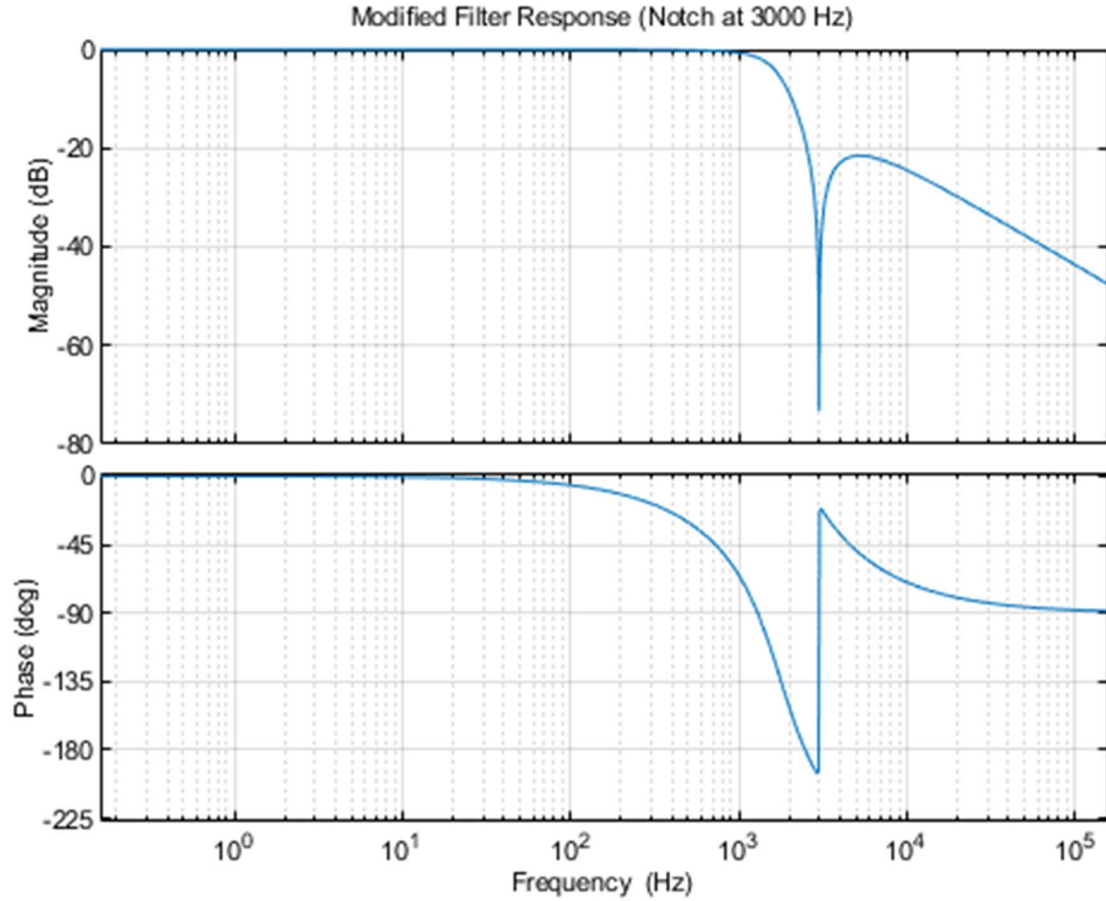


Figure 2.2: Example digital filter design bode plot

2.2 Principles of Grid-following Converter

The decoupled current controller has become widely adopted in VSC-based applications due to its flexible control of power flow and enhanced ability to limit the current. This control methodology was first proposed over three decades ago by C. Schauder and H. Mehta in a paper discussing advanced control of static VAR compensators (STATCOM) [9].

The decoupled current controller, also referred to as vector control or dq control, overcomes the time-varying nature of the sinusoidal AC power systems. This control methodology simplifies the AC time-varying quantities to DC stationary quantities through a PLL and various other mathematical

transformations. The control quantities are further decoupled between the active power and reactive power paths, which improves the flexibility and stability of the control.

Section 2.2.1 and 2.2.2 discuss mathematical transformations and core principles of a PLL, respectively.

Section 2.2.3 shows the decoupling nature of the control.

2.2.1 Key Transformations in Decoupled Current Controller

In decoupled controller, the three-phase sinusoidal quantities are converted into two DC signals for better control. The Park transformation is a mathematical transformation at the core of the control methodology. To better understand the Park transformation, it is necessary to break it down to two separate processes, the Clark transformation and the dq0 transformation,

The three-phase power system voltages can be viewed as 3-dimensional time varying signals. Assuming that the three voltage signals are independent quantities, one can imagine these voltages as the lengths of three orthogonal vectors as shown Figure 2.3. The three vectors of voltages can be further added spectrally and form a combined vector shown in grey. If the system is balanced and has minimum harmonics and zero sequence components, the sphere formed by the vector sweeping through space would cut through the origin of the coordinate system.

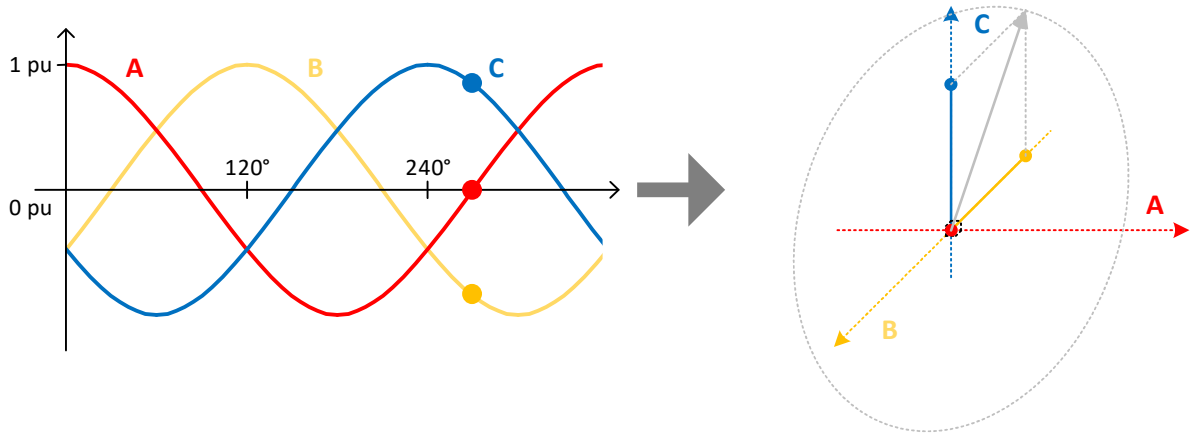


Figure 2.3: Three-phase voltages and their 3d projection.

Clarke transformation:

The three-phase voltages can be spatially rotated like a triangular pyramid where the base is pointed up towards the third dimension as shown in Figure 2.4. This is a linear transformation commonly known as Clarke transforms [18][19]. The zero-sequence component through this transformation is projected onto the vertical axis (0 axis) whereas the locus of the positive-sequence and negative-sequence components is on the 2D plane ($\alpha\beta$ plane). Although it was assumed that the three-phase voltages are independent quantities for general applications, the zero-sequence current is typically blocked at the inverter terminal. This is through various grounding technics of the inverter and the zero-sequence current is naturally suppressed. Without the need to control the zero-sequence current, the zero-sequence voltage can also be removed from control. Since the zero-sequence components only appear on the 0 axis, the projection of the 3D vector onto the $\alpha\beta$ plane is used in this control strategy as shown on the right graphs in Figure 2.4. The locus of combined vector transformed as a rotating circle from the original 3D plane to the 2D plane.

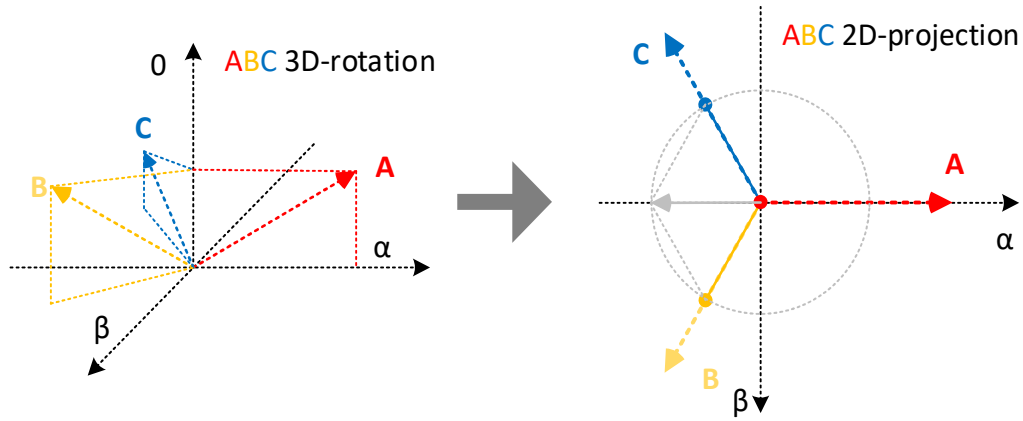


Figure 2.4: Demonstration of the $\alpha\beta$ transformation

Dq0 transformation:

As shown from Figure 2.4, the locus of a balanced positive-sequence voltage signals is a circle. This is because the combined vectors of the three-phase voltage have constant magnitudes but rotate clockwise on the $\alpha\beta$ plane at fundamental frequency. Furthermore, the locus of the negative-sequence component is also a circle on the $\alpha\beta$ plane, but the direction of its rotation opposes that of the positive sequence as depicted on the left of Figure 2.5. The locus of different harmonics is also a circle on the $\alpha\beta$ plane with the rotating speeds of multiples of fundamental frequency depending on the orders of the harmonic. Assuming that the negative-sequence component and harmonics are negligible, the combined vector of the three-phase voltage is a constant rotating vector on the $\alpha\beta$ plane as shown in the middle of Figure 2.5. Since the vector is rotating in a predictable speed typically near fundamental, $\alpha\beta$ plane can also be rotate in such speed so relative to the rotating plane, the vector appears to be stationary as shown on the right of Figure 2.5. The angle θ depends on the relative angle difference between the rotating vector and the rotating plane. This new rotating frame is the dq0 frame, and the entire transformation discussed thus far is called Park Transformation [20].

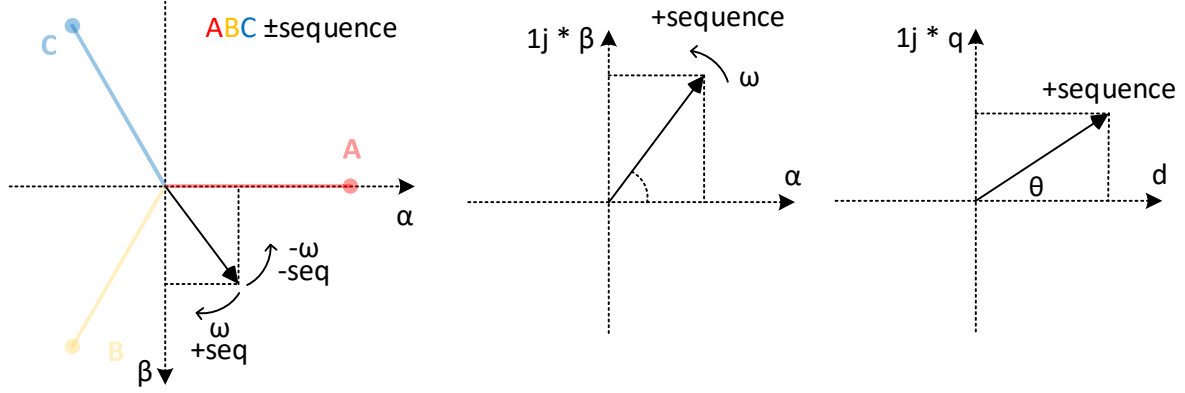


Figure 2.5: Demonstration of the dq transformation

The preceding graphs serve to demonstrate the fundamental concept of linear transformation and rotation, which enables the conversion of three time-varying signals into two stationary signals. This simplified representation does not encompass all the practical details of the transformation. The actual implementation requires further steps, including scaling and the selecting the direct or quadrature (dq) axes to align with the alpha-beta ($\alpha\beta$) axes. Although several variations of these transformations are available, the key concept is the same. This thesis will employ the formulation presented below in (2.1). These two linear transformations are shown below with the left matrix being the Clarke transformation. The Park transformation is the Clarke transformation adding a rotating frame around angle θ . The next section further discusses tracking of the angle, θ .

$$\begin{bmatrix} V_d \\ V_q \\ V_0 \end{bmatrix} = \begin{bmatrix} \cos(\theta) & \sin(\theta) & 0 \\ -\sin(\theta) & \cos(\theta) & 0 \\ 0 & 0 & 1 \end{bmatrix} \cdot \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \cdot \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \quad (2.1)$$

2.2.2 Phase Locked Loop (PLL)

In power system applications, the dominant component is the positive sequence-fundamental component. Engineers, when designing controllers, typically track the angle of the positive-sequence fundamental component. This task is typically handled by a separate unit called a phase-locked-loop (PLL) [21][22]. The PLL smooths out the distortions in the phase angle and extracts the angle of the positive-sequence component.

The angle of rotation on the $\alpha\beta$ plane is essentially an ever-growing ramp signal. To track this ramping signal, the order of the control system must be at least two for the steady state error to be zero [23]. A simple second-order PLL is shown in Figure 2.6. The control error to the PI controller is the angle of the system vectors projected onto the rotating frame as V_d and V_q . The angle is simply calculated using the atan2 command. This is a deviation from the topology originally proposed PLL in [21][22]. The atan2 command is more computationally expensive when calculating the error, but it does produce a linear relationship between the difference and the output error. If the angle is driven to zero, the input system vectors align with the d-axis of the rotating frame and the projection on the q-axis is zero. In some applications, instead of minimizing the angle, V_q is directly targeted to be driven to zero, which yields the same end result.

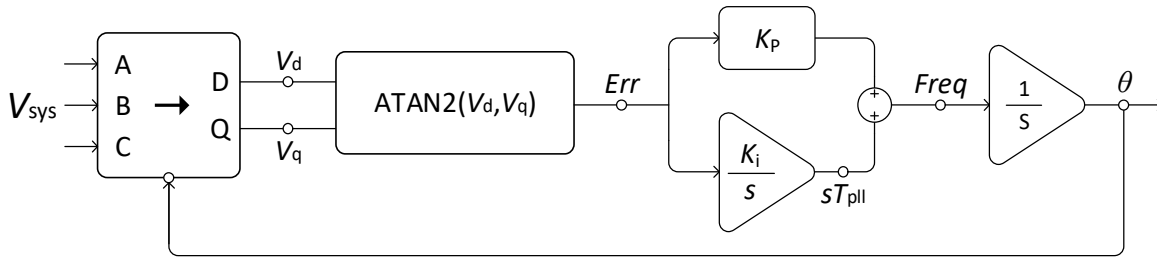


Figure 2.6: Second-order PLL

Through Park transformation and a PLL, three time-varying signals can be reduced to two stationary signals for the ease of control. These two signals are the projections onto the d and q axes of the rotating frame.

For inverter applications, both the voltage and current are projected on to the same dq frame while the PLL tracks the grid voltage's angle.

2.2.3 Decoupled Current Control Principles

The decoupled current controller was originally proposed in [9] for control of static VAR compensator (STATCOM). The methodology is built upon the foundation of Park transformation and utilizes a PLL to remain in synchronism with the grid. Figure 2.7 shows the block diagram of the common decoupled current controller.

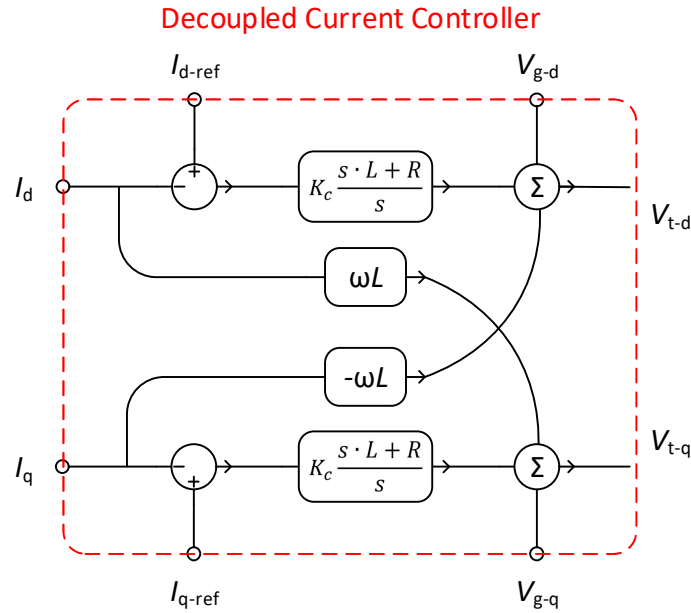


Figure 2.7: Block diagram of the decoupled current controller

The primary goal of the controller is to regulate the output current. The decoupled current controller considers the dynamics of the converter's R and L within the control loop and achieves smooth control of the output components, I_d and I_q . Figure 2.8 and Figure 2.9 further demonstrate the working principles of the controller.

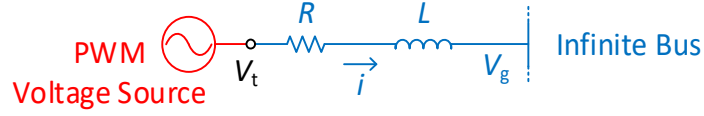


Figure 2.8: Basic circuit demonstrating the working principles of the decoupled controller

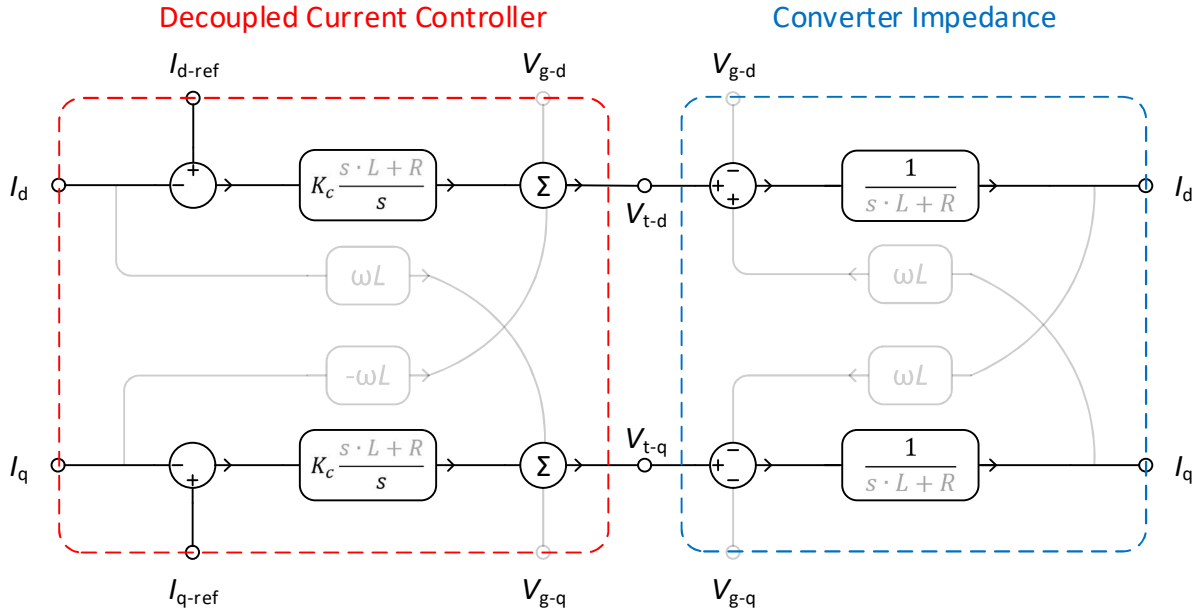


Figure 2.9: Decoupled current control cascaded with the basic circuit

Consider a simple converter shown in Figure 2.8. The converter is connected to an infinite bus; the impedance between the converter and the infinite bus is noted as R and L and the converter generates the voltage V_t . Figure 2.9 shows the complete block diagram with the decouple current controller cascaded with the converter impedance. Assuming there is no delay between inputs and outputs, the feedforward paths of the controller cancel out the coupling paths of the converter impedance. Additionally, the zeros of the PI controllers cancel out the poles of the converter impedance and leave the transfer function as shown in the following equation (2.2). In another words, the output current of the converter is simply a first-order filter of the input reference values of I_d and I_q . The speed of the controller is adjusted by K_c .

$$\frac{i_{dq}}{i_{dq-ref}} = \frac{1}{1 + s \frac{1}{K_c}} \quad (2.2)$$

Theoretically, the originally proposed decoupled controller offers great control ability for controlling the current. However, the inherent delay from the input filtering and the output transducer can make the actual transfer function complicated. In addition, the original design also assumes a strong grid. However, under weak grid conditions the controller's performance can deteriorate precipitously. These are further discussed in the next section through small-signal analysis.

The original design was still a state of art with its simplicity and robustness, but further developments have been made in [24] and [25] with additional improvements in controller's stability and decoupling capability. However, the core control loop remains intact with a direct control path to minimize steady state error and a decoupling path to suppress the coupling effect between the d and q path. This thesis will develop based on the original proposed topology but is compatible for other variances in recent development with greater control stability.

Furthermore, the decoupled current controller discussed thus far is only focused on controlling the positive sequence quantities. This makes the originally proposed topology a positive sequence controller. In recent years, there is a greater demand for the IBRs to control the negative-sequence quantities for better unbalanced fault ride through performance and unbalanced voltage management. Therefore, the tracking of negative-sequence voltage component as well as controlling the negative-sequence current is required. The scope of this thesis is focused on the positive-sequence control. Some test results presented in this thesis consider the negative sequence control loops as a proof of concept, but the extensive study of the negative-sequence control will be part of the future work. The negative-sequence control methodology considered in this thesis can be found in [39][40].

2.3 Test Models and Assumptions

A typical battery energy storage system (BESS) is developed in supporting the time-domain and small-signal analyses. A two-level converter topology is selected for the inverter, and physical parameters for the inverter are based on a combination of industry experience and literature. Some of the literature considered during the modeling process may be found in [26][27][28][29][30][31]. The full list of the inverter's physical parameters is listed in Table 2.1. From industry experience, most of the modeling parameters are similar for various applications such as wind, solar or BESS inverters and the minor discrepancies poses insignificant impact to the dynamic performance of the inverters.

A list of detailed rationality for the modeling parameters is shown below:

- **Converter impedance:** The converter impedance can have a wide range from 0.05 pu to 0.15 pu based on industry experience. A higher impedance of 0.15 pu is selected as a worst-case scenario as the converter side would have a lower fault level.
- **Converter filter:** The converter filter capacitance can vary, and the topology can also vary (such as a LC filter, LCL filter, notch filters for targeted harmonics). A simple LC filter is selected in this thesis with the filter capacitor sized to supply less than 5% of the converter MVA rating. Sensitive check will be needed to study the impact of various converter design parameters to the control performance. This will be part of the future work.
- **Converter transformer:** The converter transformer MVA size is typically similar to the size of the converter, the impedance is generally less than 0.1 pu based on industrial experience. 0.1 pu is selected here as a worst-case scenario that further limits the SCR seen by the converter.
- **DC side modeling:** The battery is directly coupled with the DC capacitor of the converter and the battery dynamics are approximated as a voltage source behind a 0.05-ohm resistor and 0.01mH inductor based on industry experience. The resistor and inductor represent the voltage drop of the battery during steady state operation and the dynamics over fast voltage transient respectively. As

the chemical reaction does not change instantaneously within the battery over sudden voltage transient and such dynamic is approximated with the dynamic of an inductor. The actual steady state and dynamic behavior of the battery is much more complex. For a directly coupled battery, a large inverter DC capacitor and a minimum measurement delay in DC link voltage can often shield the DC dynamics from the AC side especially for disturbances that do not significantly change the DC voltage such as small-signal events. Therefore, during all small-signal analyses in this thesis and time-domain simulations, the DC-side dynamics is neglected. For converters with a more sophisticated DC-side topology, such as a buck-boost converter battery, the DC-side dynamics may affect the overall control dynamics of an inverter [14]. Such DC-side converter topology is outside the investigation scope of this thesis and will be part of the future work.

Table 2.1: Plant and modeling parameters for the test system

Inverter	Value	Unit
Base Rating	1	MVA
Base Frequency	50	Hz
Terminal Voltage	0.6	kV
Converter Impedance	0.15	pu
Impedance Ratio	10	ratio
Converter Filter	300	μF
Transformer MVA	1	MVA
Transformer Voltage	0.6/0.6	kV
Transformer X	0.1	pu
Transformer R	0.01	Pu
Switching Frequency	3000	Hz
Battery Model R	0.05	Ohm
Battery Model L	0.01	mH

The limitations in real world implementation can often deteriorate the theoretical control performance. The critical limitations listed below can also affect the stability and performance of the decoupled current controller.

- Delays from the digital controllers, PWM operation, measurement devices.
- Filtering delay of voltage and current measurements

The time delay within a closed loop system can reduce the phase margin, lower the damping ratio, and adversely affect the stability of a control system [25][32]. Additional filtering can also reduce the phase margin and hinder the decoupling effect of the controller [24]. To represent these delays, a first-order filter is used as a reasonable approximation [24]. The control filter from section 2.1 has the following transfer function in (2.3).

$$G(s) = \frac{0.0144s^2 + 0.2715s + 5.117 \times 10^6}{3.493 \times 10^{-6}s^3 + 0.07716s^2 + 844.2s + 5.117 \times 10^6} \quad (2.3)$$

This transfer function is approximated at the operating frequency of the converter around 50 Hz with a phase lag of around 3° (~ 0.165 ms). Adding additional 20% in delay time, 0.2 ms delay is then used as a conservative estimate considering the overall delay from the filtering and various other types of delay.

Additionally, the bandwidth of the controller for decoupled current control is generally much lower than the switching frequency of the converter. Therefore, the switching dynamics of the controller do not significantly impact the inverter control behavior studied in this thesis. Therefore, an average-value model is used during time-domain simulation as a good approximation.

2.4 Small-Signal Analysis

This section discusses the dynamic performance and stability of a typical inverter operation under weak grid conditions. To investigate the stability of this system, a small-signal model is developed and studied. The circuit in Figure 2.10 shows a single-machine-infinite-bus (SMIB) model typically used for inverter performance evaluation. The balance of plant information is shown in Table 2.1 and the initial parameters of the controller are shown in Table 2.2. The block diagram for the PLL can be found in Figure 2.6. Figure 2.11 shows the block diagram of the decoupled current controller with its proportional gain and integral gain. The time constant of 0.2ms is the first order lag filter for the input voltage.

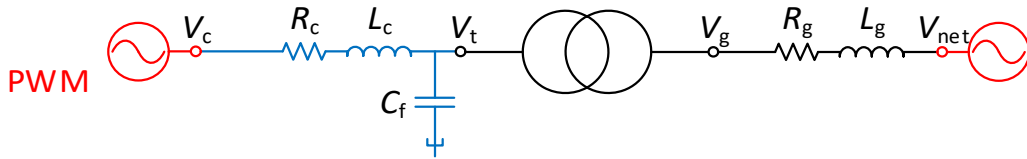


Figure 2.10: Single-machine-infinite-bus (SMIB) model of the IBR

Table 2.2: Initial parameters of the controller

Control Parameters	Quantities	Unit
PLL K_i	50	pu
PLL K_p	20	pu
Decoupled control K_i	2	pu
Decoupled control K_p	0.5	pu
Filter Time Constant	0.2	ms

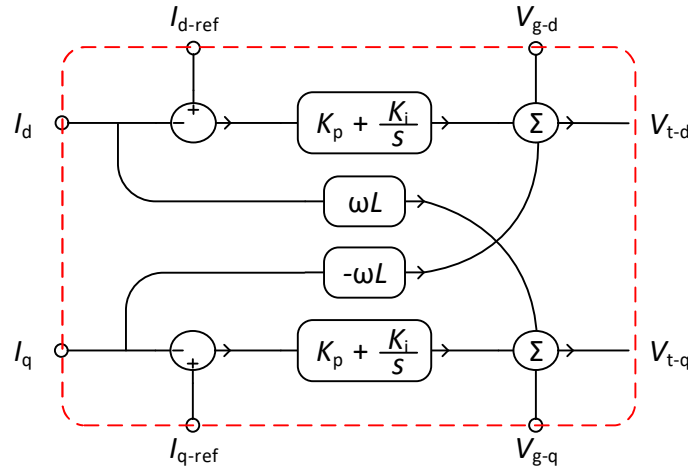


Figure 2.11: Decoupled current controller that shows the proportional and integral gain

2.4.1 Small-Signal Modeling Technique

A state-space representation of the system is developed through linearization of the system based on [33]. Since the AC system is oscillatory in nature and the oscillation frequency is around the fundamental frequency, a dynamic phasor modeling technique is used to develop a dp-frame LTI model [34]. As the number of elements increases in the AC system, a single state-space representation becomes complex. To simplify the modeling process, the AC system's small-signal model is broken down into smaller subsystems while sharing a common reference frame [35]. Each subsystem can have its linearized state space representation as shown in equations below (2.4) and (2.5) with $\mathbf{u}$ and $\mathbf{y}$ being the input and output of the system and $\mathbf{x}$ and $\mathbf{A}$ being the state variables and state matrix of the subsystem [33]. $\mathbf{B}$ is the input matrix, $\mathbf{C}$ is the output matrix, and $\mathbf{D}$ is the feedforward matrix.

$$\Delta \dot{\mathbf{x}} = \mathbf{A} \Delta \mathbf{x} + \mathbf{B} \Delta \mathbf{u} \quad (2.4)$$

$$\Delta \mathbf{y} = \mathbf{C} \Delta \mathbf{x} + \mathbf{D} \Delta \mathbf{u} \quad (2.5)$$

Furthermore, each subsystem can be interconnected to derive an overall expression for the state space representation of the entire system as shown in (2.6). $\mathbf{A}'$ in the equation are the state matrix of the entire system and can be calculated in (2.7). $\mathbf{L}$ is the connection matrix that relates the output of the subsystems to its input and represents the interactions between subsystems. This is method is commonly known as Component Connection Method (CCM) [34]. The input, output, and feedforward matrix can also be found in [36].

$$\Delta \dot{\mathbf{x}} = \mathbf{A}' \Delta \mathbf{x} + \mathbf{B}' \Delta \mathbf{u} \quad (2.6)$$

$$\mathbf{A}' = \mathbf{A} + \mathbf{B} \mathbf{L} (\mathbf{I} - \mathbf{D} \mathbf{L})^{-1} \mathbf{C} \quad (2.7)$$

The eigenvalues and eigenvectors can be further derived from $\mathbf{A}'$ for stability analysis. By using eigenvalues and eigenvectors, the small-signal analysis captures the system's dynamic behavior and provides a key perspective on the modes, participations, and contributions of its state space variables.

2.4.2 Small-Signal Model Validation

The small-signal model of the grid-following converter discussed in the previous section is shown in appendix. The small-signal model is validated by applying a small disturbance to both the original time-domain switching model, the average model, and the linearized small-signal model. Since the small-signal model approximates the full model at the chosen steady state operating point, the small-signal model's response from both the full model and the linearized model should be nearly identical. A small disturbance of 0.1 pu in the I_q reference is applied to all models. The test system's SCR is set to a low SCR of 1.5. Figure 2.12 depicts the responses from all models. The three models show reasonable alignment. An eigenvalue analysis is performed and discussed in the next section.

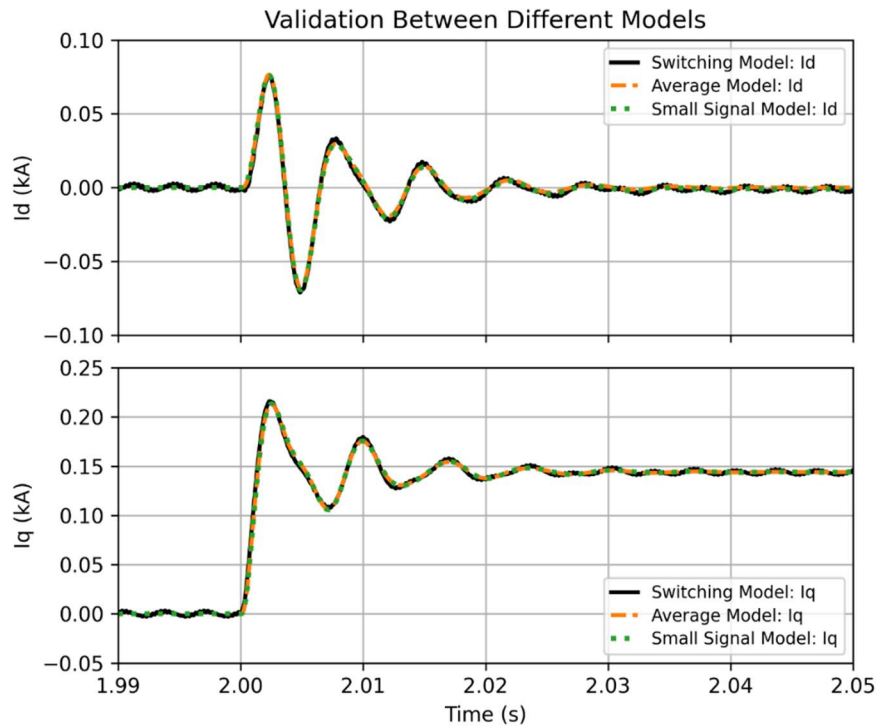


Figure 2.12: Responses from all three models to a small disturbance

2.4.3 Eigenvalue Analysis

The model validated in the previous section is used for eigenvalue analysis. A visualization tools was developed to better understand the stability of a given system. An eigenvalue diagram of the system shown in Figure 2.13. This simple diagram is helpful in identifying the stability of the system at the specific operating condition. However, the diagram itself has the limitations of identifying the stability margin or damping of the system.

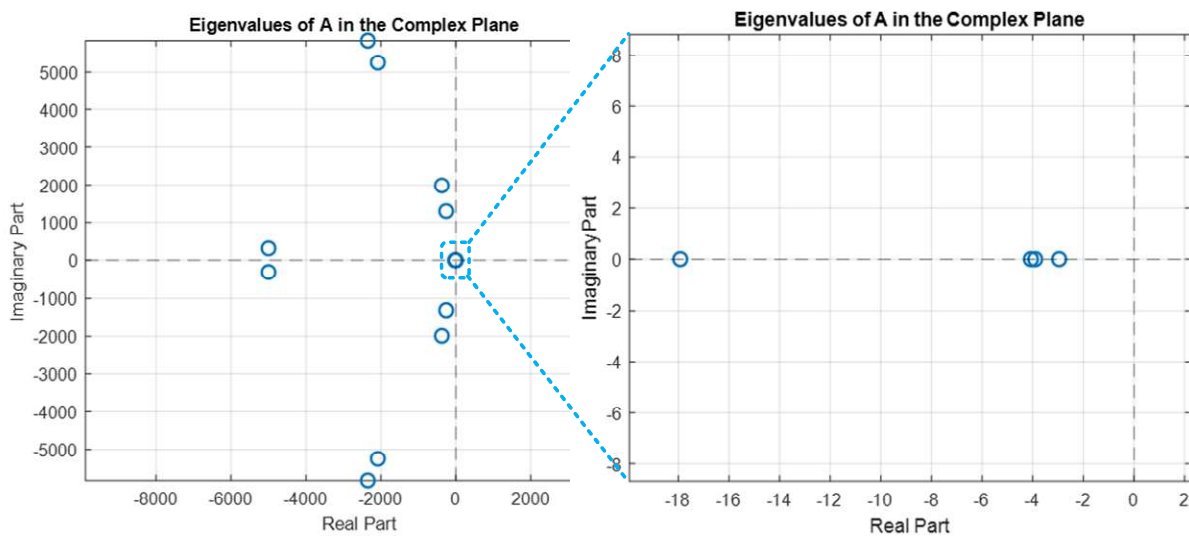


Figure 2.13: Eigenvalue diagram of the system (left diagram zoomed out, right diagram zoomed in)

In control systems engineering, study of the dynamics of a system can be done using two primary mathematical models: the transfer function (frequency-domain) and the state-space representation (time-domain). While they look different, they describe the same system, and their most important characteristics are identical.

The poles of a transfer function are the roots of its denominator, while the eigenvalues of a state-space model are the roots of its characteristic equation. The fundamental connection is that the poles of the transfer function are precisely the eigenvalues of the state-space representation. This means that analyzing the

system's stability and transient response can be done interchangeably by examining either the poles or the eigenvalues.

It is well known that for a system to be stable all the poles of the system need to be on the right-hand side of the real-imaginary plane. However, the positions of the poles also reveal the damping of a given system. This is calculated through damping ratio (ζ) based on how quickly the system damps out oscillations in respect to the oscillation frequency. Given an eigenvalue, λ , the damping ratio can be calculated using the following equation (2.8).

$$\zeta = -\frac{\text{real}(\lambda)}{|\lambda|} \quad (2.8)$$

Damping ratio is a great predictor of phase margin of a system and it is widely used by engineers to ensure a well-damped stable system [37][38]. An automated process was developed to filter all the poles from an operating state to find the minimum damping ratio. In a 3-D diagram as shown in Figure 2.14, the system's damping is presented against various system SCR and I_d injection values. The unstable region (negative damping) is marked with red. The stable region is marked with the green, and the height of the sphere shows the system damping.

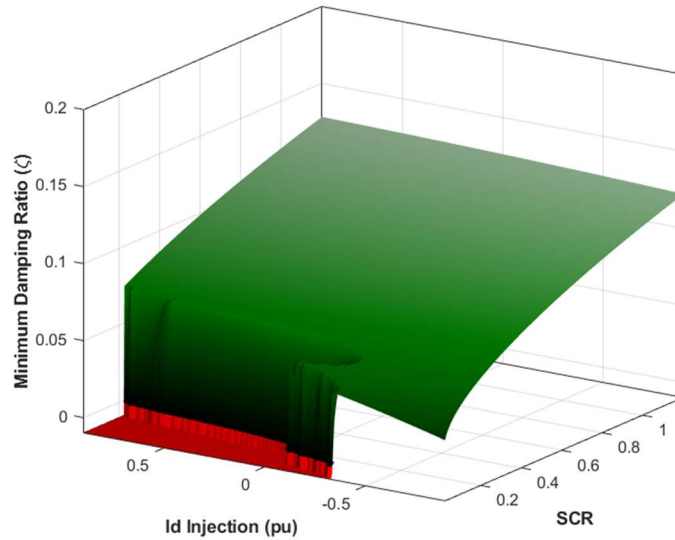


Figure 2.14: 3D view of the system damping between various system SCR and I_d injection values

Figure 2.14 can be further presented from top or side views as shown in Figure 2.15 to better illustrate the system's stability at low SCR. The axis in these figures will not be marked to simplify the presentation. All graphs will have their x-axis as SCR. The y-axis from the top view will be the injection in pu (-0.9 to 0.9 pu) and from the side view the minimum damping ratio.

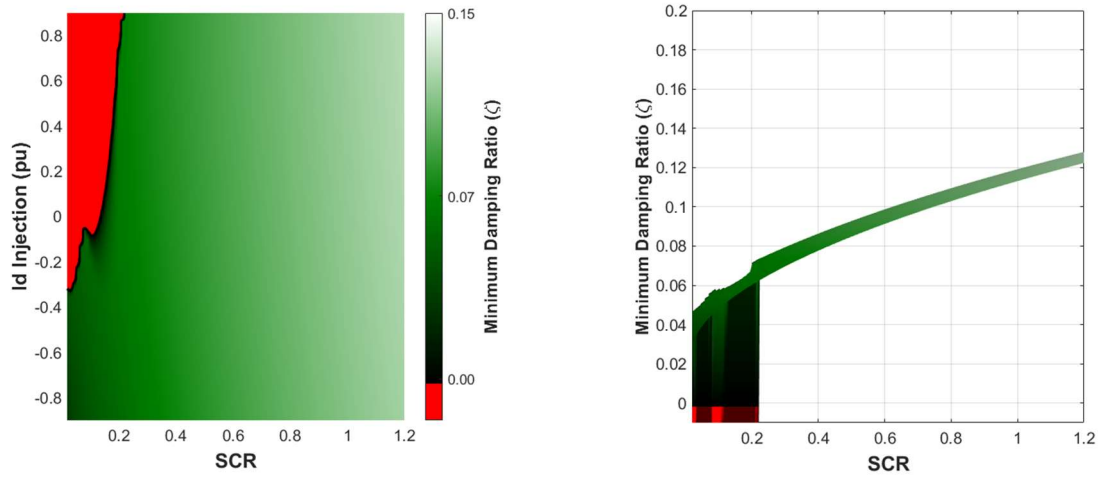


Figure 2.15: Top view and side view of the system damping

Grid X/R Ratio

The effect of grid X/R ratio is shown in Figure 2.16. For various X/R ratios tested, the decoupled current controller has difficulty operating under ultra low SCR conditions. This is when SCR drops below 0.1 - 0.2 under the initial zero power output and it gets worse with additional positive active power injection.

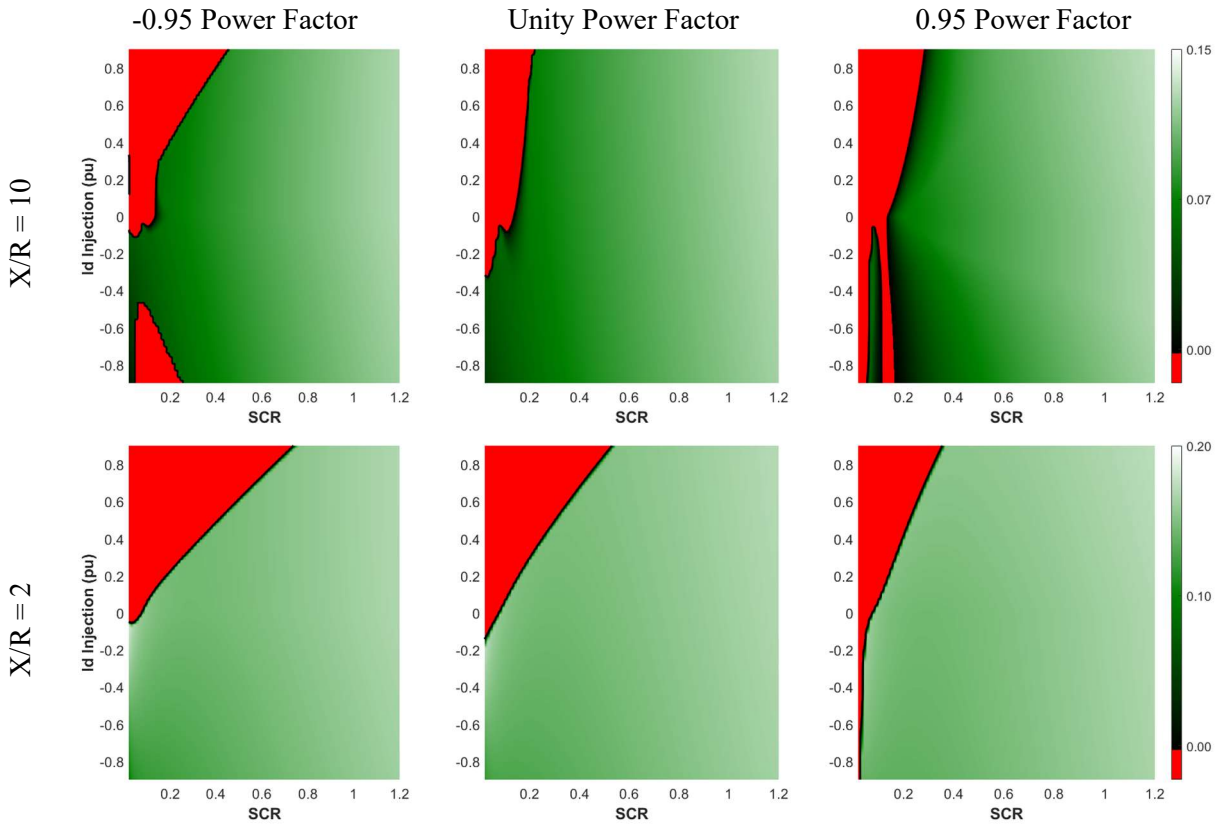


Figure 2.16: Sensitivity to grid X/R ratio

Control Parameters

The sensitivities of the control parameters are investigated at X/R ratio of 10 and unity power factor. The first variable to check is the filter time constant, T_f . The focus of this investigation is at ultralow SCR values. One of the foundations of decoupled controller is that the actual current control loop is added on top of the measured grid voltage. Since the grid voltage is filtered and all inverters have a small amount of control delay, it is expected a deterioration of stability with higher filter time constant. This is demonstrated in Figure 2.17 as the area of instability increased significantly as the filter time constant increased.

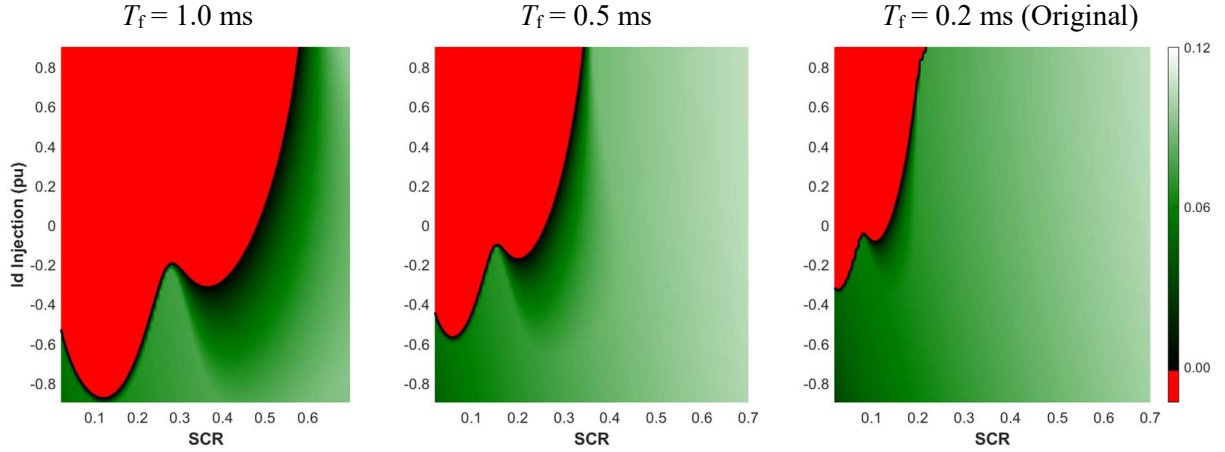


Figure 2.17: Sensitivity to input filter time constant

The PLL control parameters sensitivities are checked in Figure 2.18 and Figure 2.19. It is as expected that, as the SCR increases, the dynamics of PLL becomes less impactful as the locus of damping converges around 0.16. Additionally, lower PLL gains do provide better stability at ultra low SCR at the cost of slower tracking speed. The weak system is prone to phase angle changes and thus require a fast PLL, but a fast PLL can cause instability. Therefore, it is challenging to set the PLL parameters especially for the weak system. One additional insight can be revealed here. The decoupled current controller does not necessarily need the PLL to operate in terms of current limiting. Even when the PLL dynamics is slowed or, in an extremely case, locked in a fixed frequency, the decoupled current control will continue to limit its maximum current. However, the injected I_d and I_q will not align with the actual grid voltage in this extreme scenario, since the PLL is not tracking the grid frequency. The I_d and I_q will lie in the direction of the V_d and V_q defined by the PLL angle.

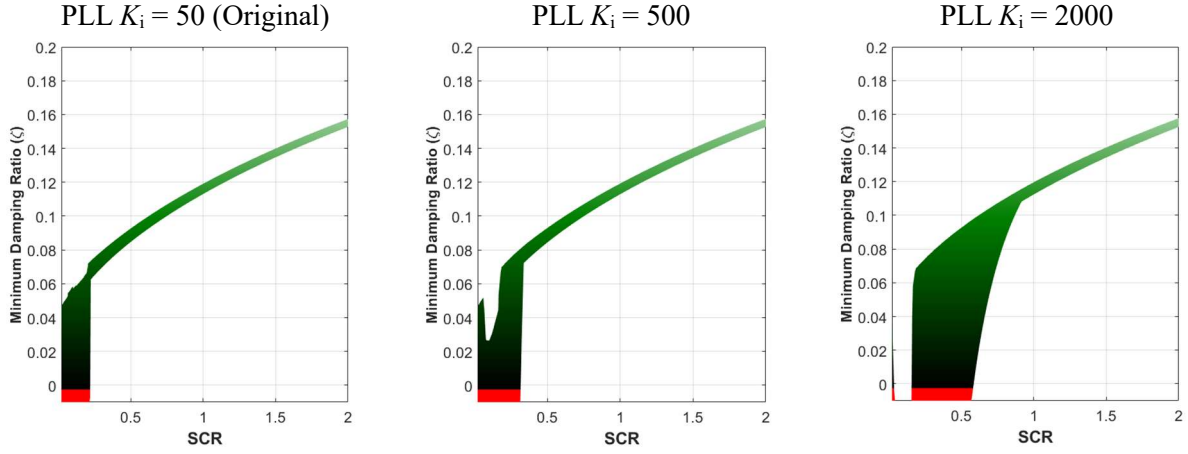


Figure 2.18: PLL integral gain control parameters sensitivities

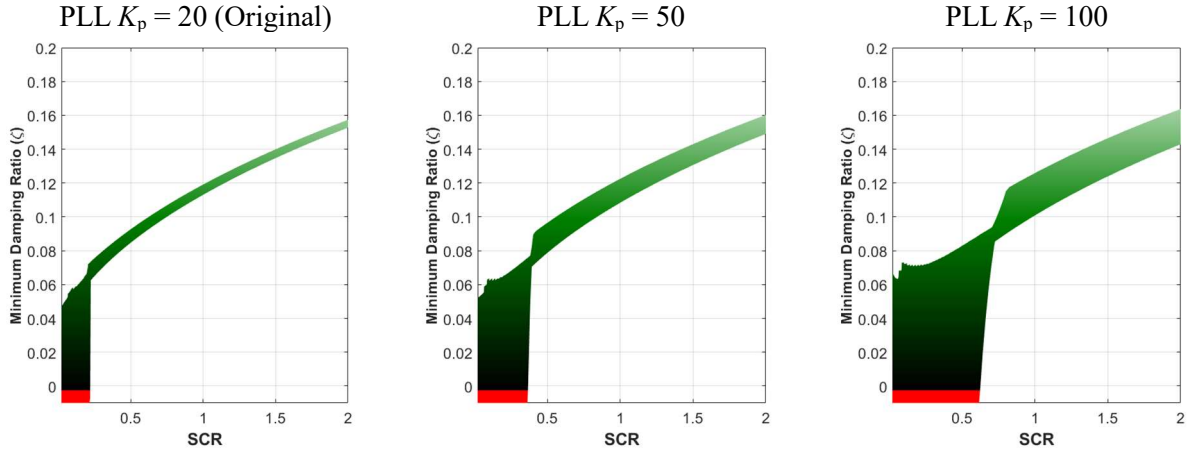


Figure 2.19: PLL proportional gain control parameters sensitivities

Figure 2.20 and Figure 2.21 depict the stability of the controller with various K_i and K_p values. Increasing both the K_i and K_p gain can reduce the overall damping ratio. For K_p specifically, reducing the gain can significantly improve the gain at higher SCR but has deteriorating effect at lower SCR.

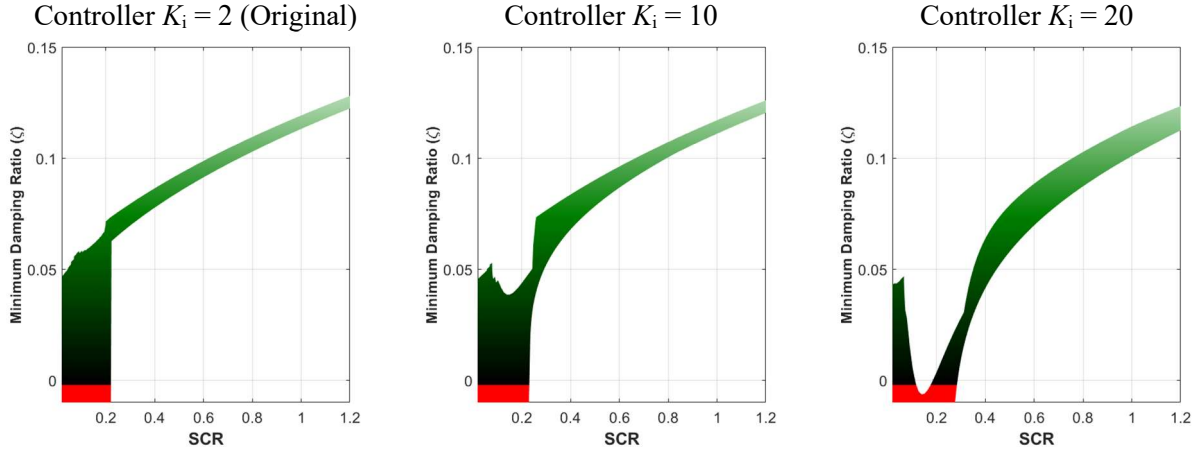


Figure 2.20: Decoupled current controller integral gain control parameters sensitives

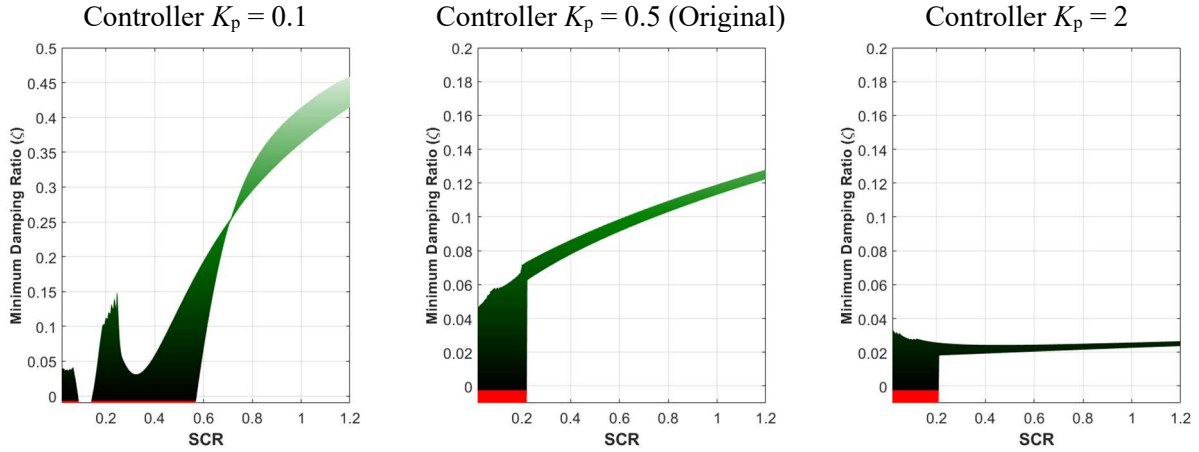
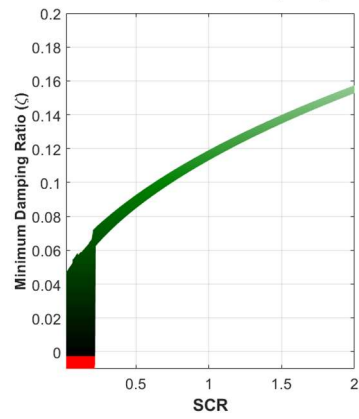


Figure 2.21: Decoupled current controller proportional gain control parameters sensitives

The effect of feedforward loop was also tested in Figure 2.22. The feedforward loop can improve the stability of the controlled system. This shows one of the main advantages of the decoupled current controller that using the feedforward loop to decouple the I_d and I_q current.

With feedforward decoupling loop



No feedforward decoupling loop

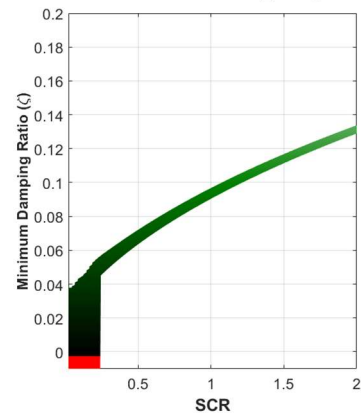


Figure 2.22: Effect of the feedforward loop in decoupled current controller

2.5 Other Critical Functions from a GFL Inverter

This section discusses a few other critical features typically implemented within a GFL to improve its performance and stability during grid events.

- FRT Blockings
- Dynamic reactive current support
- Negative-sequence current support

FRT Hold and Blockings

Since GFL relies on a stable grid voltage and angle, it can become difficult to remain stable under fault condition especially during three-phase deep faults (typically less than 0.3 pu residual voltage at the inverter terminal). One simple strategy typically employed by a GFL inverter is to hold all the control loops that are prone to saturation during a fault. This includes the reference points for I_d , I_q , and PLL frequency. Furthermore, in some extreme cases, there can be significant distortions in inverter terminal voltages and angle upon fault recovery. The GFL inverter control system may lose track of the grid under such events and causes a catastrophic control failure. To prevent such cases, the PWM operation can be blocked for a cycle or two following the fault to allow the system transient to damped out. During this time, the PLL can be released from holding position to relock to the grid. When the plant is operating under GFM mode, some of the blocking strategy can still be effective such as blocking the PLL during a deep fault.

Dynamic reactive current support

In many new standards and grid codes for grid-connected inverters, it is required for the inverter to support the grid voltage by injecting reactive currents during fault ride-through. The system impedance at the fundamental frequency from the point of view of the inverter is mostly inductive especially during faults. Therefore, supplying capacitive reactive current during faults can boost the overall system voltage and support the system recovery. For GFM controls, the natural dynamics of the GFM inner loop can have a

very similar effect of reactive current support and under such case, the additional reactive current inject is not required.

Negative-sequence current support

During recent years, grid codes have increasingly demanded inverters to have negative-sequence current control provisions. The original proposed decoupled controller is only developed for positive-sequence control and has been effective and widely adopted. However, to further optimize the plant performance during unbalanced faults, the decoupled controller, as well as the PLL, is modified for negative-sequence control. From the PLL's level, the positive- and negative-sequence components are decoupled from each other, and the positive-sequence voltage angle was tracked by the PLL [39]. With the measures separated into positive-sequence and negative-sequence components, an additional decoupled controlling loop can be used to manage the negative-sequence current [40]. This thesis only performed limited tests with the addition of negative-sequence controller as a proof of concept. The full impact assessment will be part of the future work.

Chapter 3

Limitations of Existing GFL Methods and Introduction of the Proposed GFM

This chapter investigates the limitations of existing GFL control methods, especially when operating under islanded conditions. A thought experiment of the GFL behavior under islanded operation is carried out and a new GFM control is proposed.

3.1 Introduction

The GFL inverter controls its current injection based on the I_d and I_q commands, as discussed in detail in section 2.2.3. These commands are generally converted from the active power and reactive power commands from the power plant controller's (PPC's) external control layer. Although this command may respond to the grid voltage and frequency fluctuations, the response time is generally long (in seconds). For an IBR to form the grid, the inverter must respond to the grid voltage and frequency fluctuation as fast as possible, typically within milliseconds. The current control loop also requires immediate response from the controller, and this can often cause a control conflict. Control saturation or wind up can occur when the controller tries to maintain grid power balance while limiting the current during fault conditions [10][11].

To develop a GFM control method based on the well-studied decoupled current control topology, the first step is to study the limitation of the GFL control under islanded conditions. As mentioned earlier, the I_d and I_q commands for the GFL can often be different from actual grid requirement during an islanded condition.

A thought experiment is presented in this chapter to investigate the behavior of the GFL when the I_d and I_q reference do not match the active and reactive power requirements of the grid.

3.2 Thought Experiment

Imagine a condition where an inverter tries to inject I_d current to a resistive load in an island. If the impedance magnitude (in per unit) of the load matches the current magnitude (in per unit) at rated voltage, an islanded operation will result in a stable operation in an ideal scenario.

However, if the load is a shunt reactor and the inverter is still trying to inject I_d into it, there will be a control failure. The inverter may excite an initial voltage when it attempts to inject I_d , and this I_d is based on the direction of V_d based on the internally tracked PLL angle. The reactor receives this I_d and quickly shifts its voltage angle to be 90° apart from the injected current as the natural behavior of the reactor dictates. The internal V_d of the PLL is then 90° apart from the actual system voltage. The GFL PLL will attempt to realign with the new grid angle and continue to inject I_d based on the new angle. However, the shunt reactor cannot absorb active power and will continue to repel this alignment action from the inverter's PLL and tries to be 90° apart from it. Then a chase in angle will begin and causes the angle to be ever advancing, thus increasing the overall system frequency.

Similarly, if the inverter remains the same but the load is changed to a capacitor the chase in angle will be in the opposite direction. The frequency, in this case, will decrease. This phenomenon is further tested using time-domain simulation experiments in section 3.3.

Note that during a deep fault, the inverter would lock the PLL angle to prevent wind up. This typically happens when the inverter terminal voltage drops below $0.2 - 0.3$ pu. This thought experiment assumes that there is no lock mechanism for the PLL angle as the inverter during islanded condition can have high terminal voltage depending on the load within the islanded system.

3.3 Experiment

To test the inverter behavior when the load's nature does not match the supply, the test set up in Figure 3.1 is considered. An average-value inverter model (in red) with decoupled current control is connected to an external RL load. The test procedures are as follows:

- The right-hand side of the RL branch (in blue) is connected to an infinite bus at the beginning of the test.
- The inverter injects I_d and I_q to the infinite bus through the RL branch. The magnitude of the injected current is fixed whereas the angle varies between tests.
- Once the simulation reaches steady state, a fault is applied to the right-hand side of the RL branch and forces the inverter to inject I_d and I_q to the RL branch.
- The frequency of the system is recorded and displayed.

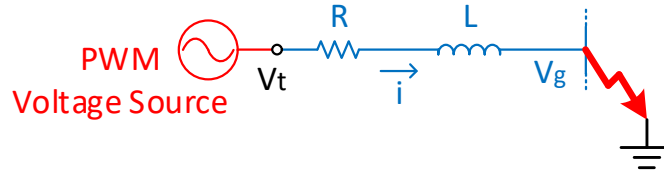


Figure 3.1: SLD of the test set-up.

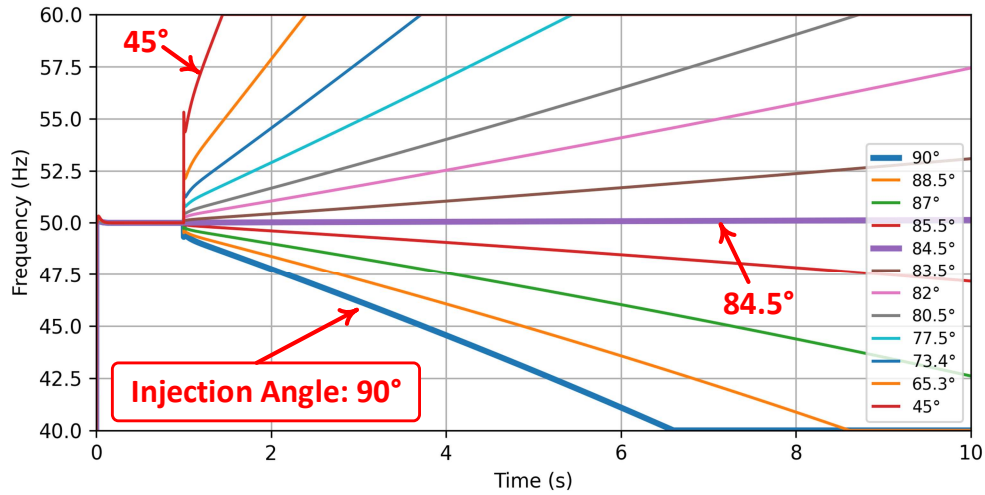


Figure 3.2: Test results of the inverter trying to inject current with various angles to a fixed RL branch.

Figure 3.2 depicts the test results of the inverter under different injection current angles. The X/R ratio of the RL branch is around 10, which corresponds to an angle of around 84.5° (load angle). The observation from the test results can be summarized as follows:

- When the reference angle is greater than the load angle, the PLL slows down.
- When the reference angle is less than the angle of the RL branch, the PLL speeds up.
- The larger the angle difference, the faster the frequency deviation.

Additional insight:

During a solid fault close to the POI of the GFL, the fault decouples the inverter from the grid and forces the inverter into an islanded state. The impedance between the fault and the GFL is mostly from the transformers, which are primarily inductive. If the GFL tries to feed the grid with active power, having a current reference comprising primarily I_d can severely deteriorate the stability of the angle and frequency of the PLL. This is because injecting I_d towards a mostly inductive impedance would escalate the PLL's instability. This is demonstrated in Figure 3.2; the frequency drifts the fastest when the current reference has a higher I_d component (i.e., at 45°). Therefore, it is generally good practice in the industry to prioritize I_q during a fault and remove I_d if necessary.

Moreover, prioritizing I_q can reduce the frequency drift but not minimize it. This is again demonstrated in Figure 3.2 where a drift in frequency is still observed when the injection angle is 90° . This is because the impedance between the inverter and the fault has losses. These losses cannot be compensated by I_q and can cause the drift in frequency. Additional optimization can be applied to maintain a small amount of I_d

reference during high I_q injection. The small I_d injection should match the typical X/R ratio of the faulted system which is typically around 10.

The insights derived from this experiment conform with the common practices in the industry to prioritize I_q during fault but maintain a small amount of I_d (5-15%). With these approaches, the rate of instability of a GFL can be minimized and help fault ride-through.

3.4 Theory development

It has been noted from the experiment in section 3.3 that the mismatch in the reference angle and the load angle can destabilize the PLL during islanded operation. This section further investigates the process during the destabilization period and formulates the theory around this phenomenon.

When the inverter is islanded with a local load (assuming an RLC load), there is no synchronization with an external grid. The frequency and angle that were originally driven by the grid are now self-excited by the inverter and the local load. Since the PLL is acting as a frequency filter of the grid, its internal integral loop maintains the inverter's operating frequency and its state does not change instantaneously. This “inertia-” liked behaviour of the PLL allows the system to continue to be excited by the inverter close to the system frequency momentarily after islanding.

Although the frequency is driven by the PLL's “inertia” during islanded operation, the actual terminal phase angle is dictated by the load characteristics. The PLL, after the islanding condition, continues to establish a rotating frame for the decoupled controller to assert control, but the load itself also needs to maintain an angle difference between the voltage and the current depending on the angle of the load's impedance.

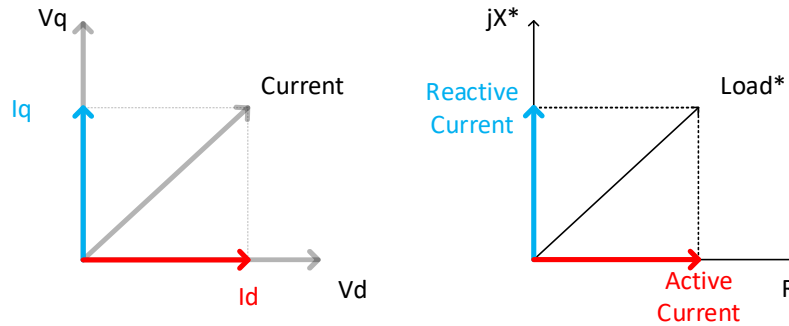


Figure 3.3: I_d and I_q injection to a load with a matching angle and current magnitude

If the injected I_d and I_q coincidentally match the load angle as shown in Figure 3.3, the load angle and the current angle will cancel each other out and result in the terminal voltage aligning perfectly with the d axis of the rotating frame. The PLL will continue to operate as normal, even under islanded conditions, if the

current angle matches the load angle. However, more commonly, if the load angle does not match the current angle, the load will react to the injected current and cause an angle deviation/jump in the terminal voltage. The magnitude of the angle jump is the difference between the load angle and the current reference angle as shown in Figure 3.4.

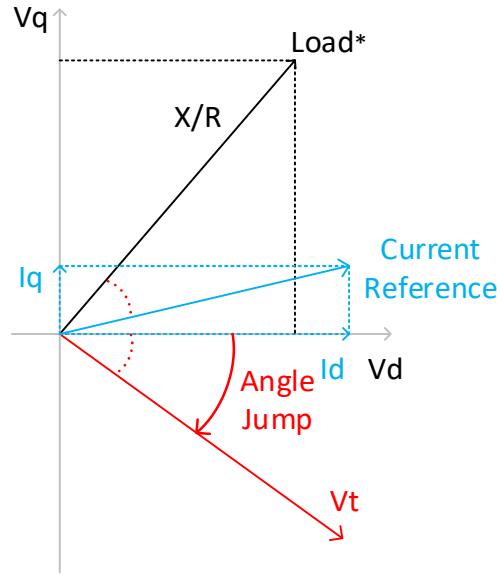


Figure 3.4: I_d and I_q injection to a load with a mismatching angle

Due to the relatively slow dynamics of the PLL compared to the RLC dynamics of the system, this rotating frame does not react quickly to the sudden change in the grid voltage's angle. The new terminal voltage would not align with the PLL's V_d . This creates an angle error at PLL input. This error to the PLL will either speed up or slow down the PLL excitation frequency. As shown in Figure 3.5, if the load angle is leading the I_d and I_q current reference angle, the PLL will observe a positive error. Similarly, the input error would be negative if the load lags the current reference angle.

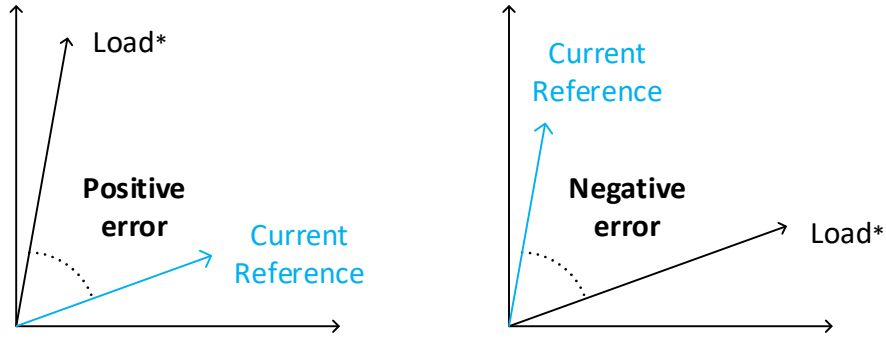


Figure 3.5: Relationship between pll input error and the relative locations of the load angle and current reference angle

Furthermore, this error will pass through the PLL's PI control loop and manifests as a step response through the proportional path and a ramp response from the integral path. The output of the PLL is the combination of the step response and the ramp response as demonstrated in Figure 3.6.

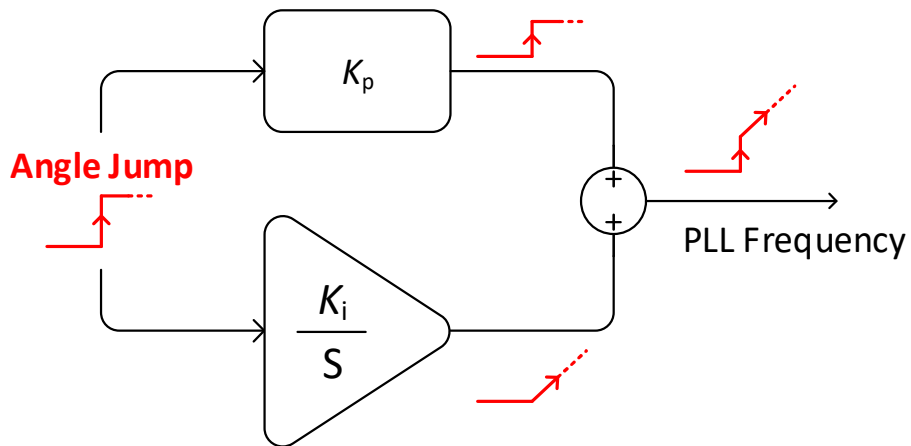


Figure 3.6: PLL inner control loop react to an angle change

The PI controller control settings will determine the magnitude of the step response and the ramp rate of the ramp response as listed below.

- A step through the proportional loop: angle difference * K_p unit in Hz
- A constant drift through the integral loop: angle difference * K_i unit in Hz/s

This is further validated with the following EMT simulation of injecting a current angle of 90° (capacitive) to a RL branch with an angle of 45° . The integral gain in this case is 30 and the proportional gain is 2. The RL branch is disconnected from the infinite bus and grounded at $t = 1$ s as shown in Figure 3.7. The left figure shows under ideal conditions, the system responds as predicted based on the theory discussed above. The simulation results show a frequency step of $(90-45)/360 \times 2 = 0.25$ Hz and a frequency ramp of $(90-45)/360 \times 30 = 3.75$ Hz/s. For the figure on the right, this is under nonideal conditions with the actual inverter switching model equipped with filters and control delays. Despite some overshoots, which is expected for an actual inverter, the PLL response generally conforms with the theoretical response.

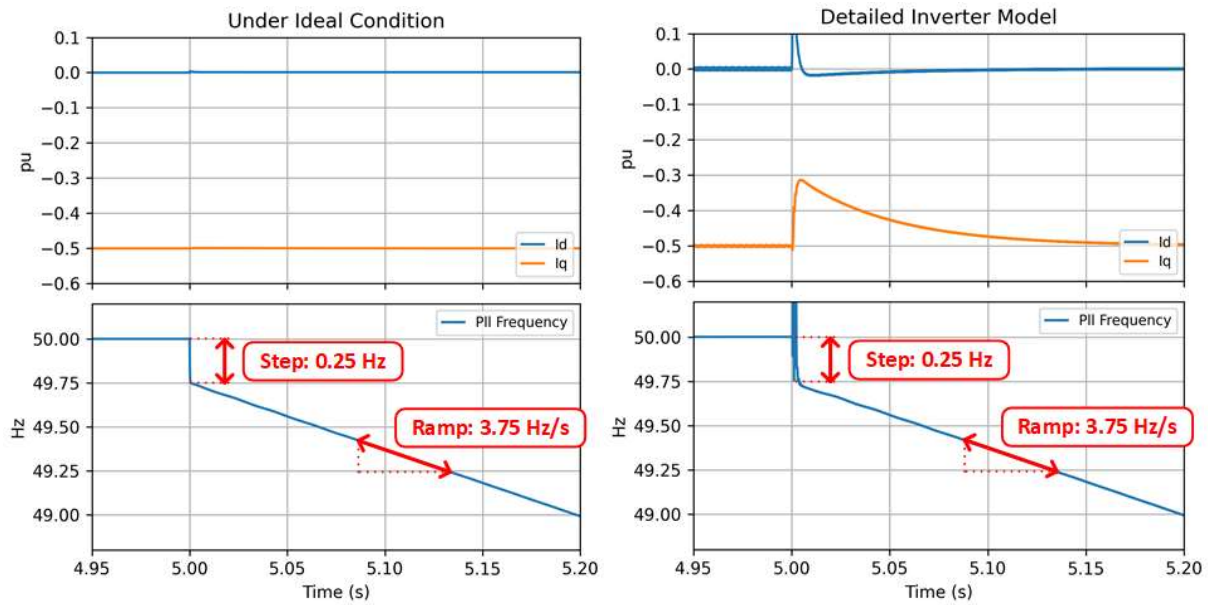


Figure 3.7: EMT simulation validation under ideal condition (left) and nonideal conditions (right).

3.5 Controller Development

In summary, for a grid-following inverter under islanded operation with a local load, the PLL turns to become destabilized if the current reference angle does not match the load angle. Furthermore, depending on the relative angle difference between the current reference angle and the load angle, the PLL will either speed up or slow down the frequency. The terminal voltage of the islanded inverter is a product of the injected current magnitude and the load impedance magnitude.

To allow a grid-following inverter to operate under islanded conditions (i.e., grid-forming mode), one must ensure that:

- The current reference angle constantly matches the load angle
- The output current magnitude matches the load demand at rated voltage.

This thesis proposes a method of matching the current reference without knowing exactly the load magnitude or angle. This method is based on the measured terminal voltage and the frequency response of the PLL.

- The inverter can determine if the reference current is leading or lagging the unknown load angle based on the PLL frequency response. A PI controller can be used to adjust the current reference angle based on the tracked frequency. Once the current reference matches the unknown load angle, the frequency of the controller is stabilized and can be brought back to the frequency set point through the PI controller.
- The inverter can determine the magnitude of the current reference through the terminal voltage. If the current injection angle matches the load angle, the magnitude of the terminal voltage would be linearly proportional to the injected current. A PI controller can be realized here to track the voltage to a set point and controls the magnitude of the current injection.
- An upper limit can be set for the voltage PI controller to limit the current reference magnitude.

Figure 3.8 shows the core control loop for the regulation of voltage and frequency. This is an extremely simplistic and minimalistic design compared to some of the GFM designs proposed in the literature. This controller can be easily augmented on top of an existing GFL. It takes the voltage and frequency output from the original GFL control in creating the I_d and I_q references for the decoupled controller. Some parameter optimizations maybe required from the original GFL control, but no topology change is required. For controllers with negative sequence loops, this thesis considered only adding the proposed GFM control loop to the positive sequence controller. Further optimization of the negative sequence controller maybe required, and this will be part of the future work.

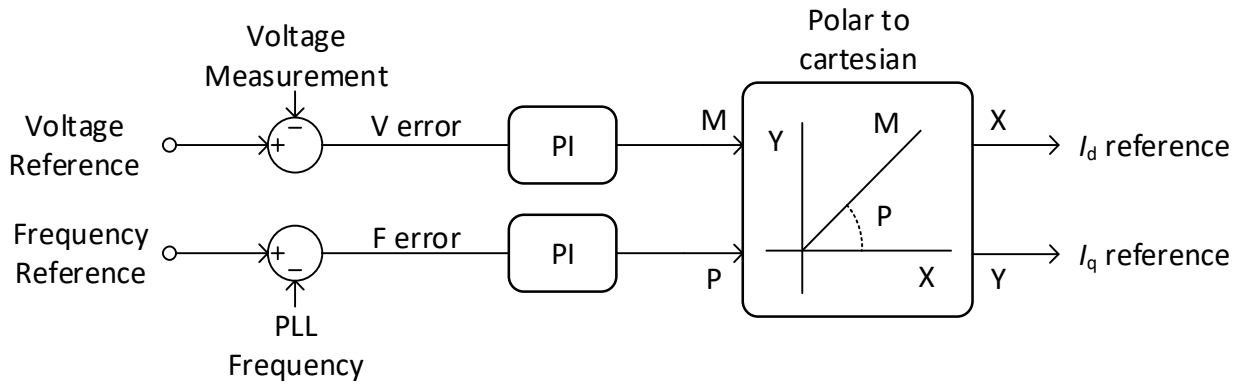


Figure 3.8: Core control loop of proposed GFM with voltage and frequency control

The above inner loop would be sufficient for islanded operation with the GFM being the only generator regulating the grid frequency. To connect to a microgrid with other generators or a utility grid, additional augmentation to the control is required to enable connection to an external PPC, which has additional droop control characteristics. The outer loop control with a PPC is generally much slower in control speed and only participates in the secondary response during a rapid transient event. The scope of this thesis will focus on the analysis of the inner loop control and the dynamics from the external secondary control is ignored. Adoption to an external PPC control is theorized here and preliminarily tested in later chapters. This thesis will not investigate and optimize it in depth, and it will be part of the future work.

- When the inverter is connected to a grid, a fixed frequency setpoint would not be possible. Instead of controlling a fixed frequency setpoint, a transient frequency error can be adopted. The input error is passed through a washout filter such that the frequency controller only provides temporary support to the grid. The support duration can be tuned through the time constant of the filter. An additional limit can be added to the filter. The purpose of the limit is to allow the controller to provide sustained frequency control if the local region surrounding the plant is islanded and is causing a much higher frequency deviation. A similar control methodology can also be applied to voltage.

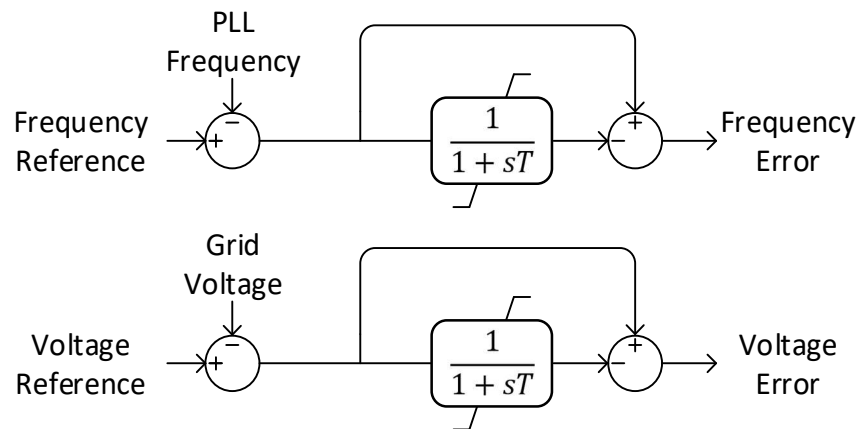


Figure 3.9: Dynamic frequency set point and dynamic voltage set point

- If the converter is connected to a grid, the converter will have additional freedom to inject arbitrary power and reactive power within its current and voltage limits. During the actual operation, an additional voltage droop and frequency droop might be requested by the external PPC control or grid operator. To achieve this additional flexibility in active and reactive power control, the PI control for the voltage and frequency in Figure 3.8 can have a dynamic reset function. This dynamic reset function can periodically reset the PI controller based on external values and move its state value towards the target value.

Chapter 4

GFM Analysis and Control Refinement

Chapter 3 discussed basic grid-forming control structures in detail. This chapter further evaluates and optimizes the proposed GFM topology. Section 4.1 investigates the proposed GFM control topology through large-signal and small-signal studies and discusses some of the limitations of the proposed topology. Section 4.2 further optimizes the proposed control topology. Section 4.3 discuss the operation of the proposed GFM control topology with conventional synchronous machine-based resources. Section 4.4 summarizes the insight from this chapter.

4.1 EMT Model Development and Testing

An EMT model of a GFM is developed based on the core GFM control loop from Figure 3.8. The proposed control loop is augmented on top of a GFL model using the decoupled current control methodology.

GFL control parameters are further tuned through trial and error to improve the overall control performance. The updates include reducing the K_p value of the current control (to 0.1) and reducing the overall tracking speed of the PLL. Reducing K_p helps to improve the control stability for the current loop at a higher SCR as suggested from investigation performed in section 2.4. However, reducing K_p does not help the controller at ultra-low SCR values and can pose some challenges to the GFM operating at low load conditions. The PLL is essentially filtering the grid angle, and this filtering effect is analogous to the effect of the rotor inertia. Reducing the PLL speed helps to provide a more stable internal angle during disturbances based on investigations from section 3.4.

A list of tests is shown below with an initial grid-connected GFM islanded with a simple RL load. The simulation schematic diagram is shown in Figure 4.1. Figure 4.2 depicts the simulation results.

- The plant of 1 MVA is initially connected to the grid with an SCR of 1 as shown in figure below with an X/R ratio of 3.
- A solid fault is applied at $t = 2$ s and at $t = 2.5$ s the fault is cleared with the GFM islanded with a local constant impedance RL load of 0.5 pu based on 1 MVA at nominal voltage.
- The RL load started as mostly reactive with X/R ratio of 3. During Islanded operation, the load angle is changed to mostly resistive (X/R ratio 0.2) at $t = 3$ s and then to equally reactive and resistive angle at $t = 3.5$ s (X/R ratio 1.0).
- The load size is increased to 1.5 pu at $t = 4$ second and a gradual load rejection test is followed with the load sizes reduced to 0.15 pu.

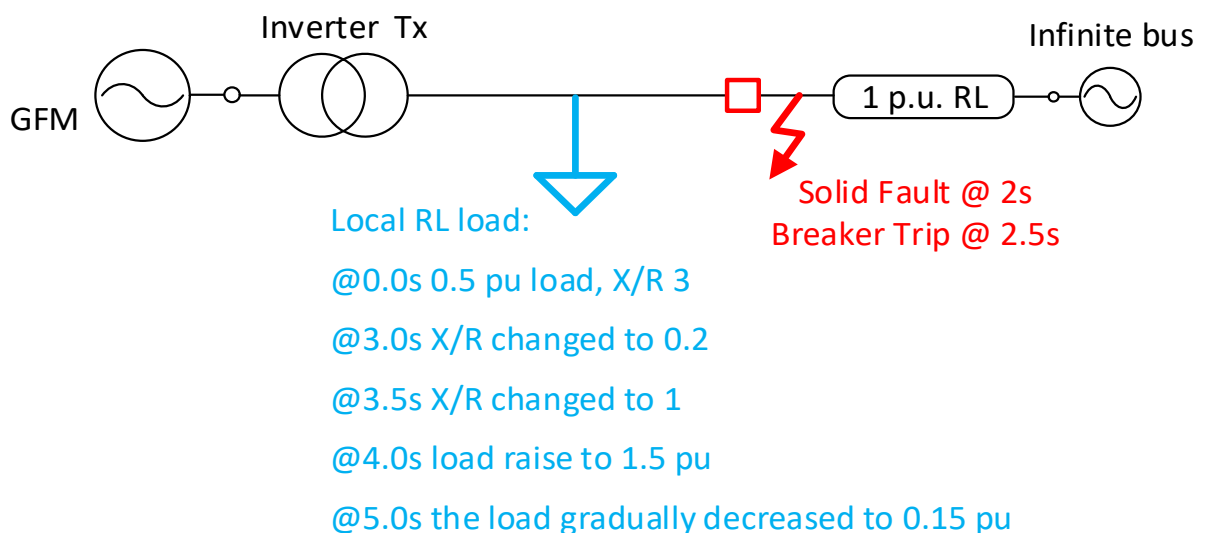


Figure 4.1: The schematic diagram of the simulation set up

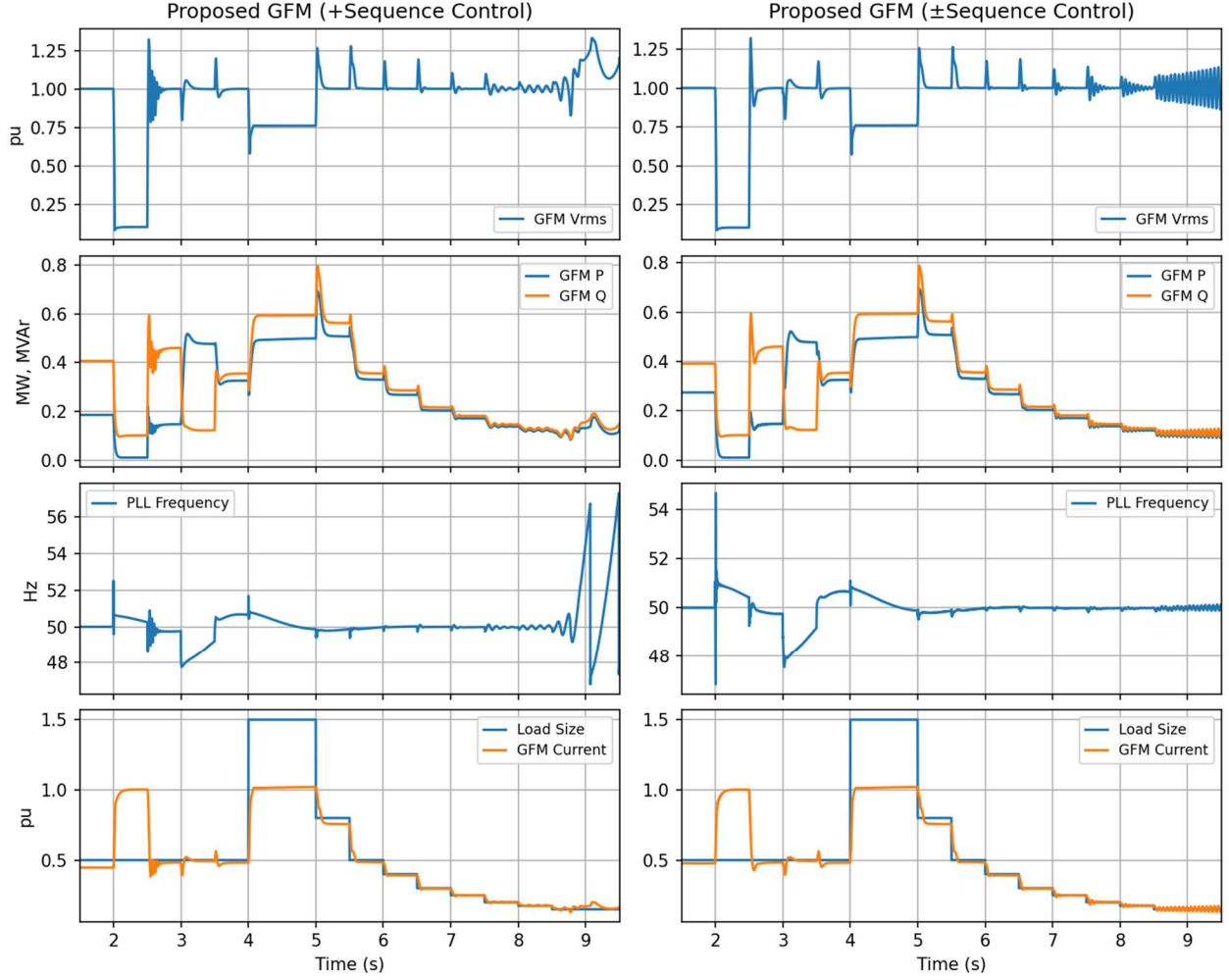


Figure 4.2: Large signal tests of the proposed GFM topology

The positive sequence decoupled current controller was tested and the results are shown on the left-hand side of Figure 4.2. The tests show stable performance during the fault, large load angle changes, and large load rejection. The augmented GFM controller can regulate the grid voltage and grid frequency under high load conditions (between $t = 1-8$ s). There is no saturation or control failure during fault ($t = 2-2.5$ s) or during the 1.5 pu large load connection ($t = 4-5$ s) and the current is closely regulated at 1 pu. The same tests are also performed with the addition of the negative-sequence control loop as shown on the right-hand side of Figure 4.2 as a proof of concept. The rest of the thesis will focus the development on the positive-sequence controller but the same should also be applicable to the control topology with the negative-

sequence loop. Further investigation and optimization with the negative-sequence control loop will be part of the further work.

Apart from the successful responses from the proposed GFM control topology, several issues were also observed as listed below:

- Small-signal stability issues are observed at low load condition.
- Additional tests were also performed and showed instabilities of the proposed GFM operating in parallel with other GFL inverters.
- Tests were also performed that indicated that the proposed GFM method cannot withstand large disturbances when operating with a synchronous machine. Pole slipping observed in most tested cases suggest that the proposed GFM has a poor performance in providing angle stability.

If the inverter is to be operated in a grid condition with system loads primarily passive, the proposed GFM topology can be a simple yet effective solution. However, the proposed GFM topology needs further improvements to operate under a more complex system load profile with GFL or rotating machines.

To further investigate the stability issue in a more systematic manner, a small signal study is carried out in section 4.2. In the same section, improvements and optimization to the original proposed GFM strategy are also discussed.

4.2 Small-Signal Study

A small-signal study is carried out in this section for the proposed GFM control topology. Since it is expected that the GFM should be operated reliably in all four quadrants as listed in Table 4.1, it must also be tested in all four quadrants. The four quadrants are split into four zones and circuits were developed for each zone to study the GFM's small-signal stability. A passive load is used for zone 1 and zone 2 when GFM is supplying power. When the GFM is absorbing power, GFL IBR models are used as generators

operating at unity power factor. The reactive power consumptions of the load are represented using reactors and capacitors. Figure 4.3 breaks down all the circuit setups designed to test the GFM in all four zones.

Table 4.1: Summary of all possible operating zones of a GFM inverter

Operating Zones	Active Power	Reactive Power
Zone 1	Supply Active Power	Supply Reactive Power
Zone 2	Supply Active Power	Absorb Reactive Power
Zone 3	Absorb Active Power	Supply Reactive Power
Zone 4	Absorb Active Power	Absorb Reactive Power

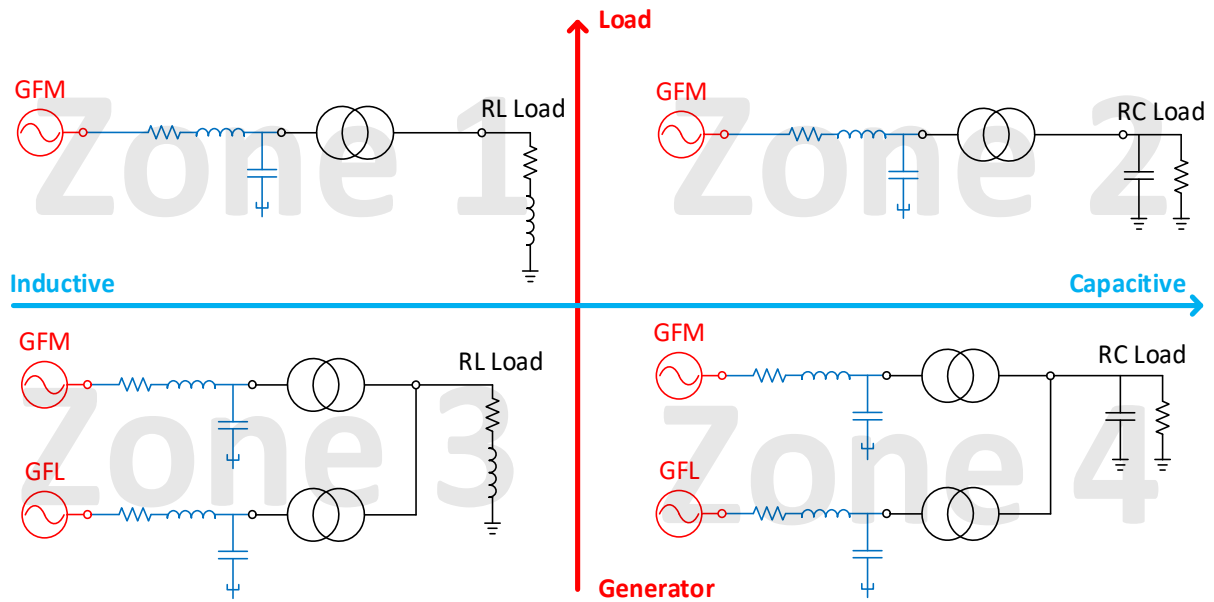


Figure 4.3: Circuit set up for small signal study under different zones of operation

4.2.1 Small-signal model and validation

A similar modeling and analysis approach from section 2.4 is adopted here and the modeling details are presented in the Appendix. The small-signal models for all four different zones all have been validated against the time-domain model, which is similar to the approach used in section 2.4.2.

During eigenvalue analysis, the GFM's small-signal model always yields a pole at the origin on the s -plane and such a pole was not observed in the GFL's small-signal model. Further checking on the participation factor of this pole indicates that this pole is contributed fully by the state of the PLL angle. For the GFM controller, the closed loop control is on the frequency of the system rather than the angle. This means that the relative angle of the system can settle at any arbitrary value, which is to be expected for GFM operation. Therefore, this pole at the origin does not indicate a marginally stable system and this pole is removed from the analysis when calculating damping.

4.2.2 Small-signal study results

Figure 4.4 shows the small-signal results of the proposed GFM operating under different zones. All combinations of operating points are tested and presented as a 2D diagram for each zone. X axis represents the amount of reactive power (capacitive or inductive), and the Y axis represents the amount of active power (positive or negative). The same approaches for eigenvalue analysis, damping ratio calculation, and representation of results from section 2.4.3 were adopted. Zone 1 and zone 2 with passive load are shown at the top and zone 3 and zone 4 with active load modeled as GFL are shown in the bottom.

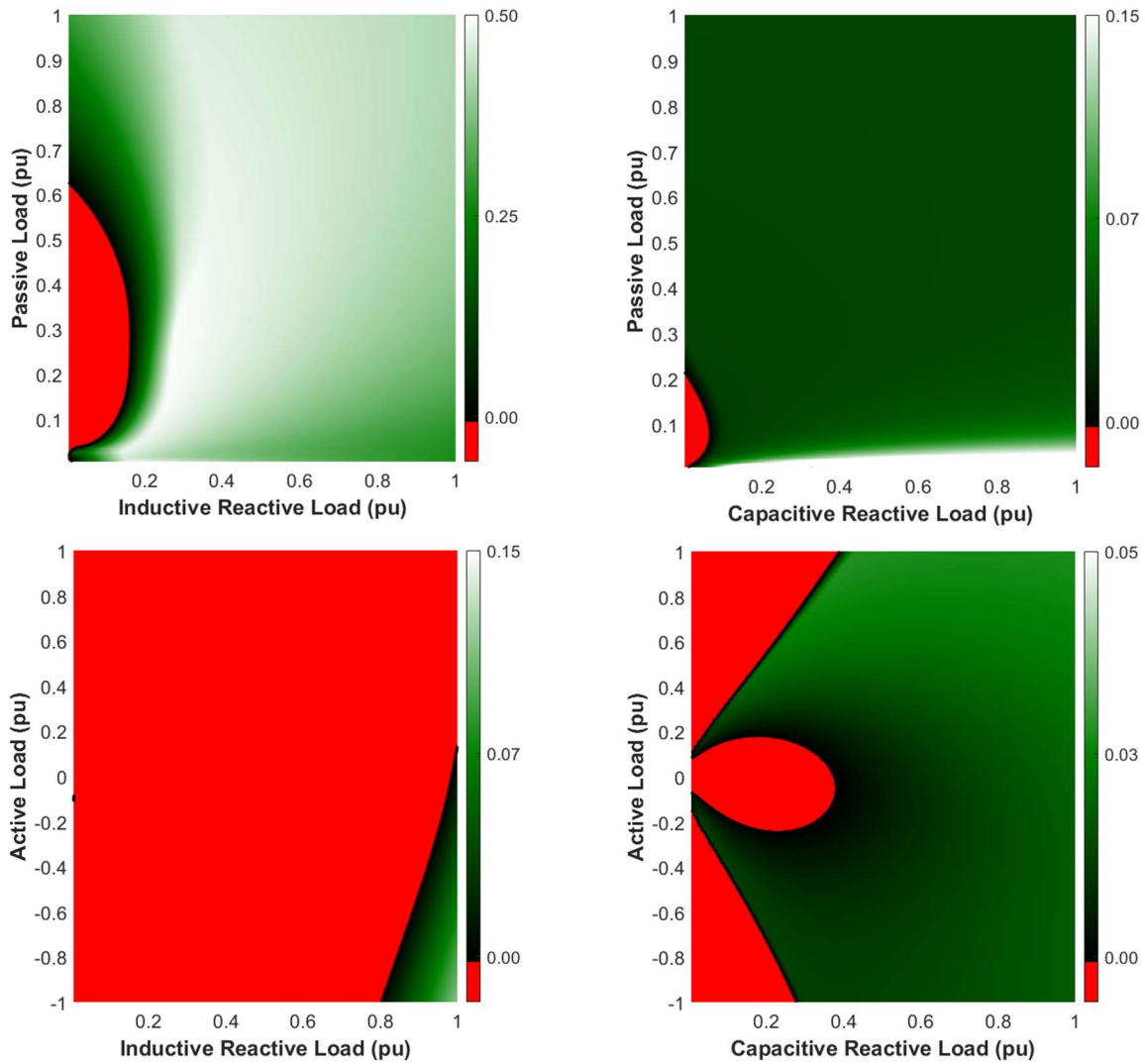


Figure 4.4: Small signal analysis results of the proposed GFM controller under different operating zones

For each zone, the minimum damping ratios are calculated for several combinations for various dispatch levels of active and reactive power. The number of combinations is selected to achieve a high-resolution view of the system stability over the whole range of operating conditions.

For zone 1 and zone 2 with the passive load, the system is not stable at low load conditions. This observation matches the observation from section 4.1 that the decoupled current controller has difficulty operating under high grid impedance. For zone 3, there are mostly no stable operating points and for zone 4 the system is also not stable at low load and high active power conditions.

In section 2.4.3 small-signal analysis for a GFL controller, the decoupled current control does have challenges operating under high grid impedance (ultra-low SCR conditions). Since the decoupled current controller is also used in the proposed GFM topology, it is expected for the inner current control loop to suffer from similar issues at low load conditions (i.e., high grid impedance).

It is observed that the controller is stable under high load conditions but unstable under low load conditions. This gives the inspiration of adding a virtual load to improve the stability of the proposed GFM control. This virtual load is introduced as virtual admittance (VA) to artificially reduce the impedance seen by the inverter. VA methodology has been proposed and investigated for GFM in the literature [41][42]. The traditional approach is to model the VA in the s -domain as $\frac{1}{R_v + sX_v}$ and transform it into the dq domain. The control loop would be similar to a decoupled current controller but in reverse direction with measured terminal voltage as inputs and VA current as output. The VA current output is then added to the current references of the decouple current controller to complete the augmentation of the virtual admittance. This implementation adds two additional control state variables and introduces significant complexity to the overall control topology. To avoid complexity, a simplified approach is used in this thesis. Based on instantaneous power theory [30], the inductor dynamics from the VA is removed and the VA control equations are simplified from the original differential equations to algebraic equations as shown in (4.1). GR and GX from the equation are the magnitude of the resistive and reactive portions of virtual load in per

unit. V_d and V_q are the measured voltage at the inverter terminal. I_{d_v} and I_{q_v} are the current reference for the VA that is to be added to the current reference of the decoupled current controller. Figure 4.5 shows the block diagram of the virtual admittance. To further improve the stability of the overall control system, the input voltages can be partially filtered. Part of GR and GX can be transferred to GR_f and GX_f path with additional filtering. This is discussed further later in this chapter.

$$\begin{bmatrix} I_{d_v} \\ I_{q_v} \end{bmatrix} = \begin{bmatrix} -GR & -GX \\ -GR & GX \end{bmatrix} \begin{bmatrix} V_d \\ V_q \end{bmatrix} \quad (4.1)$$

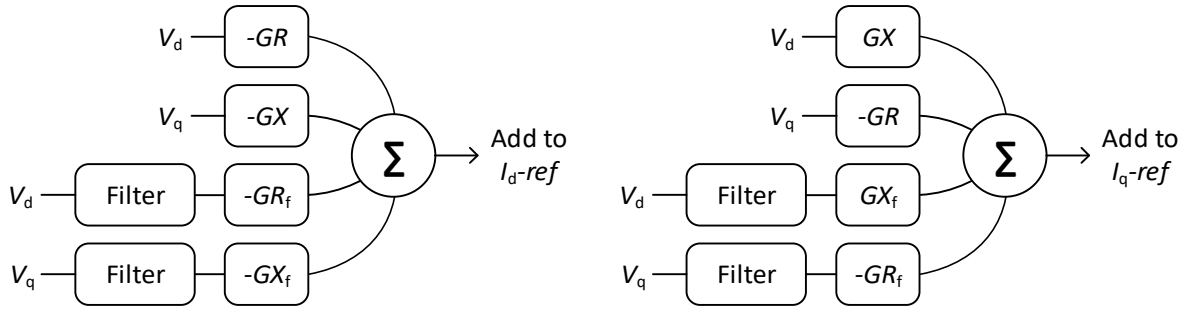


Figure 4.5: Control block diagrams of virtual admittance

4.2.3 Investigation of the VA

The stability of the GFM control with the addition of VA is evaluated in this section and a systematic way of evaluation is also developed. To better visualize the damping ratio for a given scenario the “System Stability Map” is rotated to its side as depicted in Figure 4.6 to show the load size verses the minimum damping ratio. The SCR is tested between 0.01 to 1 and the damping ratio is shown up to 0.5 on the Y axis.

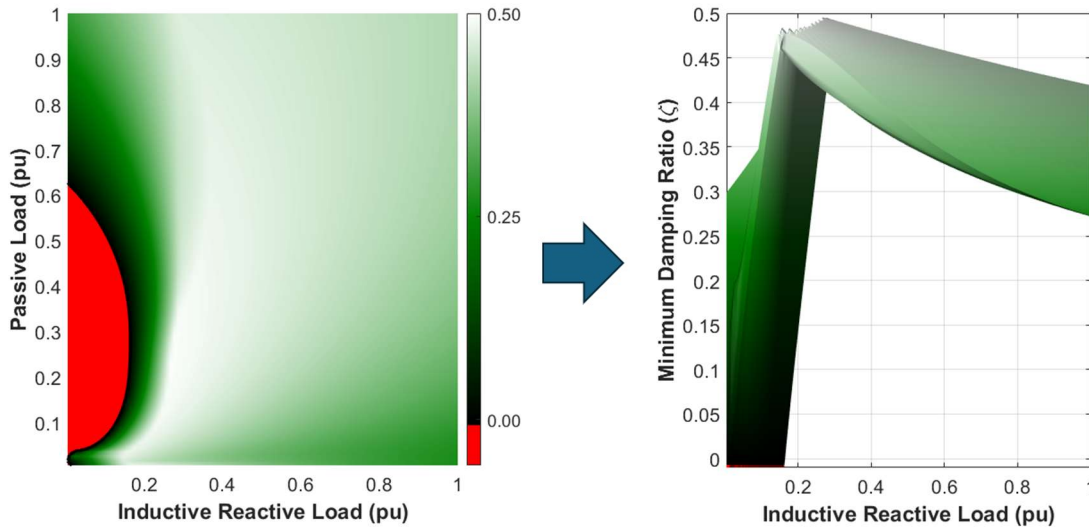


Figure 4.6: System stability map rotated to its side showing the load size vs. damping ratio

Figure 4.7 shows the stability of the system under different percentages of resistive VA (GR) and reactive VA (GX) for Zone 1. Due to the limited size of the chart, the chart labels and number marks are omitted. All figures have their X axis to be inductive load up to 1 pu and Y axis to be the minimum damping ratio up to 0.5. Both GR and GX varied between 25%, 100%, and 200%. From this test, adding the resistive VA can improve the system stability, but at a higher percentage can have diminishing effects. On the other hand, adding reactive VA generally reduces the stability of the system.

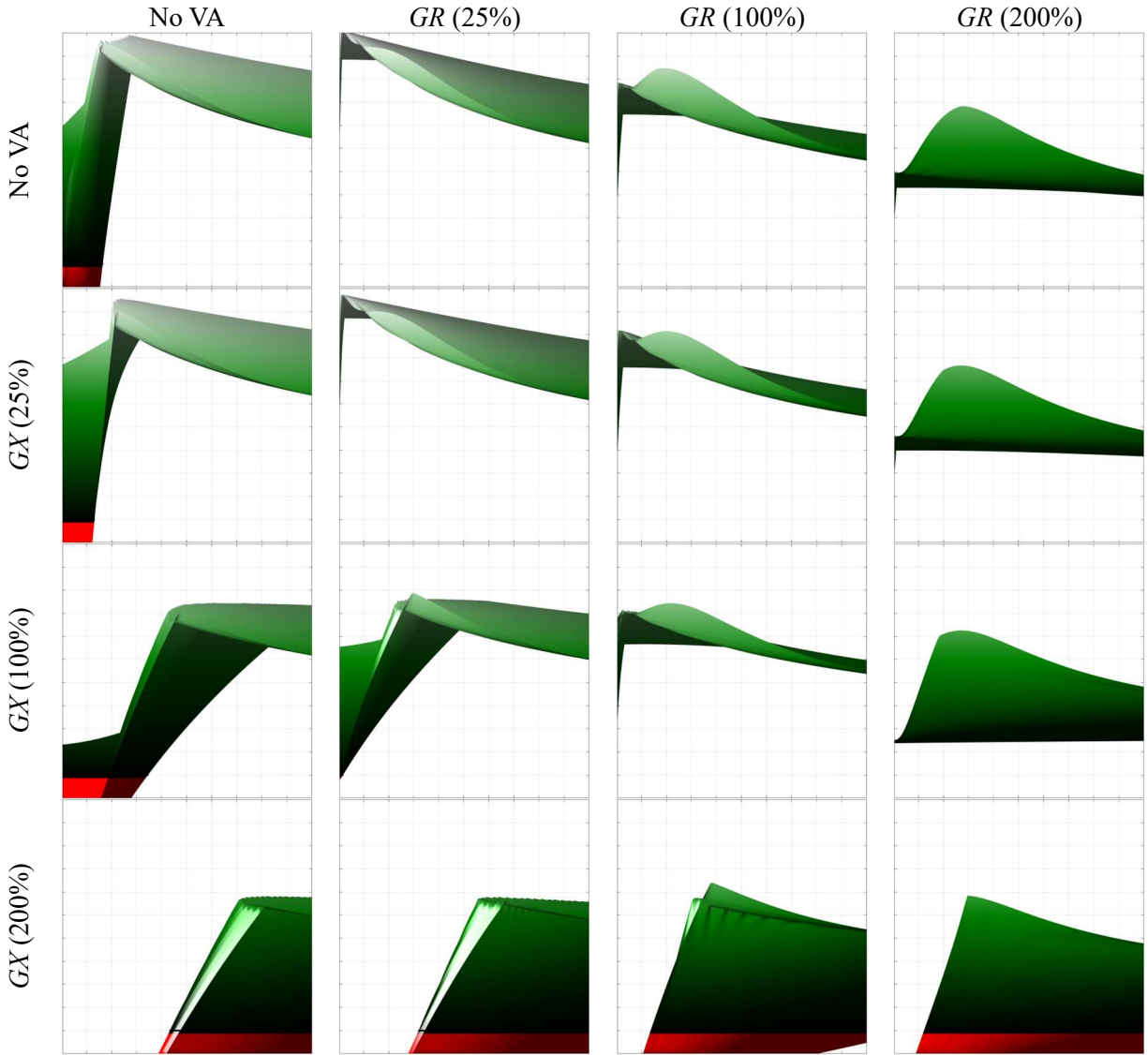


Figure 4.7: Stability maps of the system for zone 1 operation under combinations of GR and GX

A more thorough and systematic approach is developed and present in Figure 4.8. The average damping ratio for a given scenario is calculated under all tested operating conditions and projected on to a 2D plane as green. If a scenario has an operating point with a negative damping (unstable), it is marked as a red in the graph.

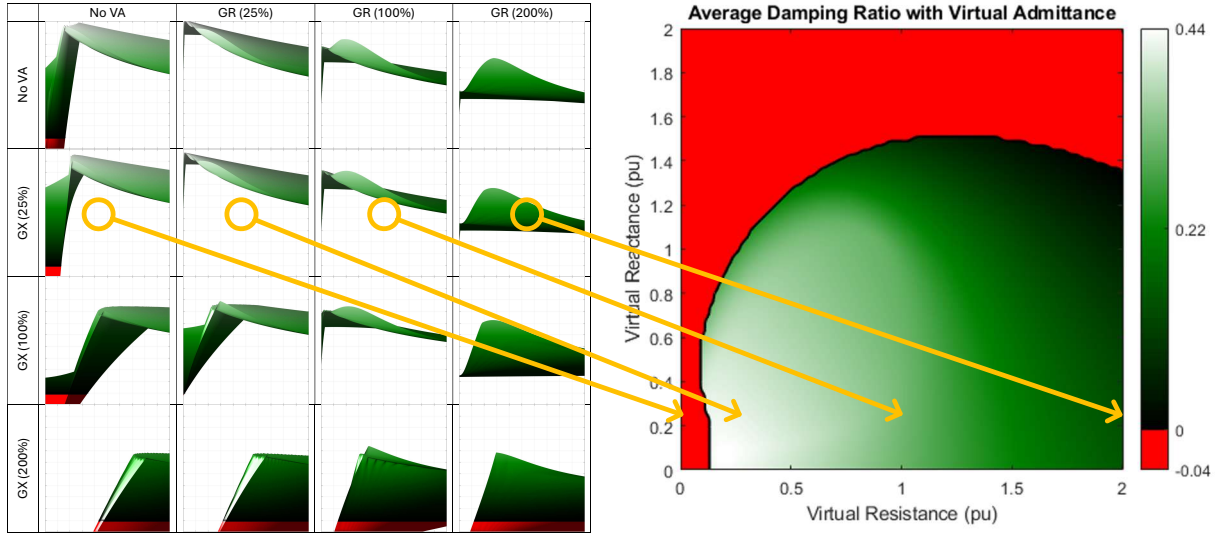


Figure 4.8: Development of the systematic approach

The systematic approach is implemented for all four zones as shown in Figure 4.9. A combination of virtual resistance and virtual reactance was tested for all four zones. For stable operation at all possible operating ranges for all 4 zones, the virtual resistance needs to be at least 1 pu and the virtual reactance needs to be less than 1 pu. Therefore, adding the VA can help the proposed GFM control topology to operate in all four zones.

Additionally, if there is design requires operating the GFM with a virtual reactance much higher than 1 pu, the input voltage to VA loop can be filtered to further improve its stability. Some sensitivity check suggests that filtering part of the voltage used for the virtual admittance calculations, especially on the reactive path, can significantly improve the control stability. For example, an GR and GX of 1.5 pu and 2.5 pu is outside the stable region from Figure 4.9. However, if the VA input voltage is partially filtered with a time constant of 1 ms for the resistive path and 10 ms for the reactive path, a region of stability common to all four zones can still be achieved.

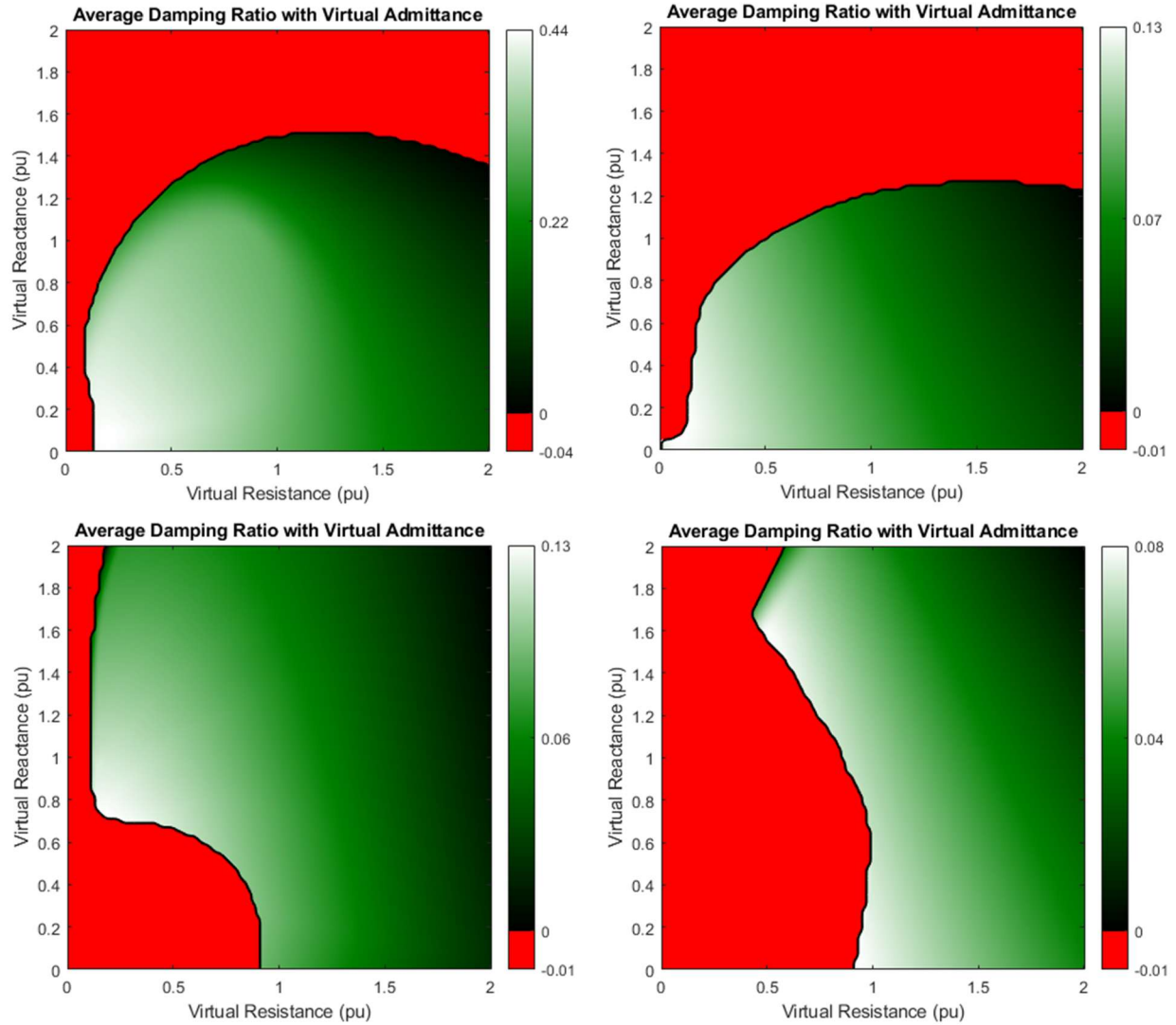


Figure 4.9: Study results of virtual admittance for GFM under all 4 zones

Figure 4.10 depicts the sensitivity checks and reveals that a 30% filtered voltage on the resistive path and 80% filtered voltage on the reactive path can achieve a stable operation in all four zones for an GR of 1.5 and G_X of 2.5 pu. Higher VA can further support the islanded operation with synchronous machine, and this is discussed in detail in the next section.

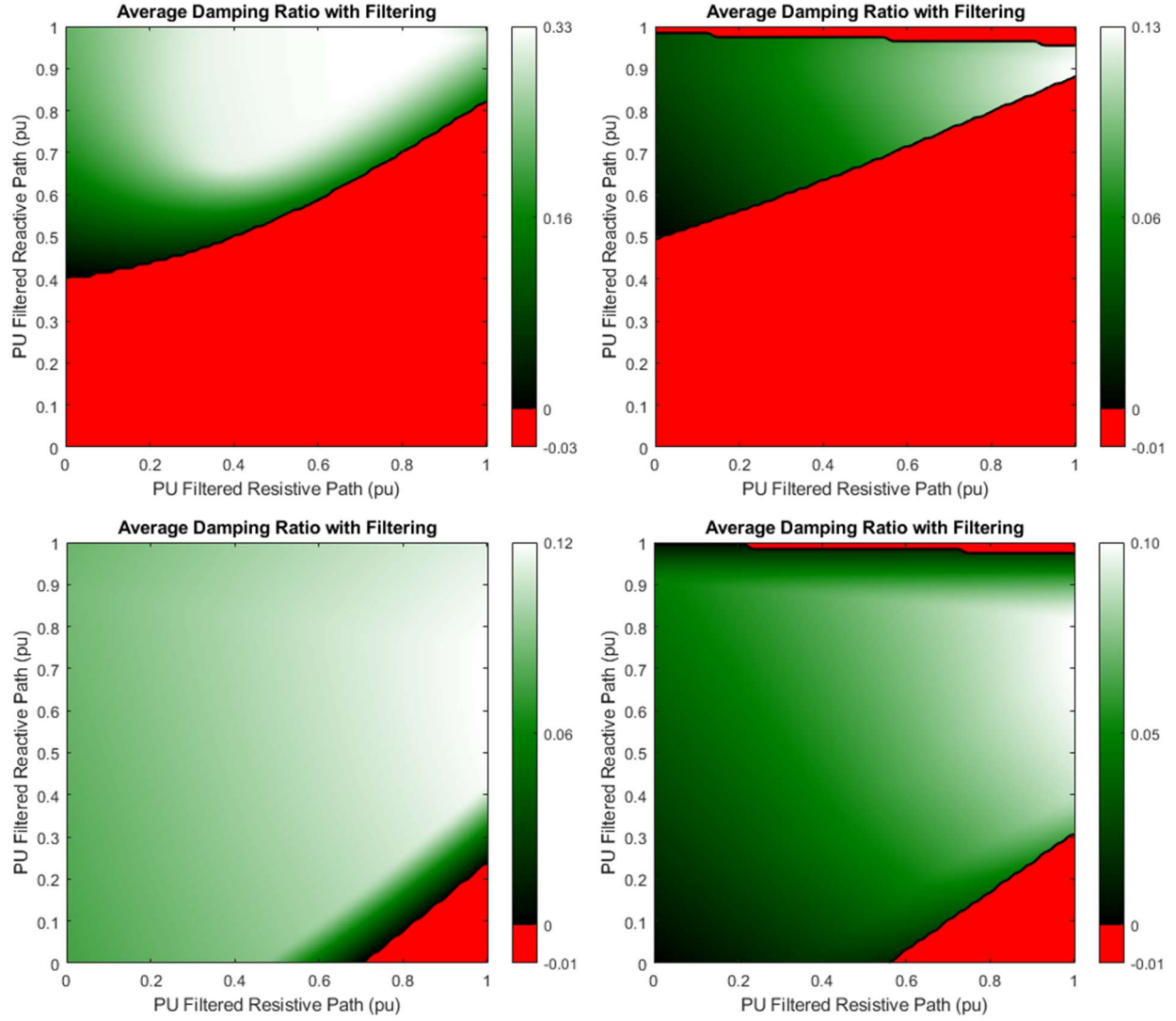


Figure 4.10: Sensitivity check on voltage filtering for the VA loops

The sensitivity of the frequency controller integral gain is also tested with the value increased to 0.05 from 0.01 and this increment did not meaningfully change the stability of the system. There is a possibility of further optimization of the VA and control parameters. This thesis focuses on finding a reasonable set of parameters to support the proposed topology and parameter optimization will be part of the future work. Apart from the small-signal study, spot checks have also been performed to ensure the small-signal results align with the time domain simulations. Furthermore, for fault ride-through with VA, the total current reference to the decoupled current controller is limited at 1 pu. The GFM voltage controller cannot integrate higher when the total current reference reaches 1 pu to prevent controller wind up.

4.3 Operation with synchronous machines

This section evaluates the performance of the proposed GFM when operating with a synchronous machine. Synchronous machines are well established electromechanical devices traditionally operated as generators and motors. Synchronous machines are designed for forming the grid and the rotating inertia of the rotor naturally resists sudden changes and helps to provide frequency stability to the grid. The damper windings of the machine can also provide sub-transient terminal voltage support, especially during faults through magnetic damping. These physical properties of the machine can help to provide short-term stability for the system; well-tuned exciters and governors help to maintain long-term stability.

A synchronous machine with generic modeling parameters is used and tested with the grid-forming controller through EMT simulations. The low value of inertia is selected to present a small synchronous machine typically found in micro grids such as a small diesel generator. Table 4.2 and Table 4.3 show the modeling parameters for the machine and the exciter, respectively. Since the dynamics from the governor is relatively slow for the study window and a well-tuned governor can potentially help prevent the pole slipping, a constant torque is used as a worst-case scenario. Some initial sensitivity check revealed that even if there is no small-signal instability for a given scenario when operating with the GFM, large-signal instability can often arise due to pole slipping of the generator upon fault recovery. Therefore, a large-signal study is carried out to determine the stability of the GFM with a generator.

Table 4.2: Machine parameters used for the study

Machine Parameters	Values
Base MVA	0.5 MVA
Base Voltage	0.6 kV
Inertia Constant	1 s
Armature Resistance	0.01 pu
X_d	2 pu
X_d'	0.3 pu
X_d''	0.2 pu
T_{d0}'	2s
T_{d0}''	0.012 s
X_q	1.1 pu
X_q''	0.2 pu
T_{q0}''	0.007 s

Table 4.3: Machine exciter ST2A modeling parameters

Exciter Parameters	Quantities
Exciter Time Constant	0.5 s
Exciter Field Constant	1 pu
Max Exciter Output	3.6 pu
Compounding Circuit Voltage Multiplier	4.9 pu
Compounding Circuit Current Multiplier	8 pu
Rectifier Loading Factor	1.8 pu

A set of tests were selected over the various dispatch levels of a synchronous machine. It is assumed that P_{max} and Q_{max} are 0.8 pu and 0.3 pu of the machine base MVA, respectively. The 0.8 pu limit reflects the typical DER maximum dispatch level (80–90%), while 0.3 pu reactive power represents a reasonable plant power factor capability. Twenty-five tests are derived from a combination of 1 pu, 0.5 pu, 0 pu, -0.5 pu and -1 pu P_{max} or Q_{max} , and three different sizes of machines are tested. These three sizes are 25% MVA of the GFM, 50% of the GFM, and 100% of the GFM. The simulation results are summarized in Table 4.4. The originally proposed GFM is tested with a fault cycle of 3 and it appears to fail for most of the simulations.

Table 4.4: Summary of FRT performance of proposed GFM operating with synchronous machine

Cyc.	P\Q	Proposed GFM														
		25%					50%					80%				
		1	0.5	0	-0.5	-1	1	0.5	0	-0.5	-1	1	0.5	0	-0.5	-1
3	1	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail
	0.5	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail
	0	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail
	-0.5	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Pass	Pass
	-1	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail	Fail

Additional investigation was carried out and revealed that the originally proposed GFM topology has a poor performance in supporting the operation of a traditional machine. This is because the proposed GFM does not follow a traditional coupling characteristic such as voltage vs. Q and angle vs. P. The traditional coupling in the power system has a droop relationship such that a dip in the grid voltage would cause an increase in the reactive power from the generator while an angle dip would cause an increase in active power. A set of time-domain simulations of the proposed GFM as summarized in Figure 4.11 does not show the desired response. A small voltage dip and an angle dip are applied to the GFM terminal, and the responding immediate reaction in power and reactive power is recorded (i.e. the delta changes in P and Q are recorded). The desired response for the GFM should be uniformly positive for delta Q (red, left) and delta P (red, right), but the actual response highly depends on the initial condition of the GFM.

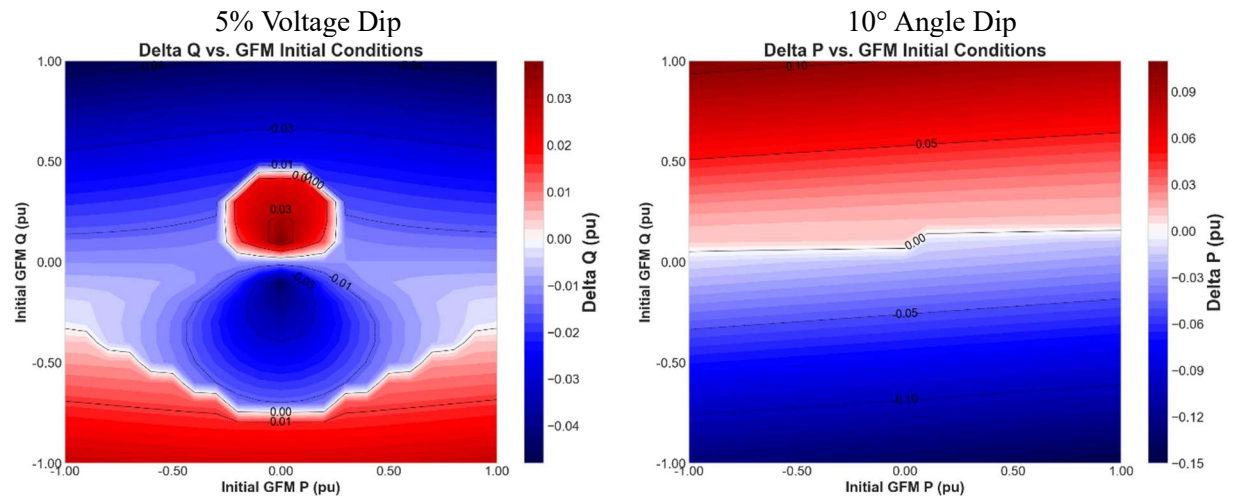


Figure 4.11: Coupling characteristics of the proposed GFM

The traditional coupling characteristics work extremely well with a power system that has mostly an inductive network and rotating machines. During a voltage sag, the increase in Q output can help boost the terminal voltage. Moreover, an angle dip or a frequency dip from the grid typically indicates a relative surge in demand and additional power support can help restore the angle and frequency.

Although the originally proposed GFM does not have such a behavior, once the VA was added as discussed from section 4.2.2 and 4.2.3, the operating point of the system can be biased towards a region where the proposed GFM has a behavior similar to the traditional coupling characteristics. The behavior of the GFM is further illustrated in Figure 4.12. The originally proposed GFM topology is shown on the left and the GFM with VA is shown on the right. The two red vectors are current references presenting two arbitrary operating conditions.

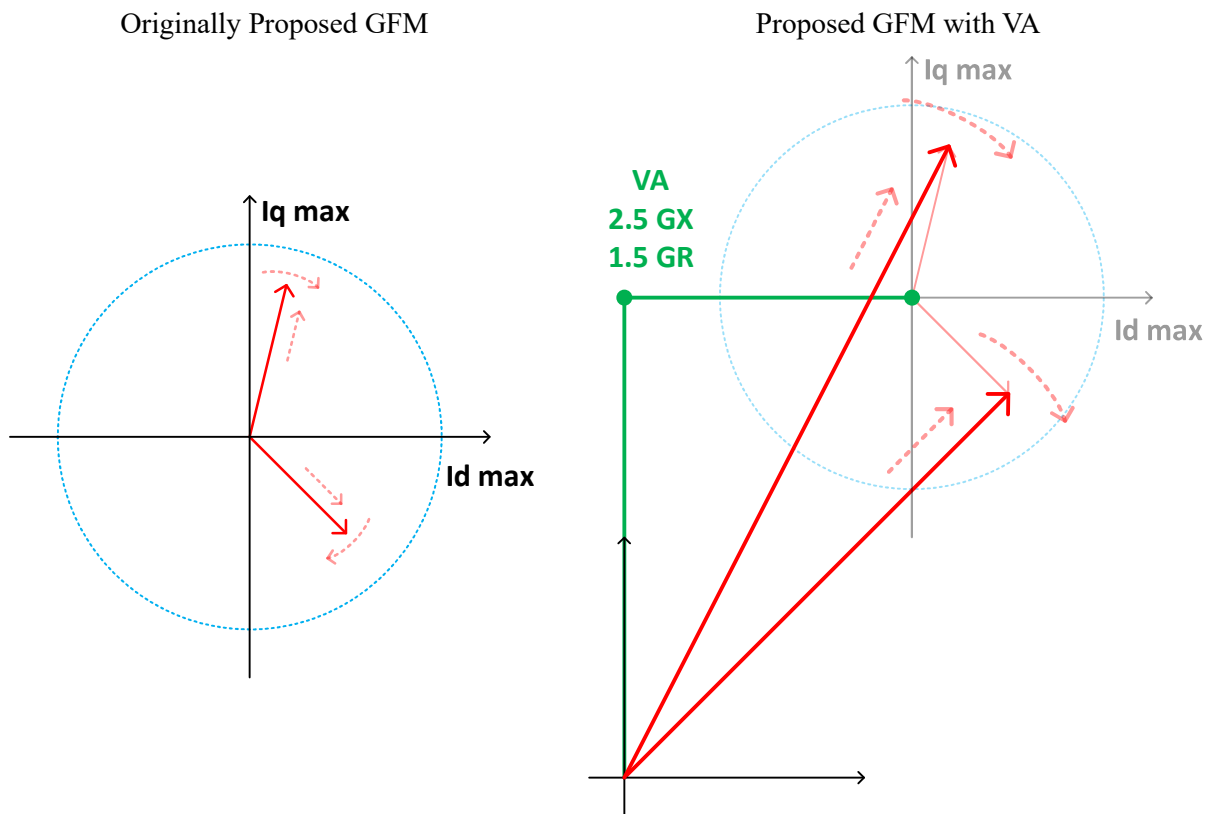


Figure 4.12: Operating behaviour of proposed GFM comparison under voltage and angle dip

Under a voltage dip, the length of the vector is increased as part of the GFM response. For the GFM control without VA, depending on the initial direction of the vector, the increase in vector can be inductive or capacitive. For GFM with VA, as the operating point is biased towards positive, the response of the GFM will always be capacitive for a voltage dip.

Additionally, for an angle dip test, the drop in grid angle would have a similar effect of a load angle dip as discussed in section 3.4. This drop in terminal angle would cause the controller to rotate clockwise. For the GFM without VA, the response of the IBR can be positive or negative in power depending on the initial Q condition. However, for GFM with VA, it will always provide active power during such an event.

The same test from Figure 4.11 was repeated for the proposed GFM topology with VA as shown in Figure 4.13. With VA, the proposed GFM also can exhibit the traditional coupling characteristic. The VA tested in this case is with a G_X of 2.5 pu and G_R of 1.5 pu. It is expected for the GFM to have a greater support with a higher G_X value. This can be deduced from Figure 4.12. Adding VA is also analogous for the GFM with a current controller to behave more like a voltage source, and more VA equivalent to less source impedance. This allows the GFM to provide synchronous-machine-like inertia support in active and reactive power for grid angle and voltage disturbances.

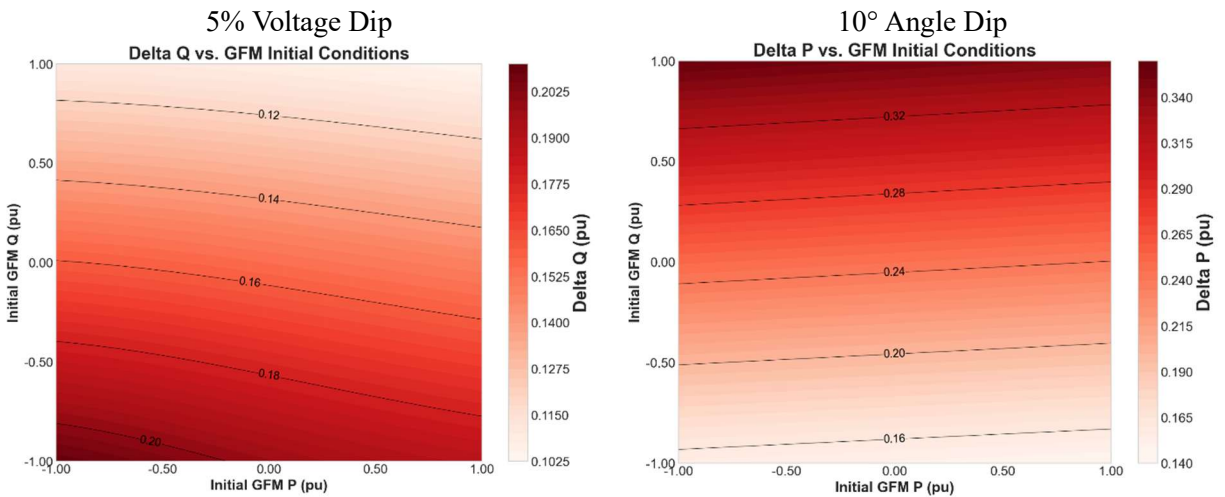


Figure 4.13: Coupling characteristics of the proposed GFM with VA

The same simulation from Table 4.4 is repeated with VA and a detailed switching model. A few fault cycles are also evaluated from 3 to 8. To have a reference for the performance of the proposed GFM, the control group GFM PWM switches are replaced with an ideal voltage source with a fixed voltage and frequency. The results are summarized in Table 4.5; the green cells indicate a successful recovery, and the red cells indicate a failed one. The proposed GFM with VA has similar performance of an ideal voltage source.

Table 4.5: Simulation summary of FRT performance GFM with VA operating with synchronous machine

Cyc.	P\Q	Ideal Source															Proposed GFM with VA														
		25%					50%					80%					25%					50%					80%				
		1	0.5	0	-1	-1	1	0.5	0	-1	-1	1	0.5	0	-1	-1	1	0.5	0	-1	-1	1	0.5	0	-1	-1	1	0.5	0	-1	-1
3	1																														
	0.5																														
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	-0.5																														
	-1																														
9	1																														
	0.5																														
	0																														
	-0.5																														
	-1																														

4.4 Summary

This chapter assessed the small and large signal performance of the proposed GFM. The originally proposed topology works well under passive load condition. With the help of a simplified virtual admittance loop, the proposed control topology can operate in parallel with other IBRs and synchronous generators and helps to provide voltage and angle inertia.

Chapter 5

Revisiting the GFL and Grid-Firming Control

A GFM control strategy has been developed in the previous chapter. With the help of VA, significant control stability improvement was observed. Although a GFL cannot form the grid, it can firm the grid by improving overall stability of its surrounding network and rejects undesirable oscillations. In this chapter, VA used for GFM is also implemented for conventional GFL control to improve its transient stability and enable its grid firming capabilities.

5.1 Virtual Admittance for Super-Synchronous Oscillations

One of the key insights from the GFM control logic development is the implementation of the virtual admittance. In the original control logic of the GFM, instability often occurs under extremely low load conditions. This is especially true when the GFM is operating under zone 1 with inductive loads. From a small signal point of view, zone 1 operation is the most similar to the typical operating environment of a GFL where the impedance seen by the inverter is mostly inductive. When adding the virtual admittance (GR) to GFM model under zone 1, a significant improvement in control stability is observed. Therefore, further experiments are carried out to determine if adding a small amount of virtual admittance to GFL control will improve the operating stability of a GFL inverter.

The tests from section 2.4.3 are repeated with the addition of GR . Unlike the tests from section 1 where the focus is on the inverter performance under extremely low SCR, the focus on these tests is around the worst-case operating conditions of the GFL inverter around an SCR of 1.2 – 2. The following tests are performed under unity power factor with the active power ranging from -0.9 pu to 0.9 pu active power. Figure 5.1 shows the system's damping ratio comparison between the original GFL control and GFL control with

virtual admittance. The system's minimum damping ratio seems to be rather uniform across different levels of I_d injection as shown from the size view below that the distribution is within a very narrow band. This indicates that further analysis performed on a particular I_d dispatch level is also applicable to others. Further eigenvalue analysis is performed under SCR of 1.2 and I_d to be 0.8 pu. As the GR increased from 0 to 20%, the stability of the system can be further improved for super synchronous modes around 100 and 200 Hz.

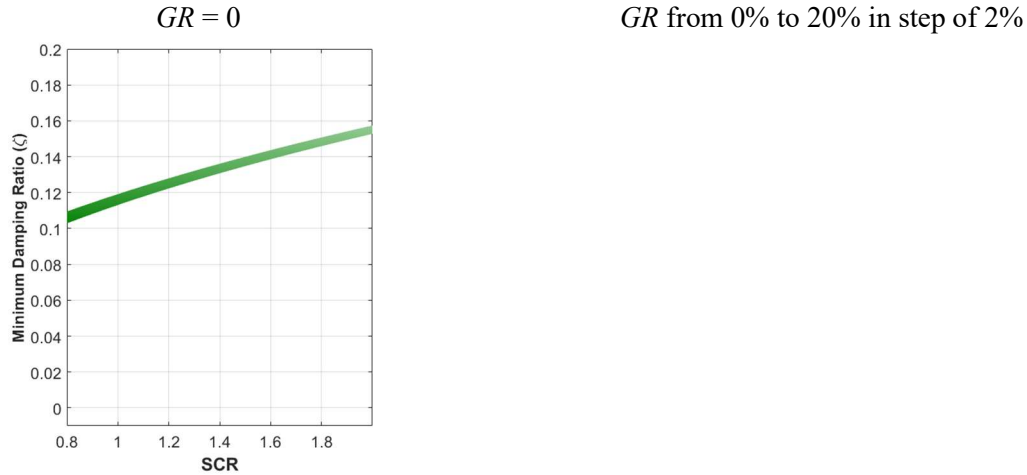


Figure 5.1: Comparison of the damping ratio for various levels of resistive virtual admittance (GR)

Additional time-domain simulation is also performed to compare the performance differences between the original GFL control and the new controls with the addition of virtual admittance. For the test, a 0.1pu I_d step change is applied to the controllers.

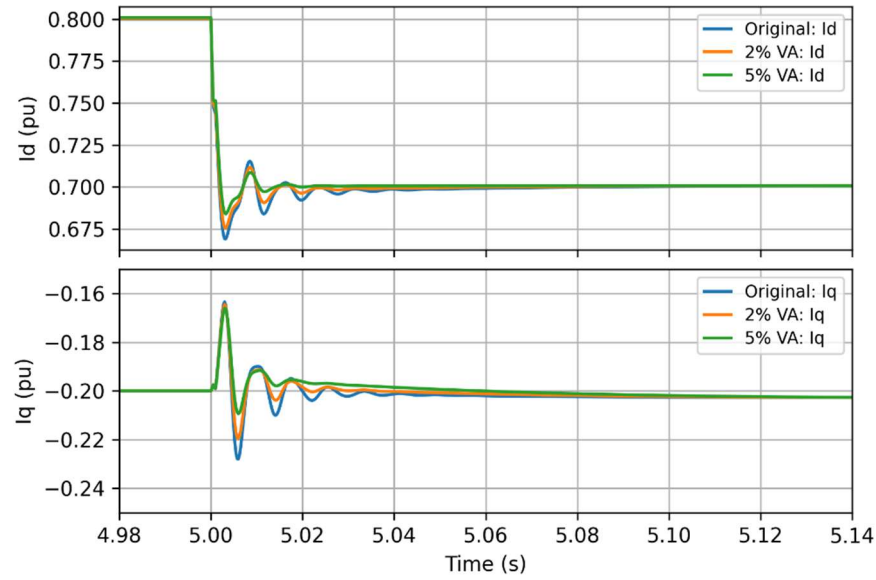


Figure 5.2: Performance comparison of GFL inverter control with resistive virtual admittance (GR)

Figure 5.2 shows the differences in control responses. It can be observed that higher the resistive virtual admittance, the better the damping in the fast transient response. These tests are performed under an SCR of 1.2. Therefore, under low SCR conditions, adding small amounts of resistive virtual admittance can help improve the controller damping for super synchronous modes.

5.2 Virtual Admittance for Sub-Synchronous Oscillations

From section 4.3, adding the virtual admittance can transform the behavior of a current source controller to a voltage source analog to Norton-Thevenin equivalent. This technique can also be applied to GFL to add grid-firming effect to the GFL and provide additional damping for sub-synchronous oscillations. Since supporting the grid's stability can be a secondary priority for GFL, a washout filter can be implemented to the VA loop so that the grid-firming effect is only temporary during grid voltage and angle transients. The duration of the effect can be adjusted through the time constant of the washout filter. Initial sensitivity study suggested that a resistive-dominated VA loop can offer slightly better damping performance compared to a reactive-dominated VA loop under the same VA magnitude. However, considering some GFL such as solar and wind may not have readily available active power head room, a resistive-dominated VA loop could overly discharge the DC-link capacitor and affect the normal operation of the plant. A reactive dominated VA loop is selected.

A small-signal analysis is carried out to study the GFL control stability with the addition of VA loops. The virtual admittance parameters are tuned through trial and error and a reasonable set of parameters is selected with GR to be 1.33 and GX to be 3. From lessons learnt in section 4.3, a higher reactive virtual admittance should provide more grid-firming effect. However, higher reactive virtual admittance without some resistive portion can also be unstable. Therefore, some resistive virtual admittance is also added. The results of the small-signal analysis for these set of virtual admittance parameters are shown in Figure 5.3. The controller is stable over the entire active power operating range for SCR over 1.

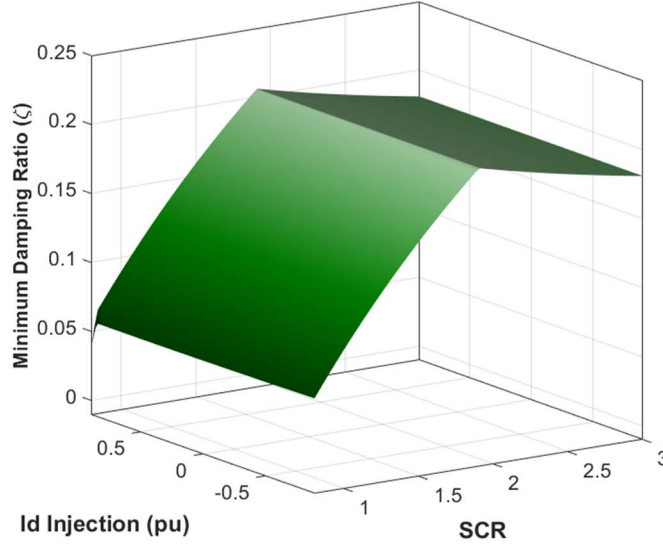


Figure 5.3: Small signal analysis for GFL with virtual admittance

To further investigate the damping of the GFL control with VA, grid oscillation rejection test is performed for both the voltage oscillation and angle oscillation. The methodology of these tests is listed below, and they are derived from [43][44].

- An oscillation under a specific frequency is modulated to the network equivalent voltage source voltage magnitude or angle.
- A disturbance magnitude of 2% and 2° is selected for the voltage and angle respectively
- The network equivalent voltage source is behind the network impedance corresponding to an SCR of 3.
- The oscillation in voltage and angle is measured both at the voltage source and the connection point of the IBR.

For the IBRs to positively damp the grid oscillation, the oscillation magnitude at the POI should be less than the oscillation at the source of modulation behind the source impedance. Sensitivity test is performed under various levels of VA to investigate the amount of VA and its corresponding effect. Assuming 100% VA is 1.33 *GR* and 3 *GX*, tests are performed to GFL with 0% VA, 25% VA, 50% VA, and 100% VA is performed.

Figure 5.4 and Figure 5.5 depict the results of the sensitivity tests. The observed voltage and angle oscillations at the POI of the GFL is shown as a percentage of the grid oscillation. Over the tested frequency of 2-45 Hz where the sub-synchronous oscillation issue is the most prevalent [44], the GFL with virtual admittance shows great rejection in grid oscillations. The higher the VA, the better the GFL can reject the sub-synchronous oscillations. Furthermore, under both rejection tests, the grid active power and reactive power is also observed to be out of phase of the modulated oscillation respectively and further indicates that the GFL control with VA can helps to reduce the oscillation in angle or voltage.

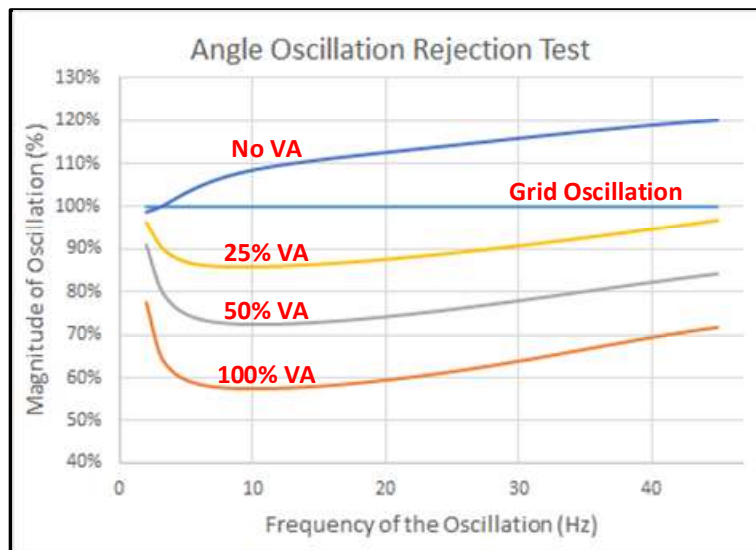


Figure 5.4: Angle oscillation rejection test

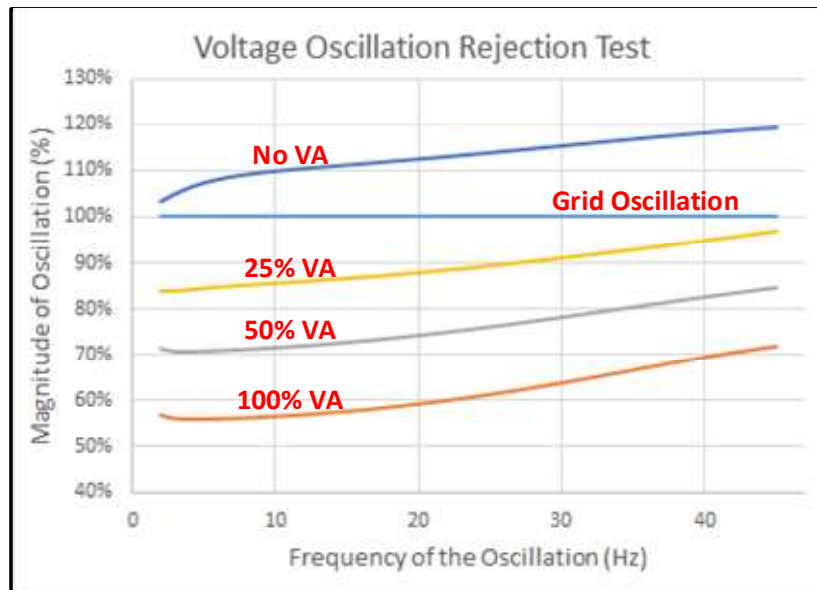


Figure 5.5: Voltage oscillation rejection test

Chapter 6

Contributions, Conclusions, and Future work

This thesis investigated the performance of the state-of-the-art decoupled current controller widely used in grid-following inverter-based resources (GFL IBRs). Small-signal analysis was conducted to evaluate the sensitivity of key control parameters, while large-signal simulations were used to examine its limitations during islanded operation. Motivated by these limitations, this thesis investigated simple yet effective control topologies that allow a conventional GFL controller to grid forming or enable its grid firming ability.

6.1 Contributions

The contributions of the thesis are listed below:

- This thesis examined the small-signal stability of the decoupled current controller under ultra-low short-circuit ratio (SCR) conditions, an area that has received limited attention in the literature. For GFM topologies that retain the decoupled current controller as the inner loop, this work provides insight into the stability limits of the controller under ultra-low SCR conditions and the influence of key control parameters on stability improvement.
- This thesis investigated the large-signal response of the decoupled current controller during islanded operation. The behavior of the controller under islanded conditions was formulated, providing important insight for the design of grid-forming control strategies.
- This thesis proposed a minimalistic GFM control topology that can be added to a conventional GFL inverter employing decoupled current control, thereby enabling grid-forming operation. Both

small-signal and large-signal studies were carried out to evaluate the performance and limitations of the proposed topology.

- A simplified virtual admittance loop was incorporated into the proposed GFM controller to improve its stability and dynamic performance.
- This thesis also investigated the grid-firming capability of the simplified virtual admittance loop when applied to a GFL inverter with decoupled current control.

6.2 Conclusions

The conclusions of this thesis are listed below:

- The decoupled current controller exhibits strong small-signal performance even under very low SCR conditions. However, the results indicate a practical stability limit at an SCR of approximately 0.1-0.2. Although parameter tuning can provide modest improvement, it cannot eliminate the instability when the SCR falls below this range.
- The decoupled current controller is unstable under islanded operation. The rate of divergence is strongly influenced by the PLL parameters and by the mismatch between the current-reference angle and the load angle.
- The proposed GFM topology can convert a GFL inverter with a decoupled current control loop into a GFM inverter through a simple control augmentation. It provides a practical and effective solution for systems in which the load is predominantly passive.
- For more complex operating conditions, such as systems containing other IBRs or rotating machines, the inclusion of the simplified virtual admittance loop is necessary to maintain control stability.
- For existing GFL inverters, the simplified virtual admittance loop also offers a simple and effective means of improving grid stability and providing grid-firming support.

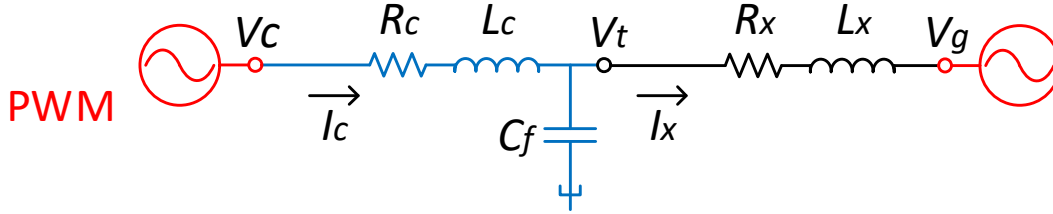
6.3 Future work

Potential avenues for future work on this topic are listed below:

- Most of the analysis and control design are performed only for positive-sequence control. Additional work can be carried out to extend the research to negative-sequence control and control parameter optimization.
- A comparison between the proposed GFM topology and other existing GFM topology in the literature and industry.
- Additional sensitive check can be carried out to study the impact of various converter design to the proposed control topologies such as various designs for grid impedance and filters, additional DC to DC converter on the DC side.
- The proposed GFM uses a simplified virtual admittance loop for the simplicity of the design and easiness of implementation. However, the detailed virtual admittance loop that considers the dynamics of the reactive path may offer additional control stability. This can be further investigated in future work.
- This thesis centered around the tests under a small size system; no external PPC dynamics are considered. Further work can be carried out to further study the proposed topology under wide area network with a more complex conditions and the inclusion of an external PPC.
- Additional test and development for GFM with energy resources with an unpredictable active power constraint such as wind and solar.

Appendix – Small signal model formulation

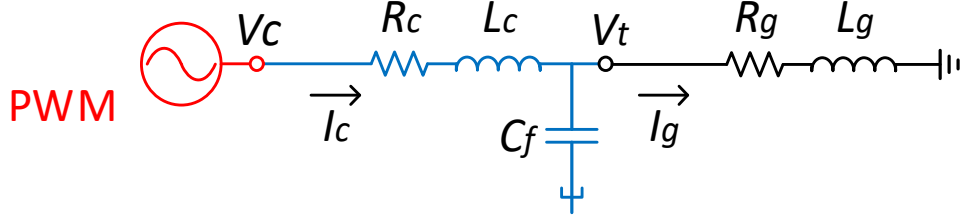
A.1 Passive Network of Grid Connected GFL



$$\begin{bmatrix} \dot{I}_{xd} \\ \dot{I}_{xq} \\ \dot{I}_{cd} \\ \dot{I}_{cq} \\ \dot{V}_{td} \\ \dot{V}_{tq} \end{bmatrix} = \begin{bmatrix} -\frac{R_x}{L_x} & \omega_{sys} & 0 & 0 & \frac{1}{L_x} & 0 \\ -\omega_{sys} & -\frac{R_x}{L_x} & 0 & 0 & 0 & \frac{1}{L_x} \\ 0 & 0 & -\frac{R_c}{L_c} & \omega_{sys} & -\frac{1}{L_c} & 0 \\ 0 & 0 & -\omega_{sys} & -\frac{R_c}{L_c} & 0 & -\frac{1}{L_c} \\ -\frac{1}{C_f} & 0 & \frac{1}{C_f} & 0 & 0 & \omega_{sys} \\ 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{xd} \\ I_{xq} \\ I_{cd} \\ I_{cq} \\ V_{td} \\ V_{tq} \end{bmatrix} + \begin{bmatrix} -\frac{V_{gd}}{L_x} \\ -\frac{V_{gq}}{L_x} \\ \frac{V_{cd}}{L_c} \\ \frac{V_{cq}}{L_c} \\ 0 \\ 0 \end{bmatrix}$$

$$\begin{bmatrix} \dot{I}_{gd} \\ \dot{I}_{gq} \end{bmatrix} = \begin{bmatrix} -\frac{R_g}{L_g} & \omega_{sys} \\ -\omega_{sys} & -\frac{R_g}{L_g} \end{bmatrix} \begin{bmatrix} I_{gd} \\ I_{gq} \end{bmatrix} + \frac{1}{L_g} \begin{bmatrix} V_{td} \\ V_{tq} \end{bmatrix} - \frac{1}{L_g} \begin{bmatrix} V_{netd} \\ V_{netq} \end{bmatrix}$$

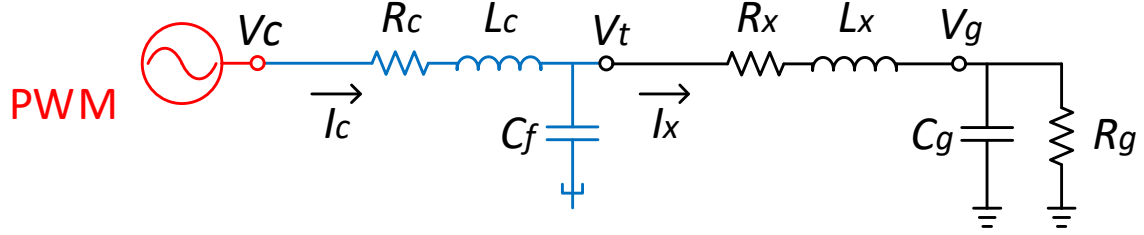
A.2 Passive Network of GFM in Zone 1



$$\begin{bmatrix} \dot{I}_{gd} \\ \dot{I}_{gq} \\ \dot{I}_{cd} \\ \dot{I}_{cq} \\ \dot{V}_{td} \\ \dot{V}_{tq} \end{bmatrix} = \begin{bmatrix} -\frac{R_g}{L_g} & \omega_{sys} & 0 & 0 & \frac{1}{L_g} & 0 \\ -\omega_{sys} & -\frac{R_g}{L_g} & 0 & 0 & 0 & \frac{1}{L_g} \\ 0 & 0 & -\frac{R_c}{L_c} & \omega_{sys} & -\frac{1}{L_c} & 0 \\ 0 & 0 & -\omega_{sys} & -\frac{R_c}{L_c} & 0 & -\frac{1}{L_c} \\ -\frac{1}{C_f} & 0 & \frac{1}{C_f} & 0 & 0 & \omega_{sys} \\ 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{gd} \\ I_{gq} \\ I_{cd} \\ I_{cq} \\ V_{td} \\ V_{tq} \end{bmatrix} + \frac{1}{L_g} \begin{bmatrix} 0 \\ 0 \\ V_{cd} \\ V_{cq} \\ 0 \\ 0 \end{bmatrix}$$

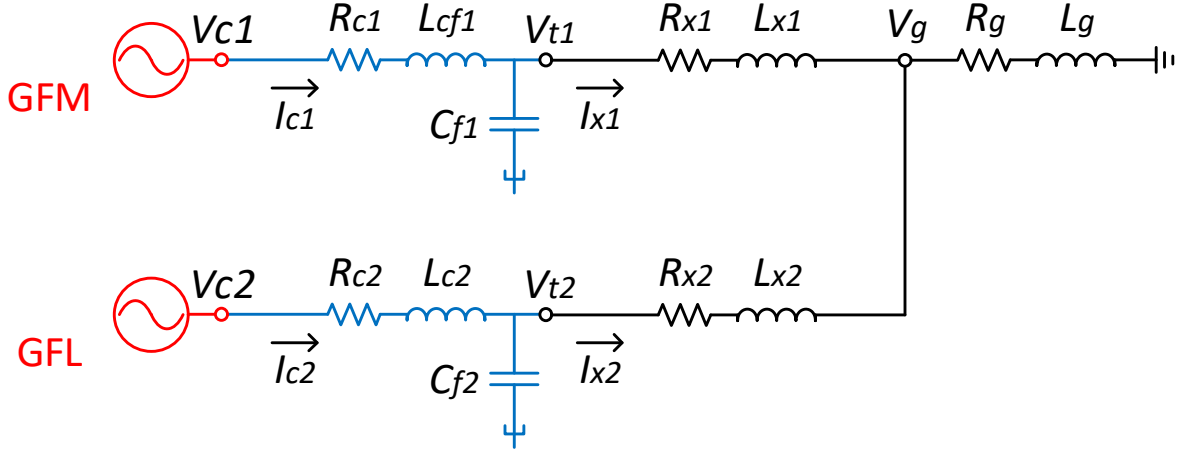
$$\begin{bmatrix} \dot{I}_{gd} \\ \dot{I}_{gq} \end{bmatrix} = \begin{bmatrix} -\frac{R_g}{L_g} & \omega_{sys} \\ -\omega_{sys} & -\frac{R_g}{L_g} \end{bmatrix} \begin{bmatrix} I_{gd} \\ I_{gq} \end{bmatrix} + \frac{1}{L_g} \begin{bmatrix} V_{td} \\ V_{tq} \end{bmatrix}$$

A.3 Passive Network of GFM in Zone 2



$$\begin{bmatrix} \dot{V}_{gd} \\ \dot{V}_{gq} \\ \dot{I}_{xd} \\ \dot{I}_{xq} \\ \dot{I}_{cd} \\ \dot{I}_{cq} \\ \dot{V}_{td} \\ \dot{V}_{tq} \end{bmatrix} = \begin{bmatrix} -\frac{1}{R_g C_g} & \omega_{sys} & \frac{1}{C_g} & 0 & 0 & 0 & 0 & 0 \\ -\omega_{sys} & -\frac{1}{R_g C_g} & 0 & \frac{1}{C_g} & 0 & 0 & 0 & 0 \\ -\frac{1}{L_x} & 0 & -\frac{R_x}{L_x} & \omega_{sys} & 0 & 0 & \frac{1}{L_x} & 0 \\ 0 & -\frac{1}{L_x} & -\omega_{sys} & -\frac{R_x}{L_x} & 0 & 0 & 0 & \frac{1}{L_x} \\ 0 & 0 & 0 & 0 & -\frac{R_c}{L_c} & \omega_{sys} & -\frac{1}{L_c} & 0 \\ 0 & 0 & 0 & 0 & -\omega_{sys} & -\frac{R_c}{L_c} & 0 & -\frac{1}{L_c} \\ 0 & 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & 0 & 0 & \omega_{sys} \\ 0 & 0 & 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} V_{gd} \\ V_{gq} \\ I_{xd} \\ I_{xq} \\ I_{cd} \\ I_{cq} \\ V_{td} \\ V_{tq} \end{bmatrix} + \frac{1}{L_c} \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ V_{cd} \\ V_{cq} \\ 0 \\ 0 \end{bmatrix}$$

A.4 Passive Network of GFM in Zone 3

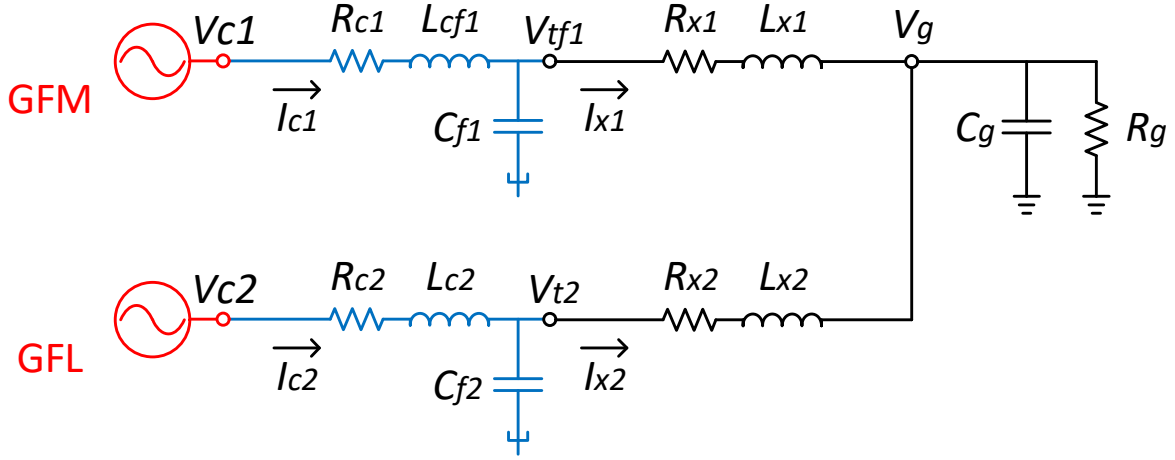


$$\begin{bmatrix} \dot{I}_{c1d} \\ \dot{I}_{c1q} \\ \dot{V}_{t1d} \\ \dot{V}_{t1q} \end{bmatrix} = \begin{bmatrix} -\frac{R_{c1}}{L_{c1}} & \omega_{sys} & -\frac{1}{L_{c1}} & 0 \\ -\omega_{sys} & -\frac{R_{c1}}{L_{c1}} & 0 & -\frac{1}{L_{c1}} \\ \frac{1}{C_{f1}} & 0 & 0 & \omega_{sys} \\ 0 & \frac{1}{C_{f1}} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{c1d} \\ I_{c1q} \\ V_{t1d} \\ V_{t1q} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_{c1}} & 0 & 0 & 0 \\ 0 & \frac{1}{L_{c1}} & 0 & 0 \\ 0 & 0 & -\frac{1}{C_{f1}} & 0 \\ 0 & 0 & 0 & -\frac{1}{C_{f1}} \end{bmatrix} \begin{bmatrix} V_{c1d} \\ V_{c1q} \\ I_{x1d} \\ I_{x1q} \end{bmatrix}$$

$$\begin{bmatrix} \dot{I}_{c2d} \\ \dot{I}_{c2q} \\ \dot{V}_{t2d} \\ \dot{V}_{t2q} \end{bmatrix} = \begin{bmatrix} -\frac{R_{c2}}{L_{c2}} & \omega_{sys} & -\frac{1}{L_{c2}} & 0 \\ -\omega_{sys} & -\frac{R_{c2}}{L_{c2}} & 0 & -\frac{1}{L_{c2}} \\ \frac{1}{C_{f2}} & 0 & 0 & \omega_{sys} \\ 0 & \frac{1}{C_{f2}} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{c2d} \\ I_{c2q} \\ V_{t2d} \\ V_{t2q} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_{c2}} & 0 & 0 & 0 \\ 0 & \frac{1}{L_{c2}} & 0 & 0 \\ 0 & 0 & -\frac{1}{C_{f2}} & 0 \\ 0 & 0 & 0 & -\frac{1}{C_{f2}} \end{bmatrix} \begin{bmatrix} V_{c2d} \\ V_{c2q} \\ I_{x2d} \\ I_{x2q} \end{bmatrix}$$

$$\begin{bmatrix} \dot{I}_{xd} \\ \dot{I}_{xq} \\ \dot{I}_{cd} \\ \dot{I}_{cq} \\ \dot{V}_{td} \\ \dot{V}_{tq} \end{bmatrix} = \begin{bmatrix} -\frac{1}{L_x} & 0 & -\frac{R_x}{L_x} & \omega_{sys} & 0 & 0 & \frac{1}{L_x} & 0 \\ 0 & -\frac{1}{L_x} & -\omega_{sys} & -\frac{R_x}{L_x} & 0 & 0 & 0 & \frac{1}{L_x} \\ 0 & 0 & 0 & 0 & -\frac{R_c}{L_c} & \omega_{sys} & -\frac{1}{L_c} & 0 \\ 0 & 0 & 0 & 0 & -\omega_{sys} & -\frac{R_c}{L_c} & 0 & -\frac{1}{L_c} \\ 0 & 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & 0 & 0 & \omega_{sys} \\ 0 & 0 & 0 & -\frac{1}{C_f} & 0 & \frac{1}{C_f} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{xd} \\ I_{xq} \\ I_{cd} \\ I_{cq} \\ V_{td} \\ V_{tq} \end{bmatrix} + \frac{1}{L_c} \begin{bmatrix} 0 \\ 0 \\ V_{cd} \\ V_{cq} \\ 0 \\ 0 \end{bmatrix}$$

A.5 Passive Network of GFM in Zone 4



$$\begin{bmatrix} \dot{I}_{c1d} \\ \dot{I}_{c1q} \\ \dot{V}_{t1d} \\ \dot{V}_{t1q} \end{bmatrix} = \begin{bmatrix} -\frac{R_{c1}}{L_{c1}} & \omega_{sys} & -\frac{1}{L_{c1}} & 0 \\ -\omega_{sys} & -\frac{R_{c1}}{L_{c1}} & 0 & -\frac{1}{L_{c1}} \\ \frac{1}{C_{f1}} & 0 & 0 & \omega_{sys} \\ 0 & \frac{1}{C_{f1}} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{c1d} \\ I_{c1q} \\ V_{t1d} \\ V_{t1q} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_{c1}} & 0 & 0 & 0 \\ 0 & \frac{1}{L_{c1}} & 0 & 0 \\ 0 & 0 & -\frac{1}{C_{f1}} & 0 \\ 0 & 0 & 0 & -\frac{1}{C_{f1}} \end{bmatrix} \begin{bmatrix} V_{c1d} \\ V_{c1q} \\ I_{x1d} \\ I_{x1q} \end{bmatrix}$$

$$\begin{bmatrix} \dot{I}_{c2d} \\ \dot{I}_{c2q} \\ \dot{V}_{t2d} \\ \dot{V}_{t2q} \end{bmatrix} = \begin{bmatrix} -\frac{R_{c2}}{L_{c2}} & \omega_{sys} & -\frac{1}{L_{c2}} & 0 \\ -\omega_{sys} & -\frac{R_{c2}}{L_{c2}} & 0 & -\frac{1}{L_{c2}} \\ \frac{1}{C_{f2}} & 0 & 0 & \omega_{sys} \\ 0 & \frac{1}{C_{f2}} & -\omega_{sys} & 0 \end{bmatrix} \begin{bmatrix} I_{c2d} \\ I_{c2q} \\ V_{t2d} \\ V_{t2q} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_{c2}} & 0 & 0 & 0 \\ 0 & \frac{1}{L_{c2}} & 0 & 0 \\ 0 & 0 & -\frac{1}{C_{f2}} & 0 \\ 0 & 0 & 0 & -\frac{1}{C_{f2}} \end{bmatrix} \begin{bmatrix} V_{c2d} \\ V_{c2q} \\ I_{x2d} \\ I_{x2q} \end{bmatrix}$$

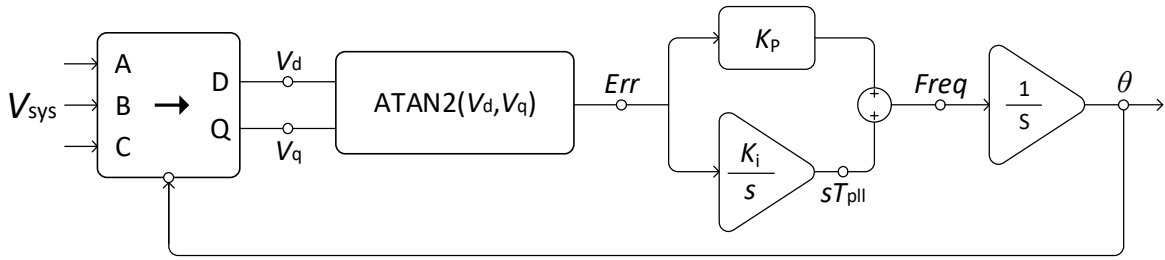
$$\begin{bmatrix} \dot{I}_{x1d} \\ \dot{I}_{x1q} \\ \dot{I}_{x2d} \\ \dot{I}_{x2q} \\ \dot{V}_{gd} \\ \dot{V}_{gq} \end{bmatrix} = \begin{bmatrix} -\frac{R_{x1}}{L_{x1}} & \omega_{sys} & 0 & 0 & -\frac{1}{L_{x1}} & 0 \\ -\omega_{sys} & -\frac{R_{x1}}{L_{x1}} & 0 & 0 & 0 & -\frac{1}{L_{x1}} \\ 0 & 0 & -\frac{R_{x2}}{L_{x2}} & \omega_{sys} & -\frac{1}{L_{x2}} & 0 \\ 0 & 0 & -\omega_{sys} & -\frac{R_{x2}}{L_{x2}} & 0 & -\frac{1}{L_{x2}} \\ \frac{1}{C_g} & 0 & \frac{1}{C_g} & 0 & -\frac{1}{R_g C_g} & \omega_{sys} \\ 0 & \frac{1}{C_g} & 0 & \frac{1}{C_g} & -\omega_{sys} & -\frac{1}{R_g C_g} \end{bmatrix} \begin{bmatrix} I_{x1d} \\ I_{x1q} \\ I_{x2d} \\ I_{x2q} \\ V_{gd} \\ V_{gq} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_{x1}} & 0 & 0 & 0 & 0 & 0 \\ 0 & \frac{1}{L_{x1}} & 0 & 0 & 0 & 0 \\ 0 & 0 & \frac{1}{L_{x2}} & 0 & 0 & 0 \\ 0 & 0 & 0 & \frac{1}{L_{x2}} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} V_{t1d} \\ V_{t1q} \\ V_{t2d} \\ V_{t2q} \\ 0 \\ 0 \end{bmatrix}$$

A.6 First-order lag Filter

$$\leftarrow \frac{1}{1 + sT}$$

$$\begin{bmatrix} \dot{V}_{df} \\ \dot{V}_{qf} \end{bmatrix} = \begin{bmatrix} -\frac{1}{T} & \omega \\ \omega & -\frac{1}{T} \end{bmatrix} \begin{bmatrix} V_{df} \\ V_{qf} \end{bmatrix} + \frac{1}{T} \begin{bmatrix} V_d \\ V_q \end{bmatrix}$$

A.7 Phase-locked Loop

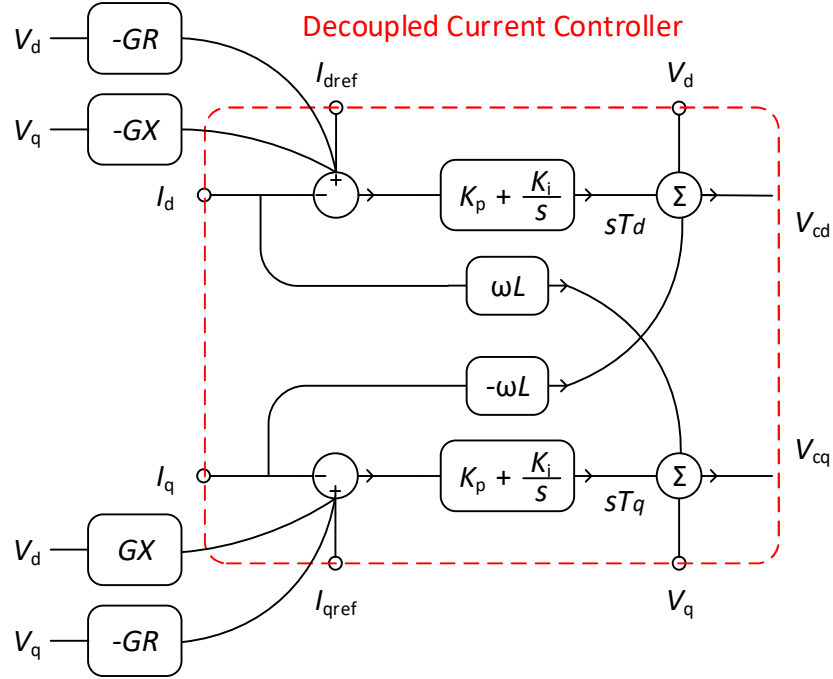


$$ATAN2(\Delta V_d, \Delta V_q) = \frac{V_{q0}}{V_{q0}^2 + V_{d0}^2} \Delta V_d - \frac{V_{d0}}{V_{q0}^2 + V_{d0}^2} \Delta V_q$$

$$V_{q0} = 0; V_{d0} = \sqrt{V_{qsys0}^2 + V_{dsys0}^2} = |V_{sys0}|$$

$$\begin{bmatrix} \Delta s\dot{T}_{pll} \\ \Delta \dot{\theta} \end{bmatrix} = \begin{bmatrix} 0 & -K_i \\ 1 & -K_p \end{bmatrix} \begin{bmatrix} \Delta sT_{pll} \\ \Delta \theta \end{bmatrix} + \begin{bmatrix} \frac{-K_i}{|V_{sys0}|} \sin(\theta_0) & \frac{K_i}{|V_{sys0}|} \cos(\theta_0) \\ \frac{-K_p}{|V_{sys0}|} \sin(\theta_0) & \frac{K_p}{|V_{sys0}|} \cos(\theta_0) \end{bmatrix} \begin{bmatrix} \Delta V_{dsys} \\ \Delta V_{qsys} \end{bmatrix}$$

A.8 Decoupled Current Controller (Considering VA)



$$\begin{bmatrix} s\dot{T}_d \\ s\dot{T}_q \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} sT_d \\ sT_q \end{bmatrix} - K_i \begin{bmatrix} I_d \\ I_q \end{bmatrix} + K_i \begin{bmatrix} I_{dref} \\ I_{qref} \end{bmatrix} + K_i G_R \begin{bmatrix} -V_d \\ -V_q \end{bmatrix} + K_i G_X \begin{bmatrix} -V_q \\ V_d \end{bmatrix}$$

$$\begin{bmatrix} s\dot{T}_d \\ s\dot{T}_q \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} sT_d \\ sT_q \end{bmatrix} - K_i \begin{bmatrix} I_d \\ I_q \end{bmatrix} + K_i \begin{bmatrix} I_{dref} \\ I_{qref} \end{bmatrix} + K_i \begin{bmatrix} -G_R & -G_X \\ G_X & -G_R \end{bmatrix} \begin{bmatrix} V_d \\ V_q \end{bmatrix}$$

$$\begin{bmatrix} V_{cd} \\ V_{cq} \end{bmatrix} = \begin{bmatrix} V_d \\ V_q \end{bmatrix} + \begin{bmatrix} sT_d \\ sT_q \end{bmatrix} + \omega L \begin{bmatrix} -I_q \\ I_d \end{bmatrix} - K_p \begin{bmatrix} I_d \\ I_q \end{bmatrix} + K_p \begin{bmatrix} I_{dref} \\ I_{qref} \end{bmatrix} - K_p G_R \begin{bmatrix} V_d \\ V_q \end{bmatrix} + K_p G_X \begin{bmatrix} -V_q \\ V_d \end{bmatrix}$$

$$\begin{aligned}
\begin{bmatrix} \Delta \dot{s}T_d \\ \Delta \dot{s}T_q \end{bmatrix} &= \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} \Delta sT_d \\ \Delta sT_q \end{bmatrix} \\
&+ K_i \left(-\frac{1}{I_{base}} \begin{bmatrix} \cos(\theta_0) & \sin(\theta_0) \\ -\sin(\theta_0) & \cos(\theta_0) \end{bmatrix} \begin{bmatrix} \Delta I_{dsys} \\ \Delta I_{qsys} \end{bmatrix} - \frac{1}{I_{base}} \begin{bmatrix} -\sin(\theta_0) & \cos(\theta_0) \\ -\cos(\theta_0) & -\sin(\theta_0) \end{bmatrix} \begin{bmatrix} I_{dsys0} \\ I_{qsys0} \end{bmatrix} \right) [\Delta \theta] \\
&+ \begin{bmatrix} \Delta I_{dref} \\ \Delta I_{qref} \end{bmatrix} + \frac{1}{V_{base}} \begin{bmatrix} -G_R & -G_X \\ G_X & -G_R \end{bmatrix} \begin{bmatrix} \cos(\theta_0) & \sin(\theta_0) \\ -\sin(\theta_0) & \cos(\theta_0) \end{bmatrix} \begin{bmatrix} \Delta V_d \\ \Delta V_q \end{bmatrix} \\
&+ \frac{1}{V_{base}} \begin{bmatrix} -G_R & -G_X \\ G_X & -G_R \end{bmatrix} \begin{bmatrix} -\sin(\theta_0) & \cos(\theta_0) \\ -\cos(\theta_0) & -\sin(\theta_0) \end{bmatrix} \begin{bmatrix} \Delta V_{dys0} \\ \Delta V_{qsys0} \end{bmatrix} [\Delta \theta]
\end{aligned}$$

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