

Robust Dynamic Resilient Power Grids Enabled By Modern Control Framework

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“Let us rise up and be thankful, for if we didn’t learn a lot today, at least we learned a little, and if we didn’t learn a little, at least we didn’t get sick, and if we got sick, at least we didn’t die; so, let us all be thankful.”.....BUDDHA

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Dedication

To ...

my Ma, Baba, Dada, Boudi and my little Tora

... for their love.

Abstract

The inexorable effects of climate change are disrupting weather patterns with disastrous consequences such as hurricanes, wildfires, ice storms, flooding, etc. The growing frequency, duration, and intensity of such catastrophes put stress on the traditional aging electrical power infrastructure and cause prolonged power outages worldwide. In many well-publicized reports of remediation plans for power outages, microgrids have been envisioned as a critical resource for supplying facilities with resilient and reliable power during such emergencies and outages. The main reason for fostering microgrids is the benefits it provides that include (1) the opportunity to deploy more zero-carbon emission electricity sources, thereby reducing greenhouse gas emissions, (2) enhanced grid resiliency to extreme weather attacks by flexibly engaging/disengaging to/from the main grid, (3) utilization of on-site distributed energy resources (DERs) for reliable and uninterrupted power supply to the local consumers. Worldwide roll-out projects and implementations indicate that from the stage of a niche novelty, today microgrid is emerging as a mainstream solution.

New challenges have emerged from this transformation in the way of making and consuming power using microgrids. Increasing installation of renewable DERs via inverter-based-resources (IBRs) in microgrids has resulted in a significant amount of decline in the inertia of the system by the displacement of high inertia-based synchronous machines. Penetration of intermittent and unpredictable renewable to such low inertia-based microgrid systems lead to undesirable fluctuations in the frequency and the voltage. Besides, there is a renewed emphasis on today's microgrids that provide operational flexibility leveraged by the control schemes for IBRs which introduces the plug-n-play capability of IBRs. As a result, today's microgrids experience a significant amount of uncertainties from the generation side due to intermittent renewable integration, from the electrical network side due to plug-n-play of IBRs, and from the load side due to constantly varying demand. It deteriorates the robustness in performance of the microgrid by influencing the control of IBRs, leading to instability in the *worst cases*.

Another topical area of development for microgrid is the enhancement of its resurrecative capability to restore power rapidly and seamlessly after outages. The improvement of the flexibility in the way of connecting/disconnecting to/from the main grid or other local microgrids, when needed, is the crucial remedial action that makes a microgrid

dynamic and resilient. The major challenge here is to develop a capability for the IBRs that enables a smooth (controlled transients and distortion in voltage waveform) and seamless (without any electrical interruption) transition of a microgrid from on-grid (grid-tied) to off-grid (islanded) conditions. Moreover, when the main grid returns to normalcy, the microgrid needs to be reconnected/resynchronized smoothly and seamlessly. Large uncontrolled transients in the output voltage and/or current of the IBRs are usually observed during these jumps in the mode of operations of microgrids, leading to power collapse in the *worst cases*. These transients are the major impediments that restrict today's microgrid from a smooth and seamless transition.

In this thesis, Chapter 2 contributes towards the development of a *robust* microgrid against various inevitable uncertainties. A μ -synthesis-based robust controllers for the IBRs in a microgrid are developed by quantifying some major sources of uncertainties using a generalized control framework. It includes the uncertainties caused by the equivalent impedance parameters at the point-of-connection of the IBRs and the equivalent loading parameters of the IBRs. The efficacy and viability of the robust controllers for IBRs are evaluated using an experimental case study involving an emulated 1- ϕ , 240V, 10kVA, 14-bus residential microgrid with two physical 2kVA IBRs. The developed robust controller for IBRs shows robust performance and dynamic response against large uncertainties of a microgrid.

Chapter 3 and Chapter 4 of this thesis contribute towards developing a *dynamic* microgrid that can engage/disengage to/from the main grid or the other microgrids very flexibly. In Chapter 3, a mode-dependant integrator-based droop controller is developed for the IBRs that facilitates a smooth and seamless on-grid to off-grid transition capability in a microgrid. In Chapter 4, a dynamic compensation-based active synchronization method is developed using droop controllers of the IBRs that enables a smooth and seamless off-grid to on-grid transition capability in a microgrid. Methods are evaluated using an emulated experimental case study involving a 3- ϕ , 480V, 500kVA, 55-bus microgrid with six IBRs and using a laboratory-scale experimental case study involving a 3- ϕ , 480-V, 270-kVA microgrid with two 3- ϕ , 480V, 125kVA IBRs. The developed controllers in the IBRs enable relatively fast and smooth connection/disconnection capability of a microgrid with less transients.

Chapter 5 introduces a large-scale demonstration only to showcase that the seamless transition capabilities of Chapter 3 is the key capability for a *resilient* microgrid to support disaster response efforts and facilitate recovery. The demonstration is conducted on a laboratory-scale 3- ϕ , 480V, 750kVA microgrid with critical loads of 550kW,

215kVAR, four 3- ϕ , 125kVA, 4-quadrant IBRs and, one 3- ϕ , 135kW diesel genset.

This thesis further investigates another emerging domain that deals with advanced and modern validation techniques for any engineering innovations. Any technological innovation is required to be validated either via offline simulation environments or via experimental demonstration before being employed in the real world. For instance, a newly developed controller for IBRs requires to be validated by conducting diverse tests before prototyping and real-world implementation. In this context, power hardware-in-the-loop (PHIL)-based digital real-time simulation (DRTS) is considered to be an attractive platform for conducting diverse tests on a laboratory scale, providing results that closely resemble real-world scenarios, all while avoiding the high costs and risks associated with experimentation on a full-scale power system. The hardware system, also called hardware-under-test (HUT), is usually interfaced with a software system that is simulated in real-time inside a real time simulator. The interfacing compensator/controller is the backbone of the PHIL-based DRTS that manages the closed-loop interaction between the hardware system (e.g., single/group of IBRs) and the software system (e.g., realistic microgrid model). Considerable challenges emerge in the process of synthesizing the PHIL interfacing controller due to the inherent uncertainties present in both the software and the hardware system. Interfacing controllers, designed without addressing these uncertainties *a priori*, often suffer from a lack of robustness in the performance. This leads to inaccurate results from the testing of the HUT using the PHIL-based DRTS that don't resemble with the expected results of the same HUT in real-world operations. Therefore, the interface between the software and the hardware system must be carefully synthesized to ensure robust and accurate PHIL-based DRTS.

Chapter 6 of this thesis makes an important contribution in addressing these challenges by synthesizing a μ -synthesized robust PHIL interface controller between the hardware and software components of a PHIL-based DRTS while employing a modern control perspective for managing inherent uncertainties. It provides a methodology that quantifies model uncertainty especially caused by the point-of-connection in the emulated power network of the software system and the model uncertainty stemming from the varying nature of the HUT. The designed interfacing controller is implemented to interface a software system based on a 110V, 1MW, 225-bus residential sub-network of the University of Minnesota and suburb Minneapolis and a hardware system consisting of two 1- ϕ 1.67kVA IBRs, one linear and nonlinear load of 1.8 kVA each. It is observed that the robust PHIL interface shows improved accuracy, and tracking performance under varying conditions of software and hardware system.

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List of Symbols

L_f	- Inductance (in H) of the inverter-side filter inductor
R_f	- Parasitic resistance (in Ω) of the inverter-side filter inductor
L_g	- Inductance (in H) of the grid-side filter inductor
R_g	- Parasitic resistance (in Ω) of the grid-side filter inductor
C_f	- Capacitance (in F) of the filter of the inverter
L_{Th}	- Thevenin equivalent inductance (in H) of the network
R_{Th}	- Thevenin equivalent resistance (in Ω) of the network
$\mathcal{G}_{inv}^{GFL}(s)$	- Open-loop internal transfer function of the GFL inverter
$\mathcal{G}_{Th}^{GFL}(s)$	- Open-loop external transfer function of the GFL inverter
L_{Load}	- Equivalent loading inductance (in H) of the network
R_{Load}	- Equivalent loading resistance (in Ω) of the network
$\mathcal{G}_{inv}^{GFM}(s)$	- Open-loop internal transfer function of the GFM inverter
$\mathcal{G}_{Load}^{GFM}(s)$	- Open-loop external transfer function of the GFM inverter
L_{Th}^{Nom}	- Nominal Thevenin equivalent inductance (in H) of the network
R_{Th}^{Nom}	- Nominal Thevenin equivalent resistance (in Ω) of the network
S_{rated}	- Rated apparent power of the inverter (in kVA)
P_{rated}	- Rated active power of the inverter (in kW)
Q_{rated}	- Rated reactive power of the inverter (in kVAr)
L_{Load}^{Nom}	- Nominal equivalent loading inductance (in H) of the network
R_{Load}^{Nom}	- Nominal equivalent loading resistance (in Ω) of the network
δ_L, δ_R	- Real parametric uncertainty related to inductor and resistor
Δ	- Structured uncertainty block
$\mathcal{I}_{gs}$	- Status of the feeder switch
$v_{o,i}^{dq}$	- dq component of output voltage of i^{th} inverter
$i_{o,i}^{dq}$	- dq component of output current of i^{th} inverter

$i_{L,i}^{dq}$	- dq component of inductor current of i^{th} inverter
p_i, q_i	- Instantaneous active, reactive power of i^{th} inverter
P_i, Q_i	- Averaged active, reactive power of i^{th} inverter
$\tau_{S,i}$	- Time constant of low pass filter of i^{th} inverter
n_i	- Droop coefficient of P - f droop law of i^{th} inverter
m_i	- Droop coefficient of Q - V droop law of i^{th} inverter
$m_{\text{int},i}$	- Integral constant of Q - V droop law of i^{th} inverter
ψ_i^Q	- The integral term in Q - V droop law of i^{th} inverter
$P_{\text{ref},i}$	- Active power set point of P - f droop equation
$Q_{\text{ref},i}$	- Reactive power set point of Q - V droop equation
ω_{nom}	- Nominal frequency (in rad/s) of the system
V_{nom}	- Nominal voltage (in volt) of the system
$\omega_{r,i}$	- Droop frequency (in rad/s) of the i^{th} inverter
$V_{r,i}$	- Droop voltage (in volt) of the i^{th} inverter
v_{PCC}^{dq}	- dq-axis component of voltage at PCC
$\theta_{r,i}$	- Phase angle of i^{th} inverter w.r.t. PCC
$\omega_{\text{grid}}, V_g$	- Grid frequency (rad/sec) and voltage (volt)
θ_{grid}	- Phase angle of the grid w.r.t. PCC
R_L, L_L	- Resistive, inductive parameter of the Z-type load
$\omega_{\text{adj}}, v_{\text{adj}}$	- Adjustments in P - f and Q - V droop laws
$v_G, v_{\mu G}$	- Voltage RMS of the grid and the microgrid
$\omega_G, \omega_{\mu G}$	- Frequency of the grid and the microgrid
$\theta_G, \theta_{\mu G}$	- Phase angle of the grid and the microgrid
k_p^ω, k_i^ω	- Phase PI controller Parameters
k_p^v, k_i^v	- Voltage PI controller Parameters
$\mathcal{G}_{\text{PA}}$	- Transfer function model of the power amplifier
$\mathcal{Z}_{\text{PA}}$	- Series impedance model of the power amplifier
$\mathcal{G}_{\text{AI}}^v$	- Transfer function of the analog input current module
$\mathcal{G}_{\text{AI}}^i$	- Transfer function of the analog input voltage module
$\mathcal{G}_{\text{AO}}^v$	- Transfer function of the analog output voltage module
$\mathcal{G}_{\text{VS}}$	- Transfer function of the voltage sensor
$\mathcal{G}_{\text{CS}}$	- Transfer function of the current sensor

List of Abbreviations

DER	- Distributed Energy Resources
PV	- Photo Voltaic
IBR	- Inverter Based Resources
GFL	- Grid-following
GFM	- Grid-forming
CI	- Critical Infrastructure
DRTS	- Digital Real Time Simulation
SIL	- Software-in-the-loop
HIL	- Hardware-in-the-loop
HUT	- Hardware Under Test
CHIL	- Controller hardware-in-the-loop
PHIL	- Power hardware-in-the-loop
EPSS	- Emergency Power Supply System
SEPSS	- Stored-energy Emergency Power Supply System
SCR	- Short Circuit Ratio
MIMO	- Multi-input-multi-output
RMS	- Root Mean Square
RTS	- Real Time Simulator
THD	- Total Harmonic Distortion
TDD	- Total Demand Distortion
PnP	- Plug and Play
PCC	- Point of Common Coupling
PoC	- Point of Connection

PWM	- Pulse Width Modulation
SOC	- State of Charge
EMT	- Electro-magnetic Transient
UV	- Under Voltage
OV	- Over Voltage
UF	- Under Frequency
OF	- Over Frequency
SRF	- Synchronous Reference Frame
PLL	- Phase Locked Loop
PI	- Proportional and Integral
PR	- Proportional and Resonant
UPS	- Uninterrupted Power Supply
NLM	- Net Load Management
NLU	- Net Load Unit
HoV	- Horizon of Viability
HVAC	- Heating, Ventilation and Air Conditioning
SOE	- State of Energy
UDP	- Universal Datagram Protocol
ITM	- Ideal Transformer Model
IA	- Interface Algorithm
PCD	- Partial Circuit Duplication
FCF	- Feedback Current Filtering
DIM	- Damping Impedance
TFA	- Time-variant First-order Approximation
TLM	- Transmission Line Model
LTI	- Linear Time Invariant
LFT	- Linear Fractional Transformation

Chapter 1

Introduction

1.1 Motivation for Robust Dynamic Resilient Microgrid

The traditional power grid provides reliable and sustainable power most of the time with the help of large centralized coal-based generation units. But when natural disasters or security breaches threaten, the ensuing blackouts can be catastrophic and costly. For instance, between 2000 and 2021, about 83% of reported major outages (over 1500 outages) in the North America and its territories were attributed to weather-related events such as winter weather, tropical cyclones, that increased by roughly 78% during 2011-2021, compared to 2000-2010 [1]. Recently, Puerto Rico has suffered an island-wide power outage nearly seven months after Hurricane Maria destroyed much of the island's infrastructure and power grid and it took roughly 11 months, for the island to restore power to all of the customers [2]. To circumvent the issues of vulnerability to large-scale outages, more reliable and resilient ways of generation, transmission and distribution of power become the need of the hour.

1.1.1 Microgrid: The Grid of the Future

With advancements in information and communication technologies in unison with *zero-carbon*-based distributed energy resources (DERs), the future promises dramatic transformations in the way people make and consume reliable, resilient, clean and economical viable energy. Many companies and utilities are turning to the concept of 'microgrids', a small/medium-scale, self-sustaining power networks unburdened by ties to the decentralized DERs scattered throughout the network. Figure 1.1 shows the schematic

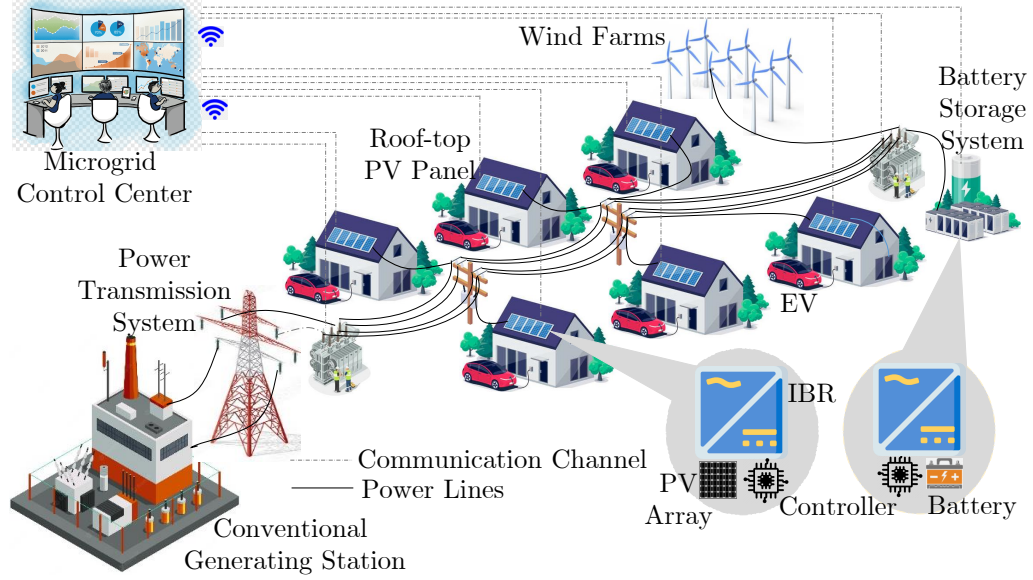


Figure 1.1: A schematic representation of a generic microgrid.

representation of a generic microgrid that consists of multiple loads such as consumer-houses, electric vehicle charging-discharging stations, multiple renewable DERs such as roof-top PV panels, local wind farms. Along with the DERs, the main power grid also participates as a source of electrical power when the microgrid is operating in *grid-tied* mode. In addition to that, it also contains battery-storage systems in case of emergency situations when the entire microgrid network is disconnected from the main power transmission system, a capability called *islanded* mode of operation. Microgrids have long been viewed as a critical resource for supplying facilities with resilient power during emergencies and outages. In many well-publicized power outages, the remediation plans have included the use of microgrids with renewable and traditional DERs. The main reason for fostering microgrids as the solution during emergencies and outages is the benefits that a microgrid provides to the environment, to the utility operators, and to the consumers. These benefits are:

- it offers the opportunity to deploy more *zero-carbon emission* electricity sources, reducing greenhouse gas emissions,
- it offers enhanced grid resiliency to extreme weather/cyber attacks by flexibly engaging/disengaging to/from the main grid,

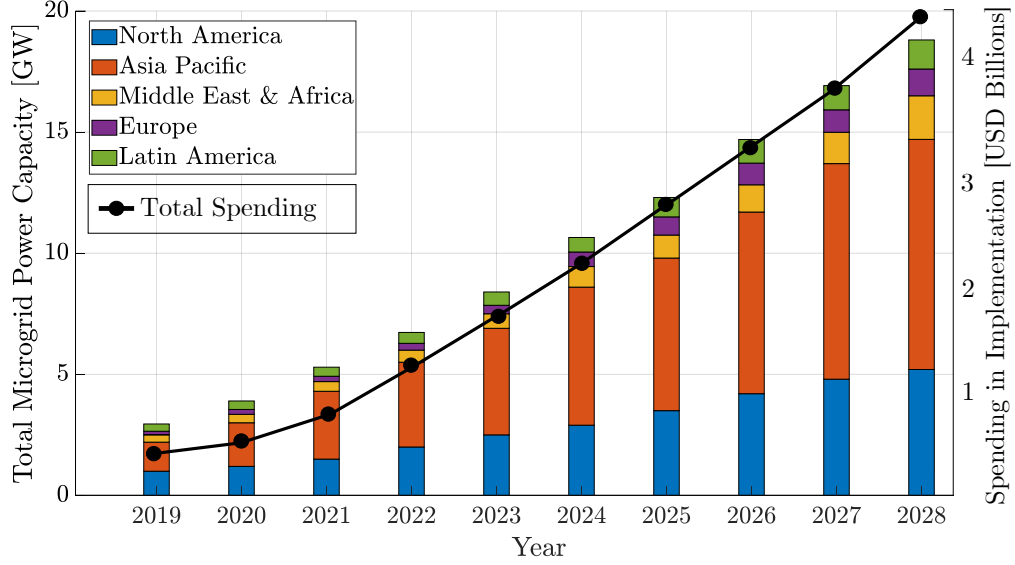


Figure 1.2: Year-wise annual total microgrid power capacity and implementation spending in various world region.

- it offers utilization of on-site energy resources for reliable and uninterrupted power supply to the local consumers.

1.1.2 Worldwide Microgrid Projects

Recent studies by various energy experts find and forecast cumulative microgrid investments in every region of the world and envision microgrid as an essential building blocks of the future electricity delivery system to support resilience, de-carbonization, and affordability [3, 4]. Figure 1.2 shows year-wise employment of annual total microgrid power capacity and implementation spending in various world region [5]. It shows that Asia Pacific and North America region are the two market leaders and Middle East, Africa, and Latin America represent the fastest growing market while Europe is lagging behind. For instance, capacity of microgrids has grown by almost 11% in the past four years and over 650 microgrids (each of them is more than 500 kW capacity) in the North America are concentrated in states like Alaska, California, Georgia, Maryland, New York, Oklahoma, and Texas [6]. Puerto Rico Electric Power Authority is re-configuring the island's grid to be made up of eight large-scale microgrids of roughly 500 MW each [7]. The vast majority of microgrids deployed in Europe are concentrated

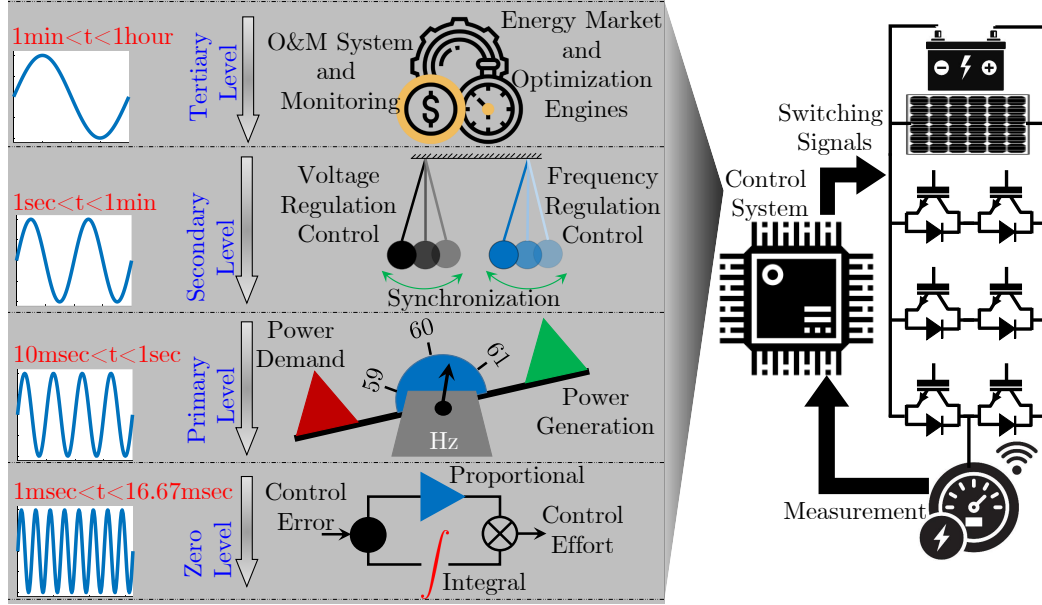


Figure 1.3: Hierarchical control levels of microgrids.

on the Mediterranean islands, the Canary Islands off the coast of Spain [8]. India has installed solar microgrids providing around 2 MW of electricity so far. The partnership between TATA Power and Rockefeller Foundation proposed to set up 10,000 microgrids in India by 2026 [9]. All these numbers tell us that from the stage of a niche novelty, today microgrid is emerging as a mainstream solution worldwide.

1.1.3 How Does A Microgrid Operate?

With the advent of power electronics and its associated digital controls, inverter-based resources (IBRs) become the backbone of any microgrids. Due to their flexible control, today, it is possible to integrate large number of renewable DERs such as photovoltaics, wind generators, and battery-storage-based DERs such as fuel cells, super-capacitors. Based on the nature and the expected functionality of the DER, these IBRs are usually operated either in grid-following (GFL) or grid-forming (GFM) mode, primarily driven by the employed control schemes. For instance, renewable DERs are interfaced usually via IBRs that are operated in GFL mode to inject maximum available power to the microgrid, whereas, IBRs are operated in GFM mode to integrate battery-storage-based DERs which are responsible to generate and maintain stable voltage and frequency of

the microgrid, when disengaged from the main grid. As an entity, the control system for an entire microgrid has multiple design and operational objectives to satisfy. For instance, regulation of voltage and frequency of the microgrid, control of active and reactive power dispatches of various IBRs, restoration of system stability following a disturbance, transition control of *grid-tied* and *islanded* mode of operation, power flow control between multiple microgrids while maintaining economic dispatch of assets, are few of these objectives. To comply with the performance criteria, microgrids usually follows a hierarchical control structure by adapting ANSI/ISA-95-based international standard [10]. This hierarchical control structure consists of four levels, where each level expects a supervisory control from higher-level and provides a supervisory control to lower-level. To ensure that the command and reference signals from one level to the lower levels will have low impact in the overall stability, the bandwidth is decreased with an increase in the control level, as shown in Figure 1.3. These levels are:

1. *Zero-level* Control: It consists of the voltage and current regulation control of each IBR-based DERs in the microgrid, e.g., *inner-current* controller for GFL IBRs and *outer-voltage-inner-current* controller for GFM IBRs. The time-scale lies usually between *sub-power-cycle* to few *power-cycles* period.
2. *Primary-level* Control: It consists of the voltage and frequency regulation control of the microgrid system and active and reactive power control of multiple IBRs, e.g., P-Q set-point generation control for GFL IBRs, droop control for GFM IBRs. The time-scale lies usually between few *milliseconds* to few *seconds*.
3. *Secondary-level* Control: It consists of the compensation control for voltage and frequency of the microgrid caused by the *primary-level*, synchronization control for engage/disengage operation of the microgrid etc. The time-scale lies usually between few *seconds* to few *minutes*.
4. *Tertiary-level* Control: It consists of the energy-production level control, economic layer control and power flow control between microgrids and the grid/other microgrids. The time-scale lies usually between few *minutes* to few *hours*.

1.1.4 Challenges That Microgrids Face Today

Increasing installation of renewable-based DERs has resulted in the displacement of high inertia-based synchronous machines by low inertia-based IBRs in today's microgrids. This causes a significant amount of decline in the inertia of the system, which

was naturally provided by the mechanical systems of synchronous machines. Another difficulty in integrating renewable-based DERs is their intermittent and unpredictable nature in power generation. As a result, interconnection of such renewable DERs to low inertia-based microgrid systems leads to undesirable fluctuations in the frequency and the voltage of the microgrid. Besides, there is a renewed emphasis on today's microgrids that provide operational flexibility leveraged by the control schemes for IBRs which introduces the plug-n-play capability of IBRs (i.e., IBRs can operate either in GFL mode or in GFM mode depending on the selection of control scheme). As a result, today's microgrids experience a significant amount of uncertainties from the generation side due to intermittent renewable integration, from the electrical network side due to plug-n-play of IBRs, and from the load side due to constantly varying demand. It deteriorates the robustness in performance of the microgrid by influencing the control in the *Zero-level* control layer, leading to instability in the *worst cases*.

Another topical area of development for today's microgrids is enhancement of its disaster response and recovery. The primary power grid is often severely impacted during natural disasters such as hurricanes, earthquakes, and floods. These disturbances lead to prolonged power outages and significant damage to essential assets of society, also called as critical infrastructures (CIs), that include medical centers and security service centers. Disruption of power to such CIs result in a debilitating impact on physical and economic security, public health and safety. Rapid and seamless recovery of power to restore CIs online, possibly after a power blackout caused by weather/climate disasters, is a crucial need arising in scenarios that are increasingly becoming more frequent. In these situations, microgrids can provide a reliable and flexible source of power that can help to support disaster response efforts and facilitate recovery. The enhancement of the flexibility in the way of connecting/disconnecting to/from the main grid or other local microgrids, when needed, is the crucial remedial action that is required to be developed and implemented in today's microgrids. The major challenge here is to develop a capability of the microgrid, especially in the *Primary-level* and *Secondary-level* control, that enables a smooth (controlled transients and distortion in voltage waveform) and seamless (without any electrical interruption) transition from on-grid (grid-tied) to off-grid (islanded) condition. Moreover, when the main grid returns to normalcy, the microgrid needs to be reconnected smoothly and seamlessly. Large uncontrolled transients in the output voltage and/or current of the IBRs are usually observed during these jumps in the mode of operations of microgrids, leading to power collapse in the *worst cases*. These transients are the major impediments that restrict today's microgrid

from a smooth and seamless transition.

1.2 Validation Before Implementation

Any technological innovation is required to be validated either via off-line simulation environments or via experimental demonstration of prototypes before employing in real world. For instance, a newly developed controller for IBRs is required to be validated by conducting diverse tests via either simulation or laboratory-scale experiments. This section introduces an attractive platform for evaluating any novel technological solution.

1.2.1 Digital Real-Time Simulation: A Modern Approach

Enabled by advances in computational technology, digital real-time simulation (DRTS) allows for conducting diverse tests on a laboratory scale, providing results that closely resemble real-world scenarios, all while avoiding the high costs and risks associated with experimentation on a full-scale power system. A DRTS applied to the domain of power systems can be broadly classified into two categories:

1. **Software-in-the-loop (SIL) DRTS:** A SIL-based DRTS requires the entire power system (including electrical network, associated control, protection) to be modeled inside the simulator and does not involve external interfacing or inputs/outputs (I/Os). A schematic of SIL-based DRTS is shown in Figure 1.4(a).
2. **Hardware-in-the-loop (HIL) DRTS:** A HIL DRTS refers to the condition where parts of the DRTS have been replaced with actual physical components. The HIL mode of the DRTS proceeds with the hardware-under-test (HUT) connected through input/output interfaces, e.g., filters, digital-to-analog (D/A) and analog-to-digital (A/D) converters, signal conditioners, etc.

A HIL DRTS is further classified into two categories:

1. **Controller hardware-in-the-loop (CHIL) -based DRTS:** It involves a real micro-controller-based hardware that interacts with the rest of the emulated system. CHIL-based DRTS is quite popular for rapid prototyping of newly designed/developed controller where the controller takes feedback signals from the real-time-simulator and processes them to produce the required output signals, which are then sent back to the the simulator, as shown in Figure 1.4(b).

2. **Power hardware-in-the-loop (PHIL) -based DRTS:** It involves a real power hardware such as power transformers, converters, and machines that interacts with the rest of the emulated system. In PHIL-based DRTS, part of the power system is internally simulated and the other part is the real hardware power apparatus connected externally, as shown in Figure 1.4(c). A power source or sink, as a part of interfacing blocks, is needed for this setup, which will either generate or absorb power. Reference signals are generated based on the solution of the virtual system inside the real-time simulator and are sent to the power amplifier that produces voltages or currents to be applied to the HUT. Feedback signals obtained from the voltage/current measurements of the HUT are appropriately scaled and brought back to the simulator to complete the simulation loop.

The popularity of DRTS are substantially increased by the advancements of the PHIL-based DRTS. Using PHIL-based DRTS, the closed-loop interaction between a HUT and a simulated power system can be evaluated under a variety of grid configurations and repeatable operational scenarios without the expense or risk of a field deployment. It serves as an efficient in-laboratory testing methodology for the power hardware under close-to-real conditions.

1.2.2 Challenges in PHIL-based DRTS

Interfacing compensator/controller is the backbone of the PHIL-based DRTS that manages the closed-loop interaction between the hardware system (e.g., single/group of IBRs) and the software system (e.g., realistic microgrid model). For the PHIL-based DRTS results to be representative of expected real world operations, this PHIL interface between the simulated system and the HUT must be carefully implemented to ensure stable and accurate closed-loop simulation. Considerable challenges emerge in the process of synthesizing the PHIL interfacing controller due to the inherent uncertainties present in both the software and the hardware system. Interfacing controllers, designed without addressing these uncertainties *a priori*, often suffer from lack of robustness in the performance. This leads to the inaccurate results from the testing of the HUT using the PHIL-based DRTS which don't resemble with the expected results of the same HUT in real-world operations. Therefore, the interface between the software and the hardware system must be carefully synthesized to ensure robust and accurate closed-loop PHIL-based DRTS.

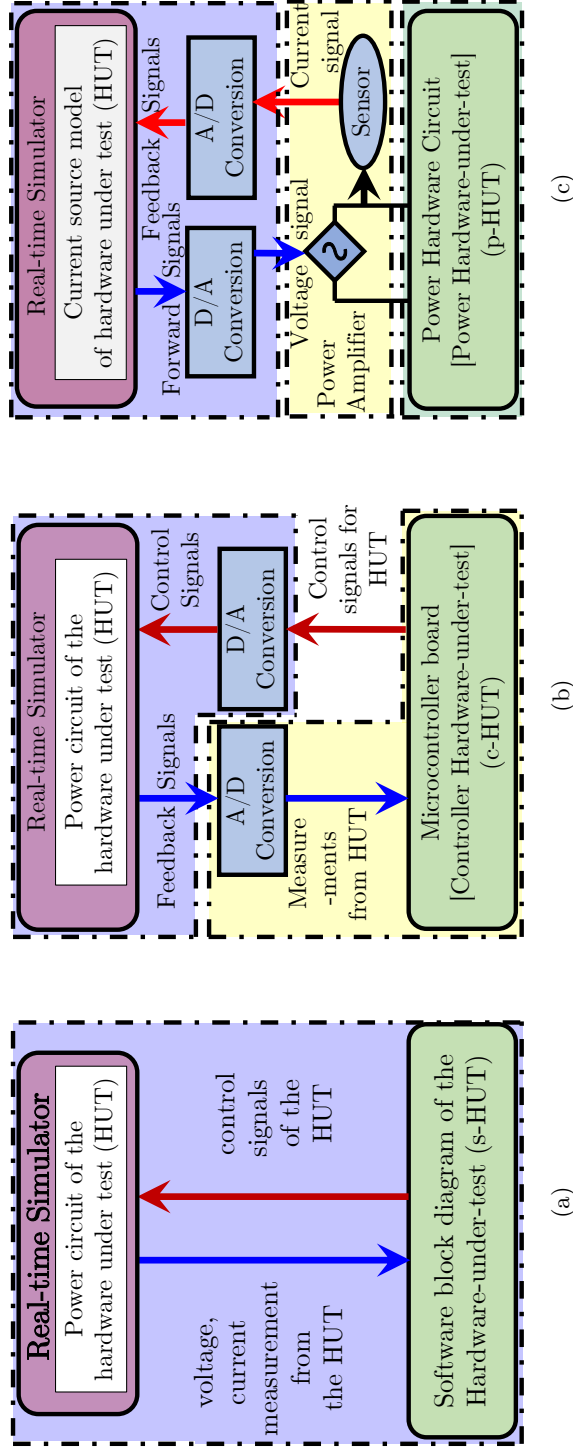


Figure 1.4: Digital real-time simulation configuration and interfacing blocks for (a) software-in-the-loop (SIL) platform, (b) controller hardware-in-the-loop platform (CHIL), and (c) power hardware-in-the-loop (PHIL).

1.3 Contributions and Overview of the Thesis

1.3.1 Contribution Towards Robust Dynamic Resilient Microgrid

Section 1.1.4 discusses the major challenges that microgrids experience today. Clearly, the need of the hour for today's microgrid is to enhance the robustness, dynamicity, and resiliency both in stability and in performance. This thesis contributes towards achieving these objectives by employing modern control approaches in the hierarchical control structure described in Section 1.1.3.

■ In this thesis, **Chapter 2** contributes towards the development of a *robust* microgrid against various inevitable uncertainties by developing a robust and optimal *Zero-level* control for microgrids. It introduces a generalized robust control framework for designing a single-loop current-controller for GFL inverters and a single-loop voltage-controller for GFM inverters. A review of existing controllers for both GFL and GFM inverters is presented to highlight the limitations of the state-of-the-art. A μ -synthesis-based robust control design methodology is developed to overcome the challenges due to uncertainties caused by varying grid impedance parameters and varying equivalent loading impedance for GFL and GFM inverters, respectively. Moreover, a generalized control framework is developed for both GFL and GFM inverter systems by leveraging the following two key points: i) the voltage-current duality in the plant dynamic model of GFL and GFM inverters, and ii) the generalized model of uncertainties by quantifying the grid impedance parameters and the equivalent loading parameters for GFL and GFM inverters respectively. The novel robust controller achieves enhanced reference tracking, disturbance rejection, and harmonic compensation capability with i) sufficient LCL resonance damping under large variations of grid impedance uncertainty for GFL inverters and ii) with enhanced dynamic response under large variations of equivalent loading uncertainty for GFM inverters. The efficacy and viability of the robust controllers for IBRs are evaluated using an experimental case study involving an emulated 1- ϕ , 240V, 10kVA, 14-bus residential microgrid with two physical 2kVA IBRs. The developed robust controller for IBRs shows robust performance and dynamic response against large uncertainties of a microgrid.

■ **Chapter 3** contribute towards developing a *dynamic* microgrid that can disengage from the main grid or the other microgrids very flexibly. It introduces a seamless transition capability in the *Primary-level* control for a microgrid using a single and unified mode-dependent droop controlled GFM inverters. By leveraging the information of the

mode of the operation of a microgrid, i.e. on-grid mode or off-grid mode, an additional integral-loop in the voltage-reactive power droop law enables a inverter to operate always in GFM mode irrespective of the availability of the grid. As a result, the control strategy achieves the following objectives: i) regulates the output active and reactive power by the GFM inverters to a desired value while operating in on-grid mode, ii) seamless transition and recovery of power injections into the load after grid failure by inverters that operates in grid-forming mode all the time, iii) regulates the voltage and frequency of the microgrid by the GFM inverters under the limits recommended by grid codes while maintaining autonomous power sharing capability during off-grid mode of operation of the microgrid. A framework for assessing the stability of the system and to guide the choice of parameters for controllers is developed using control-oriented modeling. A hardware experiment is conducted with two 3- ϕ , 480V, 125kVA grid-forming inverters, a 3- ϕ , 480V, 270kVA grid simulator, a physical grid switch, and a physical load bank.

■ **Chapter 4** contribute towards developing a *dynamic* microgrid that can engage to the main grid or the other microgrids very flexibly by developing a smooth synchronization methods in the *Secondary-level* control for microgrids. In this synchronization method, the synchronization controller compensates the phase angle and voltage magnitude difference between off-grid microgrid side and the incoming grid side to zero by adjusting the droop laws of all GFM inverters of the microgrid dynamically via secondary control layer of microgrid hierarchical control architecture. It is analytically shown that compensation on only phase angle and voltage magnitude is sufficient for synchronization as frequency is compensated by this method indirectly by the phase angle compensation. A systematic approach for tuning the proposed synchronization controllers are provided. The method is evaluated using an emulated experimental case study involving a 3- ϕ , 480V, 500kVA, 55-bus microgrid with six IBRs. The developed controllers in the IBRs enable relatively fast and smooth connection/disconnection capability of a microgrid with less transients.

■ **Chapter 5** demonstrates a case study where it is showcased that the seamless transition capabilities of Chapter 4 is the key capability for a *resilient* microgrid to support disaster response efforts and facilitate recovery. It introduces a comprehensive design and large-scale hardware demonstration of the emergency power supply system (EPSS) for uninterrupted operation of critical infrastructures (CIs) by employing novel techniques, such as 1) mode-dependent droop controlled GFM inverters for seamless transition capability; 2) fast-acting optimal net-load management engine [11, 12]

for efficient and optimal operations maintaining regulation and power quality limits; 3) optimization-based horizon of viability engine [13] for longer, sustained and viable operation in the aftermath of grid failure as recommended in standards. Two stage validations are conducted using defined test cases while considering various types of contingencies on a 3- ϕ , 60 Hz, 480 V, 450 kVA test system based on a realistic electrical network of a commercial scale medical center. The demonstration is conducted on a laboratory-scale 3- ϕ , 480V, 750kVA microgrid with critical loads of 550kW, 215kVar, four 3- ϕ , 125kVA, 4-quadrant IBRs and, one 3- ϕ , 135kW diesel genset.

1.3.2 Contribution Towards Robust PHIL DRTS Method

Section 1.2.2 discusses the major challenges in developing a stable PHIL interfacing controller for accurate PHIL-based DRTS platform. In the final part of this thesis, **Chapter 6** makes important contributions in developing new modern control-based methods to achieve stable and accurate PHIL-based DRTS platform. It introduces a PHIL interface controller that is sufficiently robust under uncertainties caused by the variations in software and the hardware system. First, the key challenges of properly interfacing the HUT (hardware system) to the simulated model (software system) through the interface controller is highlighted. It shows that considerable amount of challenges remain where often the performance of the PHIL system is not robust and the fidelity to the real-world needs to be quantified and improved. This chapter addresses these challenges by synthesizing an interface between the hardware and software components of PHIL while employing a modern control perspective for managing inherent uncertainties. The chapter describes a methodology that quantifies model uncertainty caused by the point-of-connection in the emulated power network of the software system and the model uncertainty stemming from the varying nature of the hardware-under-test. To overcome these uncertainties, the chapter proposes a robust PHIL interface controller based on μ -synthesis. This controller ensures multiple objectives are met that includes robust stability, performance, accuracy, and tracking capabilities. To assess the effectiveness and viability of the proposed PHIL interface, a PHIL experiment is conducted. The experiment involves interfacing an emulated software system based on a 225-bus, 110V, 60Hz, 1MW residential sub-network of the University of Minnesota and suburb Minneapolis. The interface is established using a real-time simulator and a physical hardware system consisting of two 1- ϕ 1.67kVA inverter systems and one linear and nonlinear load, each with a capacity of 1.8kVA.

1.4 Publications

A large share of this thesis is based on the following publications:

Journal Articles:

- [1] S. Chakraborty, S. Patel, G. Saraswat, A. Maqsood, and M. V. Salapaka, “Seamless Transition of Critical Infrastructures Using Droop-Controlled Grid-Forming Inverters,” in *IEEE Transactions on Industrial Electronics*, vol. 71, no. 2, pp. 1535-1546, Feb. 2024, doi: 10.1109/TIE.2023.3253946.
- [2] S. Chakraborty, S. Patel, and M. V. Salapaka, “ μ -Synthesis-Based Generalized Robust Framework for Grid-Following and Grid-Forming Inverters,” in *IEEE Transactions on Power Electronics*, vol. 38, no. 3, pp. 3163-3179, March 2023, doi: 10.1109/TPEL.2022.3226224.
- [3] S. Chakraborty, J. Park, G. Saraswat, T. Meyers, J. Wang, S. Tiwari, V. Khatana, A. Maqsood, A. Somani, M. V. Salapaka, “Emergency Power Supply System for Critical Infrastructures: Design and Large Scale Hardware Demonstration,” accepted in *IEEE Access*.
- [4] S. Chakraborty, M. T. Rana, and M. V. Salapaka, “A Robust Power Hardware-in-the-Loop Interface under Uncertain Software and Hardware System,” under revision in *IEEE Transactions on Industrial Electronics*.
- [5] S. Patel, S. Chakraborty, B. Lundstrom, S. M. Salapaka and M. V. Salapaka, “Isochronous Architecture-Based Voltage-Active Power Droop for Multi-Inverter Systems,” in *IEEE Transactions on Smart Grid*, vol. 12, no. 2, pp. 1088-1103, March 2021, doi: 10.1109/TSG.2020.3037159.

Conference Articles:

- [1] S. Chakraborty, S. Patel and M. V. Salapaka, “Robust and Optimal Single-loop Voltage Controller for Grid-forming Voltage Source Inverters,” 2020 IEEE Power and Energy Conference at Illinois (PECI), Champaign, IL, USA, 2020, pp. 1-7, doi: 10.1109/PECI48348.2020.9064633.
- [2] S. Chakraborty, S. Patel and M. V. Salapaka, “Design of H_∞ -based Robust Controller for Single-phase Grid-feeding Voltage Source Inverters,” 2020 52nd North American Power Symposium (NAPS), Tempe, AZ, USA, 2021, pp. 1-6, doi: 10.1109/NAPS50074.2021.9449747.

- [3] S. Chakraborty et al., “Novel Power-Hardware-in-the-Loop Interface Method for Grid-Forming Inverter Systems,” IECON 2022 – 48th Annual Conference of the IEEE Industrial Electronics Society, Brussels, Belgium, 2022, pp. 1-6, doi: 10.1109/IECON49645.2022.9968638.
- [4] S. Chakraborty, M. T. Rana and M. V. Salapaka, “Active Synchronization of Islanded Microgrid using Droop-controlled Grid-forming Inverters,” IECON 2022 – 48th Annual Conference of the IEEE Industrial Electronics Society, Brussels, Belgium, 2022, pp. 1-6, doi: 10.1109/IECON49645.2022.9968775.
- [5] S. Chakraborty, Govind Saraswat, Nischal Guruwacharya, Arnab Dey, Richard Bryce, Murti V. Salapaka, “Evaluation of Distributed Power Apportioning with Net Load Management Engine in Microgrids using Power Hardware-in-the-loop Simulation,” accepted in NAPS 2023 – 2023 North American Power Symposium.
- [6] A. Dey, S. Chakraborty and M. V. Salapaka, “Cyber-secured Microgrid Operation for Enhancing Resiliency Against False Data Injection,” 2022 North American Power Symposium (NAPS), Salt Lake City, UT, USA, 2022, pp. 1-6, doi: 10.1109/NAPS56150.2022.10012192.
- [7] A. Dey, S. Chakraborty and M. V. Salapaka, “An End-to-End Cyber-Physical Infrastructure for Smart Grid Control and Monitoring,” 2022 IEEE Power & Energy Society Innovative Smart Grid Technologies Conference (ISGT), New Orleans, LA, USA, 2022, pp. 1-5, doi: 10.1109/ISGT50606.2022.9817539.
- [8] J. Wang, S. Chakraborty, V. Khatana, B. Lundstrom, G. Sarawat and M. Salapaka, “Evaluation of Optimal Net Load Management in Microgrids Using Hardware-in-the-Loop Simulation,” 2022 IEEE Power & Energy Society Innovative Smart Grid Technologies Conference (ISGT), New Orleans, LA, USA, 2022, pp. 1-5, doi: 10.1109/ISGT50606.2022.9817515.
- [9] V. Khatana, S. Chakraborty, G. Saraswat, B. Lundstorm and M. V. Salapaka, “Evaluation of Horizon of Viability Optimization Engine for Sustained Power to Critical Infrastructure,” 2022 IEEE Power and Energy Conference at Illinois (PECI), Champaign, IL, USA, 2022, pp. 1-8, doi: 10.1109/PECI54197.2022.9744038.
- [10] S. Patel, S. Chakraborty, S. Roychowdhury and M. V. Salapaka, “Virtual Impedance

Shaping for low voltage Microgrids,” 2020 IEEE Applied Power Electronics Conference and Exposition (APEC), New Orleans, LA, USA, 2020, pp. 589-596, doi: 10.1109/APEC39645.2020.9124563.

Chapter 2

Robust Microgrid: A Generalized Control for Grid-following and Grid-forming Inverters

2.1 Gist of the Chapter

Grid-following and grid-forming inverters are integral components of microgrids and for integration of renewable energy sources with the grid. For grid-following (GFL) inverters, which need to emulate controllable current sources, a significant challenge is to address the large uncertainty of the grid impedance. For grid-forming (GFM) inverters, which need to emulate a controllable voltage source, large uncertainty due to varying loads has to be addressed. This chapter presents a generalized control framework by leveraging the voltage-current duality in the plant dynamic model of GFL and GFM inverters. The modeling of uncertainties are also generalized under the control framework by quantifying the uncertainties in grid impedance parameters and the uncertainties in equivalent loading parameters for GFL and GFM inverters respectively. Based on the generalized control framework, a μ -synthesis-based robust control design methodology is developed for both GFL and GFM inverters. The control objectives, while designing the optimal controllers, are reference tracking, disturbance rejection and harmonic compensation capability with i) sufficient LCL resonance damping under large variations of grid impedance uncertainty for GFL inverters and ii) with enhanced dynamic response under large variations of equivalent loading uncertainty for GFM inverters. A combined

system-in-the-loop (SIL), controller-hardware-in-the-loop (CHIL) and power-hardware-in-the-loop (PHIL) -based experiment is conducted to evaluate the efficacy and viability of the robust controllers.

Note: More details on this work can be found in the published articles [14].

2.2 State-of-the-art on Control Schemes for Grid-following and Grid-forming Inverters

With the proliferation of inverter-interfaced distributed energy resources, there is a renewed emphasis on local microgrids that provide operational flexibility and aid sustainability for the energy infrastructures. Both grid-forming (GFM) and grid-following (GFL) inverters have become essential components that have pivotal roles to play in such microgrids operating both in grid-tied and islanded mode. During islanded mode, GFM inverters maintain a stable voltage and frequency of the microgrid in the absence of grid. On the other hand, GFL inverters are usually operated to supply power where voltage and frequency are maintained either by the grid or other GFM inverters [15]. In the hierarchical structure of microgrid control system, *inner* voltage-control loops, regulating voltage to specified values, are responsible for GFM inverters to emulate controllable voltage sources. Similarly, GFL inverters emulate controllable current sources by regulating currents via *inner* current-control loop [16].

2.2.1 State-of-the-art on Control Schemes for Grid-following Inverters

Typically GFL inverters are connected to grid via LCL filters for high frequency attenuation caused by switching. Multiple important factors are considered in the design stage of GFL inverters including:

1. Resonances caused by low-damping of the LCL filter in GFL inverters, while connected to weak grid, that could lead to system instability. Here, proper damping of such resonance is crucial while designing [17],
2. Uncertainty in variation of grid impedance parameters significantly influence the robustness of the output current controller. For example, increase in grid inductance requires reduction in the gain and bandwidth of the current controller to keep the system stable that leads to degradation of tracking performance and disturbance rejection capability. Here, there is a need for control design that delivers

Table 2.1: A summary of major current-controllers for grid-following inverters proposed.

Ref.	LCL Damping	Robustness Against Grid Impedance Parameter Variation
[17] [20, 21, 22, 23, 24, 25, 26, 27, 28, 29]	• New filter topology • Additional feedforward of grid voltage, grid current, inner capacitor current • State feedback control • Lack robust damping	• Lack robustness under varying unknown grid impedance • Designed with known, fixed, nominal grid impedance
[18, 19] [30, 3, 31, 32, 33, 34] [35, 36, 37, 38, 39, 40]	• Active/Passive damping • Additional outer current inner current loop • Inherent active damping using controller shaping • Challenging during varying grid impedance	• Lack robustness under varying unknown grid impedance • Grid impedance uncertainty not designed for controller synthesis
Proposed	✓ Single loop control leads reduction in measurements ✓ Inherent active damping using controller shaping ✓ Robust damping under varying grid impedance	✓ Grid impedance uncertainty designed for controller synthesis ✓ Exhibit robustness under varying unknown grid impedance

optimal performance while guaranteeing stability for the range of grid impedance encountered in practice [18],

3. Grid impedance variation causes variation in resonance frequency of the inverter system that impacts active damping methods. Here, robustness of the active damping to remain effective under uncertainty is required [19],
4. Most importantly, the controller should result in good tracking performance, disturbance rejection and harmonic compensation capability while remaining implementable in low-cost micro-controllers.

Various types of control schemes and their advancements for GFL inverters are proposed in literature [41]. Major current-control schemes for GFL inverters, reported in existing literature, are compared in Table 2.1. In summary, existing methods include classical (i.e. proportional/integral/resonant controller-based), hysteresis, sliding-mode, model predictive, repetitive, linear-quadratic regulator, $\mathcal{H}_\infty$ -based control schemes as reported in [41, 17, 29, 20, 21, 22, 23, 24, 25, 26, 27, 28, 18, 19, 30, 3, 31, 32, 33, 34, 35, 36, 37, 38, 39, 40, 42, 43, 44, 45]. Most of the existing schemes emphasize on the enhancement of the

performance of the current controller under a fixed and large value of grid impedance parameter. Addition of a non-varying grid impedance parameters during controller synthesis can enhance the controller performance to an extent but unable to guarantee the robust performance criterion under varying grid impedance parameter condition. Moreover, broadly these schemes use either passive damping with or without new filter topology or active damping using either additional measurements for feedforward control action. These additional measurements increase the cost by employing multiple sensors.

2.2.2 State-of-the-art on Control Schemes for Grid-forming Inverters

Designing the voltage controller for GFM inverter is essentially a multi-objective task. The design considerations include:

1. Reference tracking, disturbance rejection and harmonic compensation in presence of various linear and non-linear loads,
2. The voltage controllers are required to provide compensation to dynamic variations of the output load current and improve the dynamic response [46],
3. Unknown nature of the output loading of GFM inverter can significantly alter the system behavior. Here during heavily loaded condition of GFM inverter, transient performance is severely compromised [47]. Therefore, the voltage controller should be robust enough against unknown loading uncertainties to perform all the aforementioned tasks.

Numerous voltage control strategies are proposed in the literature during past decades for GFM inverter system [64]. Major voltage-control schemes, reported in existing literature, are shown in Table 2.2. In summary, there are nested-loop classical (i.e. proportional/integral/resonant controller-based), sliding-mode, model predictive, repetitive, linear-quadratic regulator, $\mathcal{H}_\infty$ -based, $\mathcal{H}_2$ -based control schemes as reported in [64, 47, 48, 49, 50, 51, 52, 53, 54, 55, 56, 57, 58, 59, 60, 61, 62, 63]. Most of the advancements have focused on either multiple nested-loop structures with advanced control techniques or adding extra measurements as feedforward for enhancing dynamic performance. This results in either increased complication of the control structure that lead to difficulties in implementation or increase in the capital cost of employing multiple sensors. Also, most of the schemes face are unable to guarantee robust performance of the voltage controller under varying loading condition of the inverter. As a result, these

Table 2.2: A summary of major current-controllers for grid-following inverters proposed.

Ref.	Improved Dynamic Response	Robustness Against Equivalent Load Parameter Variation
[47] [48, 49, 50] [51, 52, 53]	• Multiloop control structure • Feedback of inductor current, inner capacitor current • Disturbance observer design • Dominant pole elimination • Need multiple measurements	• Lack robustness during variation of equivalent loading • Designed controller for rated loading condition • No loading uncertainty
[54, 55, 56] [57, 58, 59, 60] [61, 62, 63]	• Multiloop control structure with sliding mode, LQR, $\mathcal{H}_2/\mathcal{H}_\infty, \mathcal{H}_\infty/\mathcal{H}_\infty$ control • Complicated, loopy structure with multiple measurements • Challenging during varying equivalent loading conditions	• Lack robustness for full range variation of equivalent loading • Lack robust performance • Designed controller for simple load uncertainty, without non-linear load
Proposed	✓ $\mathcal{H}_\infty$-based robust control ✓ Single loop control structure ✓ No need of feedforward loop ✓ Reduction in measurements ✓ Enhancement of dynamic response by controller shaping	✓ Equivalent loading uncertainty designed for controller synthesis ✓ Exhibit robustness under the variation of equivalent loading

techniques face deteriorating dynamic response in controller performance under varying and uncertain equivalent loading. Moreover, there are some advanced control strategies applicable for GFM inverter system such as GFM inverter enabled virtual power plant control with inertia support capability, virtual synchronous generator control, Sliding-mode control-based grid-supporting inverters, virtual oscillator control etc., that can found in [65, 66, 67, 68, 69]. However, these strategies are fundamentally responsible for enhancing the “Primary Control” layer of microgrid control hierarchy by generating voltage reference signal for “Zero-level Control” layer.

2.3 Problem Formulation and Motivation

In this section control problem formulation will be discussed for inverter system operating in GFL and GFM mode. For both GFL and GFM inverters, a single-phase H-bridge inverter is considered comprising a dc bus, V_{DC} , four switching devices, S_1, S_2, S_3, S_4 , and an LCL filter with L_f, L_g, R_f, R_g and C_f as inverter and grid side filter inductors, parasitic resistances and filter capacitor respectively. A control layer is employed with

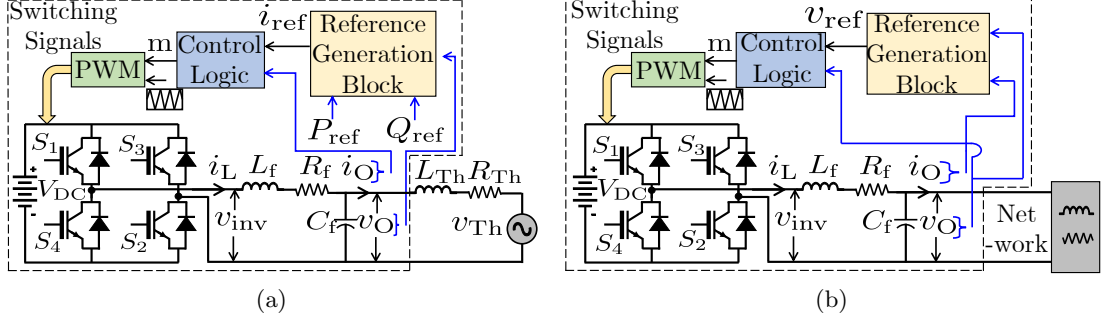


Figure 2.1: A 1- ϕ (a) grid-following inverter, and (b) grid-forming inverter.

sinusoidal PWM switching technique to generate the switching signals for power circuit.

2.3.1 Problem Formulation for Grid-following Inverter

A GFL inverter is connected to a distribution grid, represented by the Thevenin equivalent voltage source, v_{Th} , in series with the Thevenin equivalent impedance, $Z_{Th} := R_{Th} + j\omega_o L_{Th}$. ω_o (in rad/sec) is the frequency of the distribution network. For simplicity, the Thevenin equivalent impedance accounts grid side filter parameters also (i.e. L_g, R_g). Various components of GFL inverter system are shown in Fig. 2.1(a). In this operation, the inverter operates with an output current control strategy to regulate real and reactive power output where the voltage and frequency of the distribution network are determined by another source such as the grid or other GFM inverters. The goal of the current controller is to generate a controlled voltage signal, v_{inv} , by switching signals such that the output current, i_O , tracks the reference signal, i_{ref} , generated by reference generation block as discussed in Appendix A.1. The dynamics of the inverter are described as:

$$L_f \frac{d\langle i_L \rangle}{dt} + R_f \langle i_L \rangle = \langle v_{inv} \rangle - \langle v_O \rangle, \quad (2.1)$$

$$L_{Th} \frac{d\langle i_O \rangle}{dt} + R_{Th} \langle i_O \rangle = \langle v_O \rangle - \langle v_{Th} \rangle, \quad (2.2)$$

$$C_f \frac{d\langle v_O \rangle}{dt} = \langle i_L \rangle - \langle i_O \rangle, \quad (2.3)$$

where $\langle \cdot \rangle$ signifies the average values of the corresponding variable over one switching cycle (T_s) [46]. Laplace transformation and algebraic manipulation with (2.1), (2.2) and

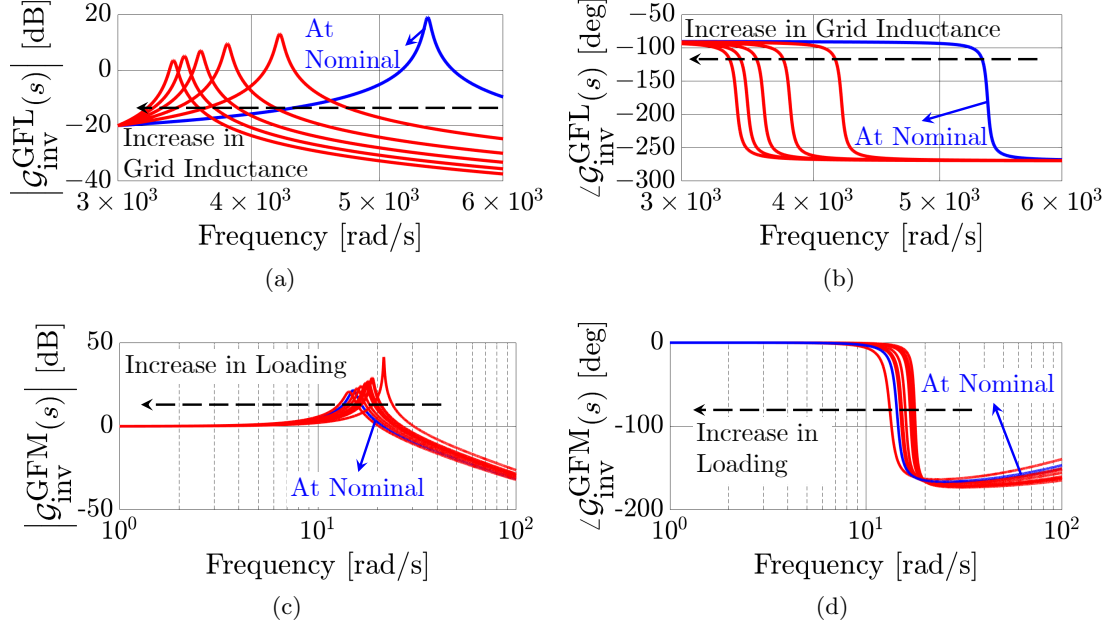


Figure 2.2: Bode plots of (a) magnitude and (b) phase of $\mathcal{G}_{\text{inv}}^{\text{GFL}}(s)$ of (2.4) and (c) magnitude and (d) phase of $\mathcal{G}_{\text{inv}}^{\text{GFM}}(s)$ of (2.13) respectively.

(2.3) results in:

$$i_{\text{O}}(s) = \mathcal{G}_{\text{inv}}^{\text{GFL}}(s)v_{\text{inv}}(s) - \mathcal{G}_{\text{Th}}^{\text{GFL}}(s)v_{\text{Th}}(s), \quad (2.4)$$

where, $\mathcal{G}_{\text{inv}}^{\text{GFL}}(s)$ and $\mathcal{G}_{\text{Th}}^{\text{GFL}}(s)$ are transfer functions parameterized by L_{f} , R_{f} , C_{f} , L_{Th} and R_{Th} , as given in (2.5) and (2.6). As observed in (2.4), distribution network has two-fold impacts on the open-loop plant dynamics. $v_{\text{Th}}(s)$ acts as an exogenous disturbance signal to the plant which can be addressed by classical disturbance rejection problem. Both $\mathcal{G}_{\text{inv}}^{\text{GFL}}(s)$ and $\mathcal{G}_{\text{Th}}^{\text{GFL}}(s)$ consist of L_{Th} and R_{Th} as parameters. Variation of these parameters due to changing distribution network topology introduces uncertainties in the plant model [36, 37]. Fig. 2.2(a) and Fig. 2.2(b) show the Bode plot of $\mathcal{G}_{\text{inv}}^{\text{GFL}}(s)$ that clearly shows the large variation in frequency response of the open-loop plant model

$$\mathcal{G}_{\text{inv}}^{\text{GFL}}(s) = 1 / [(C_{\text{f}}L_{\text{f}}L_{\text{Th}})s^3 + C_{\text{f}}(L_{\text{Th}}R_{\text{f}} + L_{\text{f}}R_{\text{Th}})s^2 + (C_{\text{f}}R_{\text{f}}R_{\text{Th}} + L_{\text{f}} + L_{\text{Th}})s + (R_{\text{f}} + R_{\text{Th}})] \quad (2.5)$$

$$\mathcal{G}_{\text{Th}}^{\text{GFL}}(s) = [(C_{\text{f}}L_{\text{f}})s^2 + (C_{\text{f}}R_{\text{f}})s + 1]\mathcal{G}_{\text{inv}}^{\text{GFL}}(s) \quad (2.6)$$

due to grid inductance variation. Fig. 2.2(a) shows that the equivalent LCL resonant peak is sensitive to grid inductance, L_{Th} . For stiff grid with small L_{Th} (leading to a large resonant frequency), the resultant resonant frequency is larger than the bandwidth of the controller. However, such a controller when used for a sufficiently weak grid with large L_{Th} , the resultant resonance peak may enter the pass-band of the current controller that in turn results in instability [36, 37]. As a result, this uncertainty in grid impedance parameters introduces challenges with respect to performance under uncertainty of grid impedance. With this motivation, this chapter designs a single-loop μ -synthesis-based stabilizing controller for GFL inverter that has robust active damping, tracking performance, disturbance rejection and harmonic compensation under grid impedance uncertainties.

2.3.2 Problem Formulation for Grid-forming Inverter

Various components of a GFM inverter are shown in Fig. 2.1(b). A GFM inverter is connected to a electrical network of a distribution system. In this mode, the inverter operates with an output voltage control strategy to generate a stable voltage and frequency for the electrical network. The goal of the voltage control logic is to generate a controlled voltage signal, v_{inv} , by switching signals such that the output voltage, v_O , tracks the reference signal, v_{ref} , generated by reference generation layer as discussed in Appendix A.2. The dynamics of the inverter are described as:

$$L_f \frac{d\langle i_L \rangle}{dt} + R_f \langle i_L \rangle = \langle v_{inv} \rangle - \langle v_O \rangle, \quad (2.7)$$

$$C_f \frac{d\langle v_O \rangle}{dt} = \langle i_L \rangle - \langle i_O \rangle, \quad (2.8)$$

where $\langle . \rangle$ signifies the average values of the corresponding variable over one switching cycle (T_s) [46]. Laplace transformation and algebraic manipulation with (2.7) and (2.8) result:

$$v_O(s) = \mathcal{G}_1(s)v_{inv}(s) - \mathcal{G}_2(s)i_O(s), \quad (2.9)$$

where, $\mathcal{G}_1(s)$ and $\mathcal{G}_2(s)$ are transfer functions parameterized by L_f , R_f , C_f , as given in (2.10) and (2.11) and described by:

$$\mathcal{G}_1(s) = \frac{1}{[L_f C_f]s^2 + [R_f C_f]s + 1} \text{ and} \quad (2.10)$$

$$\mathcal{G}_2(s) = \frac{L_f s + R_f}{[L_f C_f]s^2 + [R_f C_f]s + 1}. \quad (2.11)$$

For the modeling of i_O , it is essential to understand the impact of the external network on the open-loop plant. In this work, the disturbance is modeled by a parallel combination of two components. First component is a linear admittance, $Y_{\text{Load}}(s) := 1/(L_{\text{Load}}s + R_{\text{Load}})$, with unknown R_{Load} and L_{Load} elements in series combination; the grid side filter parameters are included in R_{Load} and L_{Load} . Another component is a parallel combination of current sources consisting of both fundamental and harmonic components [70] and defined as: $i_h(s) := \sum_k i_k(s)$, where k is odd and $i_k(s)$ is k^{th} harmonic current. As a result, $i_O(s)$ is characterized as:

$$i_O(s) = Y_{\text{Load}}(s)v_O(s) + i_h(s). \quad (2.12)$$

Combining (2.9) and (2.12), system's closed-loop is described by:

$$v_O(s) = \mathcal{G}_{\text{inv}}^{\text{GFM}}(s)v_{\text{inv}}(s) - \mathcal{G}_{\text{Load}}^{\text{GFM}}(s)i_h(s), \quad (2.13)$$

where, $\mathcal{G}_{\text{inv}}^{\text{GFM}}(s)$ and $\mathcal{G}_{\text{Load}}^{\text{GFM}}(s)$ are transfer functions parameterized by L_f , R_f , C_f , L_{Load} and R_{Load} , as given in (2.14) and (2.15). As observed in (2.13), load has impacts on the open-loop plant dynamics of the GFM inverter. Firstly, both $\mathcal{G}_{\text{inv}}^{\text{GFM}}(s)$ and $\mathcal{G}_{\text{Load}}^{\text{GFM}}(s)$ consist of L_{Load} and R_{Load} as parameters. that introduce uncertainties in the plant model. Variation of these parameters due to loading of the inverter due to changing generation-load imbalance in the distribution network introduces uncertainties in the plant model [54, 55]. Secondly, $i_h(s)$, imposed by the fundamental and harmonic current load, acts as an exogenous disturbance signal to the plant. Where the latter one is a classical disturbance rejection problem, the former one poses robustness issues. Fig. 2.2(c)

$$\mathcal{G}_{\text{inv}}^{\text{GFM}}(s) = [L_{\text{Load}}s + R_{\text{Load}}]/[(C_f L_f L_{\text{Load}})s^3 + C_f(L_{\text{Load}}R_f + L_f R_{\text{Load}})s^2 + (C_f R_f R_{\text{Load}} + L_f + L_{\text{Load}})s + (R_f + R_{\text{Load}})] \quad (2.14)$$

$$\mathcal{G}_{\text{Load}}^{\text{GFM}}(s) = \mathcal{G}_{\text{inv}}^{\text{GFM}}(s)[(L_{\text{Load}}L_f)s^2 + (L_{\text{Load}}R_f + L_f R_{\text{Load}})s + (R_{\text{Load}}R_f)]/[L_{\text{Load}}s + R_{\text{Load}}] \quad (2.15)$$

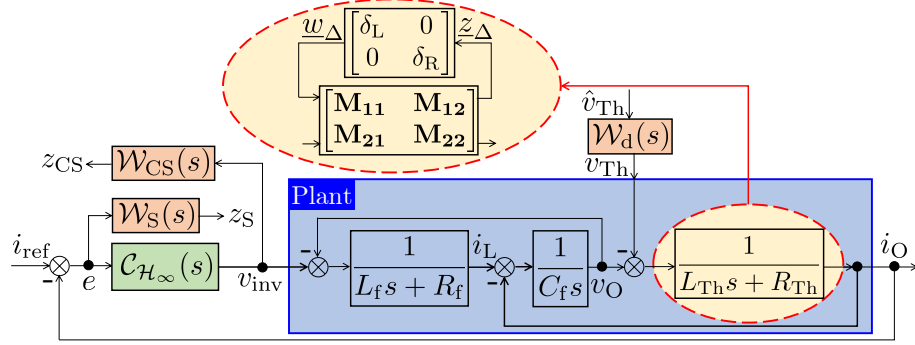


Figure 2.3: Proposed μ -synthesis-based robust controller for GFL Inverter.

and Fig. 2.2(d) depict the Bode plot of $\mathcal{G}_{\text{inv}}^{\text{GFM}}(s)$ that shows the variation in frequency response of the open-loop plant model due to variation in equivalent loading. It is also evident that the uncertain loading variation across the inverter causes significant change in the plant dynamics, especially the effective resonant frequency of the inverter system as shown in Fig. 2.2(c). This deteriorates the transient response of the GFM inverter severely as well as overall system stability [57, 47]. with the above motivation, this chapter designs a single-loop μ -synthesis-based stabilizing controller that has robust tracking performance, disturbance rejection and harmonic compensation with improved transient operation under the full range variation in equivalent loading.

2.4 μ -synthesis-based Generalized Framework for Controller Synthesis

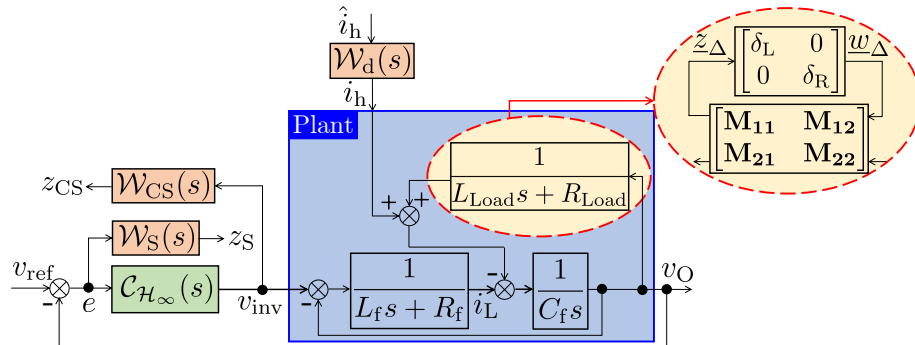


Figure 2.4: Proposed μ -synthesis-based robust controller for GFM Inverter.

Motivated by Section 2.3.1 and 2.3.2, this section will describe the generalized control framework and required modeling for μ -synthesis-based controller synthesis.

2.4.1 μ -synthesis-based Generalized Framework

The following observations are common for both GFL and GFM inverter system:

1. The open-loop plant model, $\mathcal{G}_{\text{inv}}^{\text{GFL}}(s)$ of (2.5) and $\mathcal{G}_{\text{inv}}^{\text{GFM}}(s)$ of (2.14) are 3rd order model with different numerators for GFL and GFM inverter respectively. The disturbance model, $\mathcal{G}_{\text{Th}}^{\text{GFL}}(s)$ of (2.6) and $\mathcal{G}_{\text{Load}}^{\text{GFM}}(s)$ of (2.15), are 3rd order models for GFL and GFM respectively. Moreover, there is a voltage-current duality in the plant dynamic model of GFL and GFM inverter. Here for both the models, v_{inv} is the *plant input* signal, i_{O} , v_{O} are the *plant output* and v_{Th} , i_{h} are the *disturbance* signal for GFL and GFM inverter respectively as given in (2.4) and (2.13).
2. The open-loop plant and disturbance models have similar parametric uncertainties imposed by L_{Th} , R_{Th} and L_{Load} and R_{Load} for both GFL and GFM inverter respectively. The similarity is evident in Fig. 2.3 and Fig. 2.4 where the open-loop plant models are shown using block diagram inside the blue boxes and corresponding uncertainty block using dashed ovals for both GFL and GFM inverter respectively.
3. Robust tracking performance, disturbance rejection and harmonic compensation capabilities are desired for controllers of GFL and GFM inverters under the uncertainties.

Item 1 and 2 motivate designing a generalized control framework for GFL and GFM inverters based on these similarity and duality property. Third point warrants the μ -synthesis-based robust controller synthesis for addressing multiple objectives. A systematic approach is presented below for designing controller, $\mathcal{C}_{\mathcal{H}_{\infty}}(s)$ of Fig. 2.3 and Fig. 2.4 for GFL and GFM inverter respectively using the generalized μ -synthesis-based robust control framework.

2.4.2 Modeling of Uncertainty

In both the GFL and GFM inverter open-loop plant model, the model uncertainty is present as parametric uncertainties in 1st order transfer function (dashed ovals in Fig. 2.3 and Fig. 2.4 respectively). A systematic approach is followed here for characterizing and modeling the uncertainty for both the cases as discussed below:

Characterizing the Uncertainty for GFL Inverter

Variations in grid impedance (i.e. L_{Th} and R_{Th}) results in real-parametric uncertainties in the GFL plant model. The short-circuit ratio (SCR) is often used to characterize the grid stiffness and can be employed in determining the Thevenin equivalent impedance of the grid at the point-of-connection. SCR is defined as $(V_{PC}^N)^2 / [S_B \sqrt{(\omega_N L_{Th})^2 + R_{Th}^2}]$, where V_{PC}^N and ω_N are the nominal voltage and frequency at point-of-connection, and S_B is the rated apparent power of the GFL inverter. Usually the grid at point-of-connection is considered as weak when the SCR is less than 3 [71, 70]. In this work, with a pre-specified SCR (< 3) and X/R ratio (< 10), the nominal grid impedance parameters, L_{Th}^{Nom} and R_{Th}^{Nom} , are determined. By considering $\pm 100\%$ variations over nominal values, it is assumed that $L_{Th} \in [\underline{L}_{Th}, \bar{L}_{Th}]$ and $R_{Th} \in [\underline{R}_{Th}, \bar{R}_{Th}]$. It is to be noted that stiff to extremely weak grid conditions are accommodated with this uncertainty characterization. As a result,

$$L_{Th} := L_{Th}^{Nom} + w_{Th}^L \delta_L, \quad R_{Th} := R_{Th}^{Nom} + w_{Th}^R \delta_R, \quad (2.16)$$

where, $\delta_L, \delta_R \in [-1, 1]$, $L_{Th}^{Nom} = \frac{1}{2}[\bar{L}_{Th} + \underline{L}_{Th}]$, $R_{Th}^{Nom} = \frac{1}{2}[\bar{R}_{Th} + \underline{R}_{Th}]$, $w_{Th}^L = \frac{1}{2}[\bar{L}_{Th} - \underline{L}_{Th}]$, $w_{Th}^R = \frac{1}{2}[\bar{R}_{Th} - \underline{R}_{Th}]$.

Characterizing the Uncertainty for GFM Inverter

In this case, variation in equivalent loading (i.e. R_{Load} and L_{Load}) results in real-parametric uncertainties in the GFM plant model. In this work, the linear part of the loading is modeled by a series combination of equivalent unknown R_{Load} and L_{Load} element. These elements are at nominal while the GFM inverter loading is at rated condition. Considering rated VA loading as S_{rated} , with active power, P_{rated} , and reactive power, Q_{rated} , the following hold for nominal values:

$$R_{Load}^{Nom} = V_N^2 \frac{P_{rated}}{S_{rated}^2}, \quad L_{Load}^{Nom} = V_N^2 \frac{Q_{rated}}{S_{rated}^2} \frac{1}{\omega_N}, \quad (2.17)$$

where V_N and ω_N are the nominal voltage and frequency of the network respectively. By considering no-loading and overloading (200% loading) of GFM, it is assumed that

$L_{\text{Load}} \in [\underline{L}_{\text{Load}}, \bar{L}_{\text{Load}}]$, $R_{\text{Load}} \in [\underline{R}_{\text{Load}}, \bar{R}_{\text{Load}}]$. As a result,

$$L_{\text{Load}} := L_{\text{Load}}^{\text{Nom}} + w_{\text{Load}}^{\text{L}} \delta_{\text{L}}, \quad (2.18)$$

$$R_{\text{Load}} := R_{\text{Load}}^{\text{Nom}} + w_{\text{Load}}^{\text{R}} \delta_{\text{R}}, \quad (2.19)$$

where, $\delta_{\text{L}}, \delta_{\text{R}} \in [-1, 1]$, $L_{\text{Load}}^{\text{Nom}} = \frac{1}{2}[\bar{L}_{\text{Load}} + \underline{L}_{\text{Load}}]$, $R_{\text{Load}}^{\text{Nom}} = \frac{1}{2}[\bar{R}_{\text{Load}} + \underline{R}_{\text{Load}}]$, $w_{\text{Load}}^{\text{L}} = \frac{1}{2}[\bar{L}_{\text{Load}} - \underline{L}_{\text{Load}}]$, and $w_{\text{Load}}^{\text{R}} = \frac{1}{2}[\bar{R}_{\text{Load}} - \underline{R}_{\text{Load}}]$.

Generalized Representation of Uncertainty

In synthesizing the controller with defined uncertainties in $R \in [\underline{R}, \bar{R}]$ and $L \in [\underline{L}, \bar{L}]$ when appearing in the form of $1/(Ls + R)$, linear fractional transformation (LFT)[72] can be utilized to convert the model into an upper LFT, $F_{\text{U}}(\mathbf{M}, \Delta)$, given as:

$$F_{\text{U}}(\mathbf{M}, \Delta) = \frac{1}{sL + R} = \mathbf{M}_{22} + \mathbf{M}_{21}\Delta[\mathbf{I} - \mathbf{M}_{11}\Delta]^{-1}\mathbf{M}_{12},$$

$$\text{with } \mathbf{M} = \begin{bmatrix} \mathbf{M}_{11} & \mathbf{M}_{12} \\ \mathbf{M}_{21} & \mathbf{M}_{22} \end{bmatrix}, \Delta = \begin{bmatrix} \delta_{\text{L}} & 0 \\ 0 & \delta_{\text{R}} \end{bmatrix}, \quad \begin{matrix} \delta_{\text{L}} \in [-1, 1], \\ \delta_{\text{R}} \in [-1, 1], \end{matrix}$$

$$\mathbf{M}_{11} = \mathbf{M}_{22} \begin{bmatrix} sw^{\text{L}} & w^{\text{R}} \\ sw^{\text{L}} & w^{\text{R}} \end{bmatrix}, \quad \mathbf{M}_{12} = \mathbf{M}_{22} \begin{bmatrix} 1 & 1 \end{bmatrix}^{\text{T}},$$

$$\mathbf{M}_{21} = \mathbf{M}_{22} \begin{bmatrix} sw^{\text{L}} & w^{\text{R}} \end{bmatrix},$$

where $\mathbf{M}_{22} = -1/[sL_{\text{Nom}} + R_{\text{Nom}}]$, $w^{\text{L}} = \frac{1}{2}[\bar{L} - \underline{L}]$, $w^{\text{R}} = \frac{1}{2}[\bar{R} - \underline{R}]$. It is important to note here that it is a generalized representation of uncertainty for both GFL and GFM inverter where $L = L_{\text{Th}}$, $R = R_{\text{Th}}$ for GFL inverter and $L = L_{\text{Load}}$, $R = R_{\text{Load}}$ for GFM inverter.

2.4.3 Shaping of Transfer Functions

The closed-loop objectives in designing the feedback control based on proposed robust controller, $\mathcal{C}_{\mathcal{H}_{\infty}}(s)$, as shown in Fig. 2.3 and Fig. 2.4 for GFL and GFM inverter, are as follows:

- i) **Reference Tracking:** i_{O} tracks i_{ref} for GFL and v_{O} tracks v_{ref} for GFM inverter with minimum tracking error.
- ii) **Disturbance Rejection:** Effects of v_{Th} on i_{O} for GFL and effect of i_{h} on v_{O} for GFM inverter are attenuated.
- iii) **Control Effort Reduction:** v_{inv} satisfies the respective bandwidth limitations for

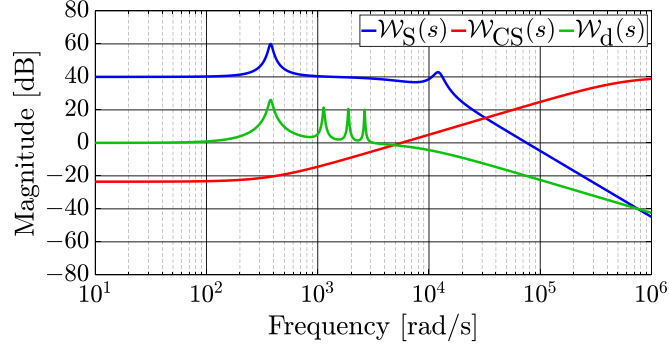


Figure 2.5: Bode plots of magnitudes of selected weighting transfer functions.

GFL and GFM inverter. Based on the objectives, user-defined weighting transfer functions, $\mathcal{W}_S(s)$, $\mathcal{W}_{CS}(s)$, $\mathcal{W}_d(s)$, are designed. The guidelines for designing the weighting functions are provided below.

Selection of $\mathcal{W}_S(s)$

To shape the sensitivity transfer function, the weighting function, $\mathcal{W}_S(s)$, is introduced so that **i)** The tracking error, e , ($e := i_{\text{ref}} - i_O$ and $e := v_{\text{ref}} - v_O$ for GFL and GFM inverter respectively) at fundamental frequency is small; **ii)** Resonance phenomenon of the system is actively damped. $\mathcal{W}_S(s)$ is modeled to have peaks around ω_N and system's resonant frequency, ω_r (different in GFL and GFM plant), with 2nd order roll-off, $k_{S,1}(s)$ and formed as:

$$\mathcal{W}_S(s) = k_{S,1}(s) \frac{s^2 + 2k_{S,2}\zeta\omega_N s + \omega_N^2}{s^2 + 2\zeta\omega_N s + \omega_N^2} \frac{s^2 + 2k_{S,3}\zeta\omega_r s + \omega_r^2}{s^2 + 2\zeta\omega_r s + \omega_r^2},$$

where, $k_{S,2}$ and $k_{S,3}$ are selected to exhibit peaks and ζ addresses the off-nominal frequency around the nominal values.

Selection of $\mathcal{W}_{CS}(s)$

$\mathcal{W}_{CS}(s)$ is designed to suppress high-frequency control effort to shape the performance of v_{inv} for both GFL and GFM controller. Hence, it is designed as a high-pass filter

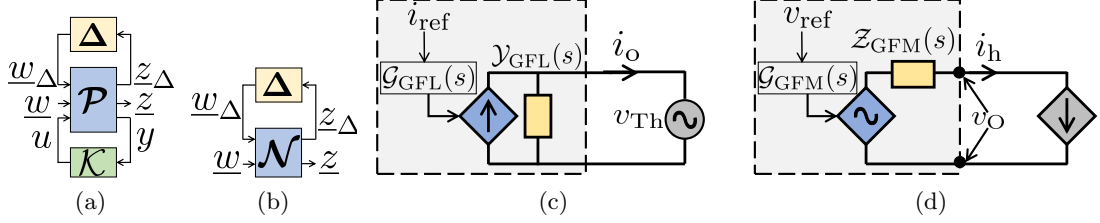


Figure 2.6: Generalized (a) Δ - $\mathcal{P}$ - $\mathcal{K}$ control configuration, (b) Δ - $\mathcal{N}$ control configuration, of Fig. 2.3 or Fig. 2.4 and closed-loop equivalent circuit model of (c) grid-following inverter system, (d) grid-forming inverter system.

with cut-off frequency at switching frequency with the form:

$$\mathcal{W}_{CS}(s) = k_{CS} \frac{s + k_{CS,1}\omega_N}{s + k_{CS,2}\omega_N}, \text{ where } k_{CS,1} \ll k_{CS,2}.$$

Selection of $\mathcal{W}_d(s)$

$\mathcal{W}_d(s)$ emphasizes the disturbances at fundamental and harmonic frequencies imposed by v_{Th} and i_h and emphasized by exogenous signal $\hat{v}_{Th}$ and $\hat{i}_h$ for GFL and GFM inverter respectively, as shown in Fig. 2.3 and Fig. 2.4. It is designed by a low-pass filter, $k_d(s)$, with peaks at selected frequencies and is given by:

$$\mathcal{W}_d(s) = k_d(s) \prod_{h=1,3,5,7} \frac{s^2 + 2k_{d,h}\zeta h\omega_N s + h^2\omega_N^2}{s^2 + 2\zeta h\omega_N s + h^2\omega_N^2},$$

where, the values of $k_{d,h}$ are selected based on the regulated limits of 3rd, 5th, 7th harmonics in network voltage and current injection with respect to fundamental [73]. A representative of selected weighting functions are shown in Fig. 6.5.

2.4.4 Preparing the Generalized Plant

In preparation for robust controller design, $\mathcal{C}_{\mathcal{H}_\infty}(s)$, the multi-loop closed-loop block diagram in Fig. 2.3 and Fig. 2.4 for GFL and GFM inverter respectively are consolidated into the general control configuration in Fig. 2.6(a) [72]. Here, $\mathcal{P}(s)$ is the generalized multi-input-multi-output (MIMO) plant, $\mathcal{K}(s)$ is the proposed $\mathcal{C}_{\mathcal{H}_\infty}(s)$ controllers to be designed for GFL and GFM inverter and Δ is the structured uncertainty. $\underline{w}$ is a vector of the exogenous inputs (e.g., reference, disturbance), $\underline{z}$ are the exogenous outputs (e.g.,

signals to be regulated). y and u are the controller input and output signals respectively. $\underline{z}_\Delta$ and $\underline{w}_\Delta$ are the vector of input and output signals of structured uncertainty block. Note that in this continuous-time modeling framework, all variables are functions of the Laplace variable, s ; not explicitly shown for notational convenience. As a result, the generalized MIMO plant maps $\begin{bmatrix} \underline{w}_\Delta & \underline{w} & u \end{bmatrix}^\top$ to $\begin{bmatrix} \underline{z}_\Delta & \underline{z} & y \end{bmatrix}^\top$ as follows:

$$\begin{bmatrix} \underline{z}_\Delta \\ \underline{z} \\ y \end{bmatrix} = \begin{bmatrix} \mathcal{P}_{\Delta\Delta} & \mathcal{P}_{\Delta w} & \mathcal{P}_{\Delta u} \\ \mathcal{P}_{z\Delta} & \mathcal{P}_{zw} & \mathcal{P}_{zu} \\ \mathcal{P}_{y\Delta} & \mathcal{P}_{yw} & \mathcal{P}_{yu} \end{bmatrix} \begin{bmatrix} \underline{w}_\Delta \\ \underline{w} \\ u \end{bmatrix}, \quad (2.20)$$

where, the input and output signals are tabulated in Table 2.3 for both GFL and GFM inverter system. The detailed MIMO transfer function models of (6.7) for both GFL and GFM inverter systems are provided in Table 4.12 where

$$\begin{aligned} \mathcal{A}_{\text{GFL}} &= \frac{1}{[L_f C_f]s^2 + [R_f C_f]s + 1}, \\ \mathcal{B}_{\text{GFL}} &= \frac{L_f s + R_f}{[L_f C_f]s^2 + [R_f C_f]s + 1}, \\ \mathcal{A}_{\text{GFM}} &= \frac{1}{[L_f C_f]s^2 + [R_f C_f + L_f]s + [1 + R_f]}, \\ \mathcal{B}_{\text{GFM}} &= \frac{L_f s + R_f}{[L_f C_f]s^2 + [R_f C_f + L_f]s + [1 + R_f]}. \end{aligned}$$

$\mathcal{P}$ and $\mathcal{K}$ can be combined together by a lower LFT to form $\mathcal{N}$ in Fig. 2.6(b) as follows:

$$\begin{aligned} \mathcal{N} &= \begin{bmatrix} \mathcal{P}_{\Delta\Delta} & \mathcal{P}_{\Delta w} \\ \mathcal{P}_{z\Delta} & \mathcal{P}_{zw} \end{bmatrix} + \frac{\mathcal{K}}{[1 - \mathcal{P}_{yu}\mathcal{K}]^{-1}} \begin{bmatrix} \mathcal{P}_{\Delta u} \\ \mathcal{P}_{zu} \end{bmatrix} \begin{bmatrix} \mathcal{P}_{y\Delta} \\ \mathcal{P}_{yw} \end{bmatrix}^\top \\ &= \begin{bmatrix} \mathcal{N}_{\Delta\Delta} & \mathcal{N}_{\Delta w} \\ \mathcal{N}_{z\Delta} & \mathcal{N}_{zw} \end{bmatrix}, \text{ where } \begin{bmatrix} \underline{z}_\Delta \\ \underline{z} \end{bmatrix} = \mathcal{N} \begin{bmatrix} \underline{w}_\Delta \\ \underline{w} \end{bmatrix}. \end{aligned} \quad (2.21)$$

Therefore, the uncertainty closed-loop transfer function from $\underline{w}$ to $\underline{z}$, $\underline{z} = \mathcal{M}\underline{w}$, is related to $\mathcal{N}$ and Δ by an upper LFT where $\mathcal{M} = \mathcal{N}_{zw} + \mathcal{N}_{z\Delta}\Delta(\mathbf{I} - \mathcal{N}_{\Delta\Delta}\Delta)^{-1}\mathcal{N}_{\Delta w}$.

Table 2.3: Input-Output Mapping of the Generalized Plant

Signals	GFL Inverter	GFM Inverter
$\underline{w}$	$[i_{\text{ref}} \ \hat{v}_{\text{Th}}]^\top$ of Fig. 2.3	$[v_{\text{ref}} \ \hat{i}_{\text{h}}]^\top$ of Fig. 2.3
$\underline{z}$	$[z_{\text{S}} \ z_{\text{CS}}]^\top$ of Fig. 2.3	$[z_{\text{S}} \ z_{\text{CS}}]^\top$ of Fig. 2.4
u	v_{inv} of Fig. 2.3	v_{inv} of Fig. 2.4
y	e of Fig. 2.3	e of Fig. 2.4
$\underline{w}_\Delta$	$\underline{w}_\Delta$ of Fig. 2.3	$\underline{w}_\Delta$ of Fig. 2.4
$\underline{z}_\Delta$	$\underline{z}_\Delta$ of Fig. 2.3	$\underline{z}_\Delta$ of Fig. 2.4

Table 2.4: The generalized MIMO transfer function models, $\mathcal{P}(s)$, of (6.7) for both GFL inverter control of Fig. 2.3 and GFM inverter control of Fig. 2.4.

Type	$\mathcal{P}(s)$ [$\mathcal{P}_{\Delta\Delta} \in \mathbb{C}^{2 \times 2}$, $\mathcal{P}_{\Delta w} \in \mathbb{C}^{2 \times 2}$, $\mathcal{P}_{\Delta u} \in \mathbb{C}^{2 \times 1}$, $\mathcal{P}_{z\Delta} \in \mathbb{C}^{2 \times 2}$, $\mathcal{P}_{zw} \in \mathbb{C}^{2 \times 2}$, $\mathcal{P}_{zu} \in \mathbb{C}^{2 \times 1}$, $\mathcal{P}_{y\Delta} \in \mathbb{C}^{1 \times 2}$, $\mathcal{P}_{yw} \in \mathbb{C}^{1 \times 2}$, $\mathcal{P}_{yu} \in \mathbb{C}$]				
GFL	$\mathcal{P}_{\Delta\Delta} = \mathbf{M}_{11} - \frac{\mathcal{B}_{\text{GFL}}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}} \mathbf{M}_{12}\mathbf{M}_{21}$	$\mathcal{P}_{\Delta w} = \mathbf{0}_{2 \times 1}$	$\frac{\mathcal{B}_{\text{GFL}}\mathcal{W}_d\mathbf{M}_{22}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}} \mathbf{M}_{12}$	$\mathcal{P}_{\Delta u} = \left(\mathcal{A}_{\text{GFL}} - \frac{\mathcal{B}_{\text{GFL}}\mathcal{A}_{\text{GFL}}\mathbf{M}_{22}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}} \mathbf{M}_{12} \right) \mathbf{M}_{12}$	
	$\mathcal{P}_{z\Delta} = \begin{bmatrix} -\frac{\mathcal{W}_s}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}} \mathbf{M}_{21} \\ \mathbf{0}_{1 \times 2} \end{bmatrix}$	$\mathcal{P}_{zw} = \begin{bmatrix} \mathcal{W}_s \\ 0 \end{bmatrix}$	$\frac{\mathcal{W}_s\mathcal{W}_d\mathbf{M}_{22}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}}$	$\mathcal{P}_{zu} = -\frac{\mathcal{W}_s\mathbf{M}_{22}\mathcal{A}_{\text{GFL}}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}}$	$\mathcal{W}_{\text{CS}}$
	$\mathcal{P}_{y\Delta} = -\frac{1}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}} \mathbf{M}_{21}$	$\mathcal{P}_{yw} = \frac{1}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}}$	$\frac{\mathbf{M}_{22}\mathcal{W}_d}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}}$	$\mathcal{P}_{yu} = -\frac{\mathbf{M}_{22}\mathcal{A}_{\text{GFL}}}{1 + \mathbf{M}_{22}\mathcal{B}_{\text{GFL}}}$	
GFM	$\mathcal{P}_{\Delta\Delta} = \mathbf{M}_{11} - \mathcal{B}_{\text{GFM}}\mathbf{M}_{12}\mathbf{M}_{21}$	$\mathcal{P}_{\Delta w} = \mathbf{0}_{2 \times 1}$	$-\mathcal{W}_d\mathcal{B}_{\text{GFM}}\mathbf{M}_{12}$	$\mathcal{P}_{\Delta u} = \mathcal{A}_{\text{GFM}}\mathbf{M}_{12}$	
	$\mathcal{P}_{z\Delta} = \begin{bmatrix} -\mathcal{W}_s\mathcal{B}_{\text{GFM}}\mathbf{M}_{21} \\ \mathbf{0}_{1 \times 2} \end{bmatrix}$	$\mathcal{P}_{zw} = \begin{bmatrix} \mathcal{W}_s \\ 0 \end{bmatrix}$	$-\mathcal{W}_s\mathcal{W}_d\mathcal{B}_{\text{GFM}}$	$\mathcal{P}_{zu} = -\mathcal{W}_s\mathcal{A}_{\text{GFM}}$	$\mathcal{W}_{\text{CS}}$
	$\mathcal{P}_{y\Delta} = \mathcal{B}_{\text{GFM}}\mathbf{M}_{21}$	$\mathcal{P}_{yw} = \mathbf{1}$	$\mathcal{W}_d\mathcal{B}_{\text{GFM}}$	$\mathcal{P}_{yu} = -\mathcal{A}_{\text{GFM}}$	

2.5 Controller Synthesis and Stability Analysis

With reference to the general control configuration of Fig. 2.6(a), the standard μ -synthesis-based optimal control problem is to find all stabilizing controllers $\mathcal{K}(s)$ by solving

$$\min_{\mathcal{K}(s) \text{ stabilizing}} \|\mathcal{N}\|_{\infty}, \quad (2.22)$$

where, $\|\cdot\|_{\infty}$ refers to the μ -synthesis norm. This problem can be readily solved using the MATLAB Robust Control Toolbox. An algorithm for solving (6.9) along with the theoretical underpinnings of this optimization problem can be found in [72]. Upon finding a stabilizing controller, the requirement of stability and performance of the closed-loop system are needed to be checked and can be summarized as follows:

$$\text{Nominal Stable(NS)} : \mathcal{N} \text{ is internally stable}, \quad (2.23)$$

$$\text{Nominal Performance(NP)} : \|\mathcal{N}_{zw}\|_{\infty} < 1 \text{ \& NS}, \quad (2.24)$$

$$\text{Robust Stable(RS)} : \mu_{\Delta}(\mathcal{N}_{\Delta\Delta}) < 1, \forall \omega, \text{ \& NS}, \quad (2.25)$$

$$\text{Robust Performance(RP)} : \mu_{\bar{\Delta}}(\mathcal{N}) < 1, \forall \omega, \text{ \& NS}, \quad (2.26)$$

where, $\mu_{\Delta}(\mathcal{N}_{\Delta\Delta})$ and $\mu_{\bar{\Delta}}(\mathcal{N})$ are the structured singular values of $\mathcal{N}_{\Delta\Delta}$ and $\mathcal{N}$ for the allowed structure of Δ and $\bar{\Delta} := \text{diag}(\Delta, \Delta_P)$ respectively with Δ_P being an unstructured uncertainty [72]. It is necessary to check whether stabilizing controller, $\mathcal{K}(s)$ of (6.9) satisfies all the conditions of (6.10)-(6.13) to analyze robust performance of the controller. In this work, an iterative approach is followed for μ -synthesis problem (i.e. finding the stabilizing controller that minimizes a given μ -condition). Design guidelines for selecting parameters of $\mathcal{W}_S(s)$, $\mathcal{W}_{CS}(s)$, $\mathcal{W}_d(s)$ are provided below:

- During design of the $\mathcal{W}_S(s)$, $k_{S,1}(s)$, $k_{S,2}$, $k_{S,3}$ and ζ are required to be selected. A desirable selection of $k_{S,1}(s)$ is a 2nd order roll-off low-pass filter with cut-off frequency of around 10-15 times the fundamental frequency. It primarily determines the bandwidth of the resulting controller. The choice of high value of $k_{S,2}$, $k_{S,3}$ ensure to have significantly high value of $\mathcal{W}_S$ at fundamental power frequency (this leads to significantly less tracking error) and at system's resonance frequency (this lead to sufficient active damping of resonance). Moreover, in microgrid scenarios, where frequency can vary significantly, it may be necessary to increase ζ to broaden the frequency range where $\mathcal{W}_S$ is large.

- During design of the $\mathcal{W}_{\text{CS}}(s)$, k_{CS} , $k_{\text{CS},1}$, and $k_{\text{CS},2}$ are required to be selected. It is important to note that $\mathcal{W}_{\text{CS}}(s)$ corresponds to shaping the performance of u of Fig. 2.6(a), i.e. v_{inv} of Fig. 2.3 or Fig. 2.4. Therefore, $\mathcal{W}_{\text{CS}}(s)$ is usually designed to suppress high-frequency control effort produced by the resulting controller. The suitable choice of $k_{\text{CS},1}$, and $k_{\text{CS},2}$ ($k_{\text{CS},1} \ll k_{\text{CS},2}$) should be such that the cut-off frequency is around the switching frequency of the inverter. Moreover, selection of k_{CS} will be determined based on the desired gain of the high pass filter.
- During design of the $\mathcal{W}_{\text{d}}(s)$, $k_{\text{d}}(s)$, $k_{\text{d},1}$, $k_{\text{d},3}$, $k_{\text{d},5}$ and $k_{\text{d},7}$ are required to be selected. $\mathcal{W}_{\text{d}}(s)$ characterizes the response of expected disturbances. For systems with linear and non-linear loads (such as diode rectifier loads and switching power supplies etc), disturbance currents have components primarily at fundamental frequency and harmonics at higher orders. The value of $k_{\text{d},1}$, $k_{\text{d},3}$, $k_{\text{d},5}$ and $k_{\text{d},7}$ can be determined based on the relative harmonic distortion limits of various non-linear loads as provided in [73]. By designing the $k_{\text{d}}(s)$ as a low pass filter with cut-off frequency less than the fundamental power frequency, low-order disturbance rejection is emphasized.

The selected weights are shown in Appendix A.3. The μ -synthesis-based optimal controller design procedure based on the generalized robust framework contains the following steps:

Step-1. Based on the design guidelines, select the performance weights (i.e. $\mathcal{W}_{\text{S}}(s)$, $\mathcal{W}_{\text{CS}}(s)$, $\mathcal{W}_{\text{d}}(s)$),

Step-2. Based on Section 6.4.1, characterize the uncertainties of the model (i.e. to determine the $\mathbf{M}$ of Section 2.4.2)

Step-3. Construct the generalized plant, $\mathcal{P}$. An analytical form of the generalized plant for both GFL and GFM inverter system is given in Table 4.12.

Step-4. Solve the μ -synthesis problem, as stated in (6.9), with the following iterations:

i) Synthesis-stage: Synthesize an $\mathcal{H}_{\infty}$ controller for the generalized plant, $\mathcal{P}$, with initial choice of parameters as guided by Step 1-4.

ii) Checking-stage: Check the conditions of (6.10)-(6.13) are all satisfied or not. If conditions are satisfied, μ -synthesized optimal controller is achieved as the problem in (6.9) is solved. If the conditions are not satisfied, return to Step-1 and adjust either performance weights (i.e. $\mathcal{W}_{\text{S}}(s)$, $\mathcal{W}_{\text{CS}}(s)$, $\mathcal{W}_{\text{d}}(s)$) or uncertainty weights (i.e. w^{L} , w^{R}) and continue until conditions of (6.10)-(6.13) are all satisfied.

Table 2.5: Stability Assessment of Resulting Controller.

Controller	NS	NP	RS	RP
GFL Inverter	✓	0.49	0.91	0.94
GFM Inverter	✓	0.47	0.94	0.98

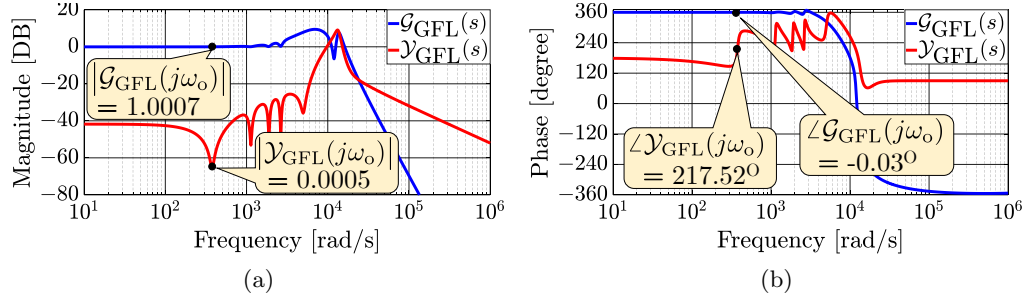


Figure 2.7: Bode plots, (a) magnitudes, (b) phase, of $\mathcal{G}_{\text{GFL}}$, $\mathcal{Y}_{\text{GFL}}$ of Fig. 2.6(c).

The optimal controller $\mathcal{K}(s)$ will have an order similar to the order of $\mathcal{P}$. Thus, before implementation in actual inverter control board, model order reduction is used to obtain a lower order controller using MATLAB's *balred* command. Moreover, bilinear transformation is used in the discretization stage of the resulting controller.

2.5.1 Analysis of Resulting Controller for GFL Inverter

Following the procedure of synthesizing the optimal controller for GFL inverter, a 13th order $\mathcal{C}_{\mathcal{H}_\infty}(s)$ of Fig. 2.3 is found to perform well. The closed-loop stability and desired performances are met as summarized in Table 2.5. The closed-loop model for GFL inverter with negative feedback loop transfer function with resulting controller, $\mathcal{C}_{\mathcal{H}_\infty}(s)$, in Fig. 2.3 can be derived by substituting $v_{\text{inv}}(s) = \mathcal{C}_{\mathcal{H}_\infty}(s)[i_{\text{ref}} - i_{\text{O}}]$ in (2.4). It can be written as $i_{\text{O}} = \mathcal{G}_{\text{GFL}}(s)i_{\text{ref}} - \mathcal{Y}_{\text{GFL}}(s)v_{\text{Th}}$ and represented as Norton's equivalent model connected to a voltage source as shown in Fig. 2.6(c). For an example, at nominal plant condition with resulting optimal controller, the Bode plot of $\mathcal{G}_{\text{GFL}}(s)$ and $\mathcal{Y}_{\text{GFL}}(s)$ are shown in Fig. 2.7(a) and Fig. 2.7(b) respectively. It is observed that $|\mathcal{G}_{\text{GFL}}(j\omega_{\text{N}})|$, $\angle \mathcal{G}_{\text{GFL}}(j\omega_{\text{N}})$ and $|\mathcal{Y}_{\text{GFL}}(j\omega_{\text{N}})|$ are ≈ 1 , $\approx 0^\circ$ and ≈ 0 respectively that leads to $i_{\text{O}} \approx i_{\text{ref}}$ at fundamental frequency.

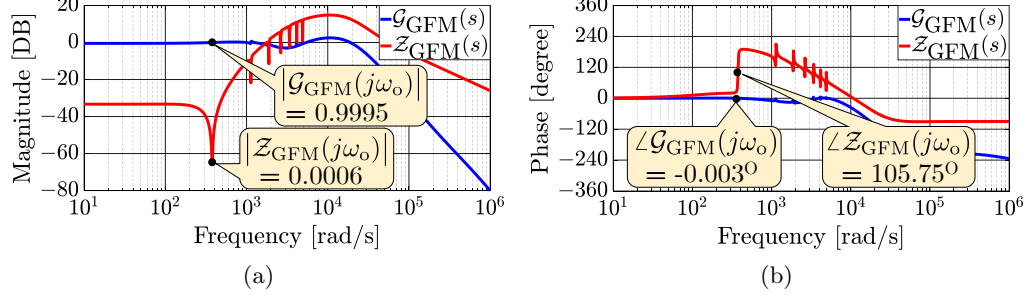


Figure 2.8: Bode plots, (a) magnitudes, (b) phase, of $\mathcal{G}_{\text{GFM}}$, $\mathcal{Z}_{\text{GFM}}$ of Fig. 2.6(d).

2.5.2 Analysis of Resulting Controller for GFM Inverter

Following the procedure of synthesizing the optimal controller for GFM inverter, a 14th order $\mathcal{C}_{\mathcal{H}_{\infty}}(s)$ of Fig. 2.4 is found to be sufficient and performs well. The closed-loop stability and desired performances are met as summarized in Table 2.5. The closed-loop model for GFM inverter with negative feedback loop transfer function with resulting controller, $\mathcal{C}_{\mathcal{H}_{\infty}}(s)$, in Fig. 2.4 can be derived by substituting $v_{\text{inv}}(s) = \mathcal{C}_{\mathcal{H}_{\infty}}(s)[v_{\text{ref}} - v_{\text{O}}]$ in (2.13). It can be written as $v_{\text{O}} = \mathcal{G}_{\text{GFM}}(s)v_{\text{ref}} - \mathcal{Z}_{\text{GFM}}(s)i_{\text{h}}$ and represented as Thevenin's equivalent model connected across a current source as shown in Fig. 2.6(d). For an example, at nominal plant condition with resulting optimal controller, the Bode plot of $\mathcal{G}_{\text{GFM}}(s)$ and $\mathcal{Z}_{\text{GFM}}(s)$ are shown in Fig. 2.8(a) and Fig. 2.8(b) respectively. It is observed that $|\mathcal{G}_{\text{GFM}}(j\omega_{\text{N}})|$, $\angle \mathcal{G}_{\text{GFM}}(j\omega_{\text{N}})$ and $|\mathcal{Z}_{\text{GFM}}(j\omega_{\text{N}})|$ are ≈ 1 , $\approx 0^\circ$ and ≈ 0 respectively that leads to $v_{\text{O}} \approx v_{\text{ref}}$ at fundamental frequency.

2.6 Experimental Results and Verification

2.6.1 Experimental Configuration

A combined system-in-the-loop (SIL), controller hardware-in-the-loop (CHIL) and power hardware-in-the-loop (PHIL) based experimental validation is conducted in order to evaluate the efficacy and viability of the proposed μ -synthesis-based controller for single-phase GFL and GFM inverters. The ratings and parameters of the inverter systems are tabulated in Table 5.4. Moreover, the efficiency curve of the inverter system under study operating at 240 volt RMS is shown in Fig. 2.9. The laboratory-based experimental setup is shown in Fig. 2.10. The experimental configuration is shown in Fig. 2.11.

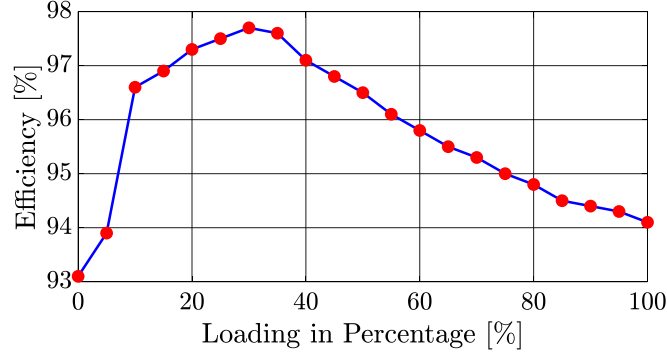


Figure 2.9: Efficiency curve of the 1- ϕ , 60 Hz, 1.67 kVA inverter under study operating at 240 volt RMS.

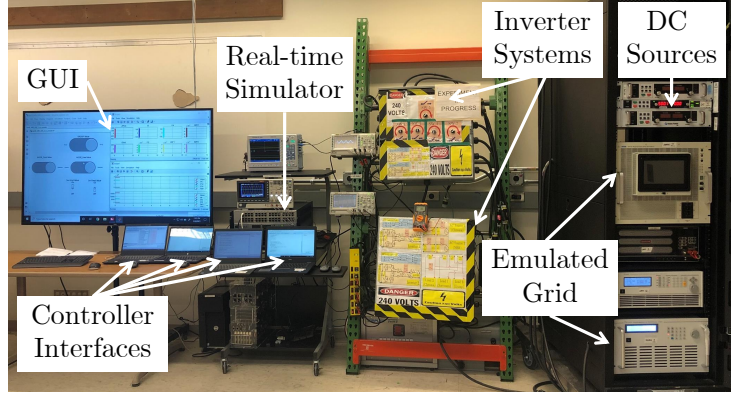


Figure 2.10: Figure of the laboratory-based experimental hardware setup.

Real-time Simulation and SIL Configuration

A residential sub-network of North American low voltage distribution feeder from CIGRE Task Force C6.04.02 [74], affiliated with CIGRE Study Committee C6 is emulated using eMEGASIM platform inside the OP5700 RT-simulator (RTS) manufactured by OPAL-RT. The original ratings of load at each bus and line parameters are modified in order to make it compatible with the voltage rating and power capacity of the laboratory. Moreover, the test system is modified by including sufficient non-linear loads at various buses while respecting the recommended limits of harmonic distortions mentioned in [73]. Usually real-time simulation applied to the domain of power engineering is classified into two categories: 1) fully real-time simulation (e.g., system-in-the-loop

Table 2.6: 1-phase Inverter System Under Study

Inverter	Value
Ratings ($1-\phi$)	240 V, 60 Hz, 1.67 kVA, 0.9 pf
Inverter Parameters	$V_{dc} = 500$ V, $f_{sw} = 20$ kHz
Filter Parameters	$L_f = 2$ mH, $R_f = 0.2$ Ω , $C_f = 20$ μ F

(SIL)), and 2) hardware-in-the-loop (HIL) real-time simulation. A fully real-time simulation (or SIL) requires the entire system (including control, protection, and other accessories) to be modeled inside the simulator and does not involve external interfacing or inputs/outputs [75]. As part of SIL-setup, one GFM and one GFL inverter are emulated entirely (i.e. both power circuit and the control with proposed μ -synthesis-based controller) inside the RTS, connected at Bus₁ and Bus₁₂ respectively as shown in Fig. 2.11.

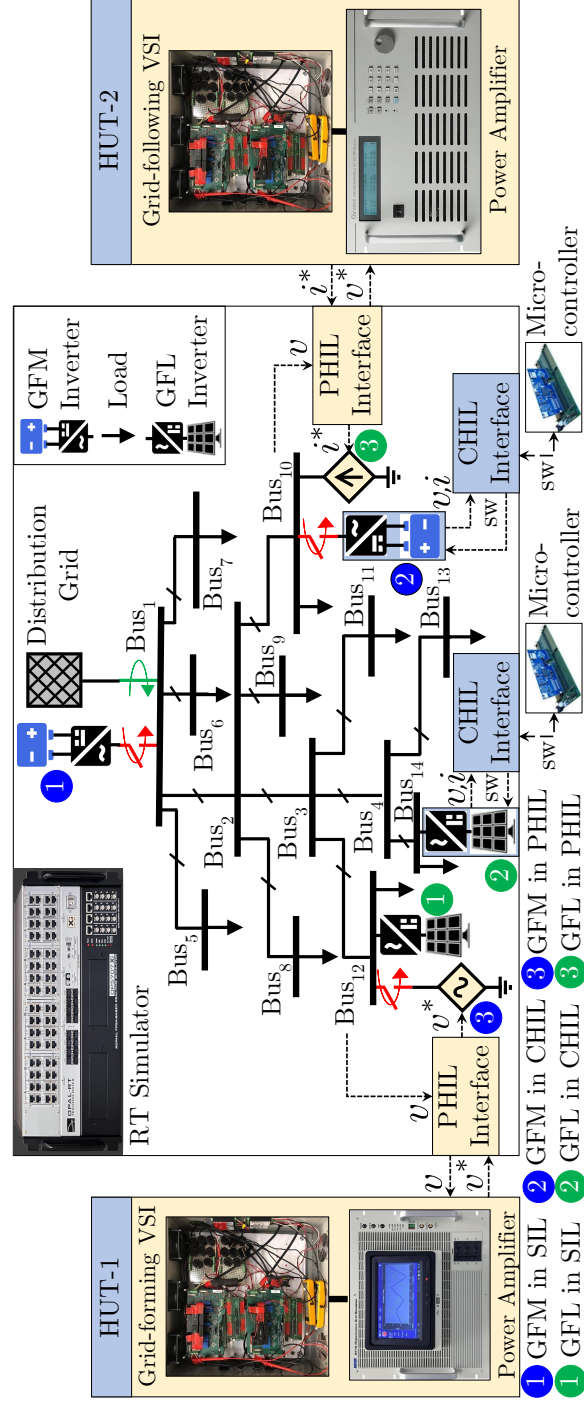


Figure 2.11: Figure of the configuration of SIL-CHIL-PHIL-based experimental setup.

Controller Hardware-in-the-loop Configuration

This stage of validation is used in this work for rapid controller prototyping with the proposed μ -synthesis based robust controller for both GFL and GFM inverter systems. As part of CHIL-setup, one GFM and one GFL inverter system are emulated with only power circuit inside the RTS, connected at Bus₁₀ and Bus₁₄ respectively as shown in Fig. 2.11. The proposed μ -synthesis-based control logic of both GFL and GFM inverter systems are realized on two Texas-Instruments TMS320F28379D, 16/12-bit floating-point 200 MHz Delfino micro-controller boards interfaced with RTS.

Power Hardware-in-the-loop Configuration

This stage of validation is used in this work for real hardware validation with the proposed μ -synthesis based robust controller for both GFL and GFM inverter systems. As part of PHIL-setup, one GFM (HUT-1 in Fig. 2.11) and one GFL inverter (HUT-2 in Fig. 2.11), connected at Bus₁₂ and Bus₁₀ respectively, are physically realized. In HUT-1 the physical inverter system, fed by MAGNA-POWER programmable DC power supply (configured with user-defined I – V characteristic of battery cell), is interfaced with low-cost Texas-Instruments TMS320F28379D, 16/12-bit floating-point 200 MHz Delfino micro-controller boards employed with proposed μ -synthesis control logic for GFM inverter with “Overload Mitigation Controller” based overcurrent protection scheme [76]. On the other hand, in HUT-2 the physical inverter system, fed by another MAGNA-POWER programmable DC power supply (configured with user-defined I – V characteristic of solar photovoltaic array), is interfaced with another low-cost Texas-Instruments TMS320F28379D, 16/12-bit floating-point 200 MHz Delfino micro-controller boards employed with proposed μ -synthesis control logic for GFL inverter with “Instantaneous Current Limiter” based overcurrent protection scheme [77]. The ideal transformer model (ITM) based PHIL interface logic [78] is adapted for both the hardware-under-tests (HUTs’). The power terminals of the inverters are connected with power amplifiers. NHR 9410 regenerative grid simulator and Chroma 61605 programmable ac power source are used as the power amplifiers. The sensors to measure inverter output current are realized by AEMC MN-260 current probe. Moreover, a diode-bridge rectifier based load, realized using a “Chroma 63800 Series AC-DC Electronic Load Bank”, is also considered as non-linear load under test.

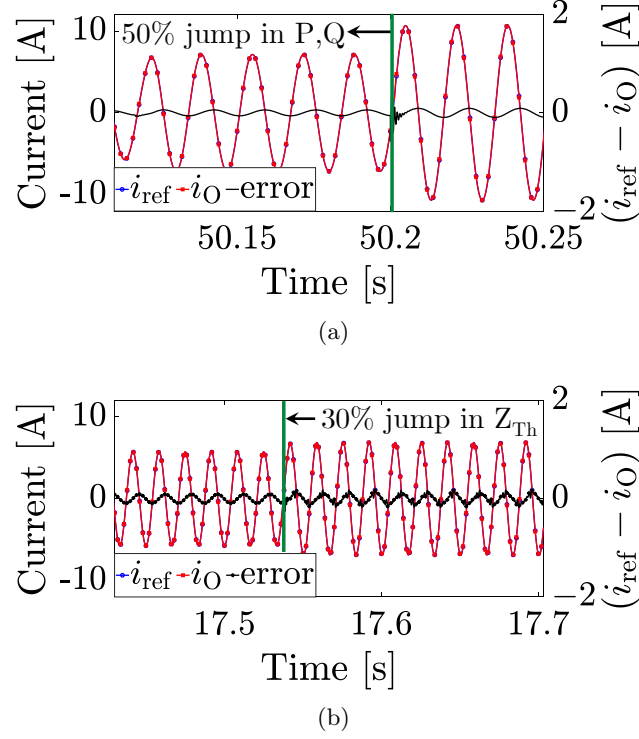


Figure 2.12: The current reference tracking capability of GFL inverter with proposed μ -synthesis-based optimal controller in CHIL setup at Bus₁₄ of Fig. 2.11 in (a) CASE-1, (b) CASE-2.

2.6.2 Results and Discussions

Four test cases (two test cases each for GFL and GFM inverters) are demonstrated by emulating a sequence of events. Two test cases for GFL inverters are as follows:

- **CASE-1:** The emulated distribution network of Fig. 2.11 is running in off-grid mode and P-Q reference of GFL inverters jumps up by 50% due to increased demand.
- **CASE-2:** The network is running in off-grid mode and experiences a topology change which results in 30% increase in equivalent Thevenin impedance at PCC of GFM inverters.
- **CASE-3:** The emulated distribution network of Fig. 2.11 has an on-grid to off-grid mode transition. GFM inverters will have a maximum jump in loading from no-load condition (during on-grid mode) to full-load condition (during off-grid mode).
- **CASE-4:** The same network has an off-grid to on-grid mode transition. GFM inverters will have another maximum jump in loading from full-load condition (during off-grid

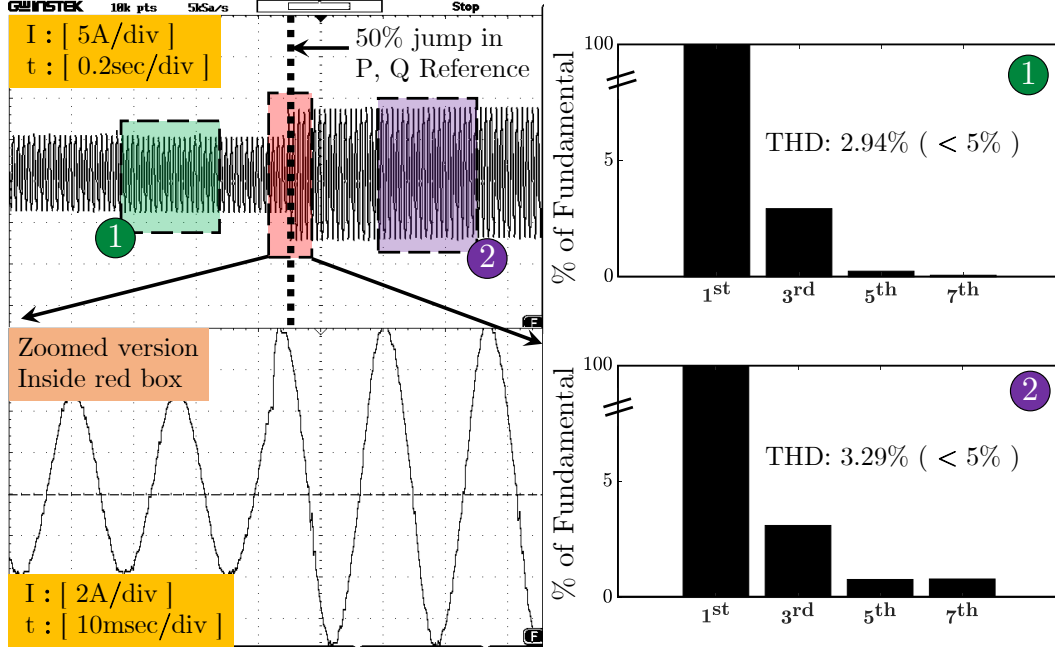


Figure 2.13: The current reference tracking capability of GFL inverter in PHIL setup at Bus₁₀ of Fig. 2.11 in CASE-1 with the proposed optimal controller.

mode) to no-load condition (during on-grid mode).

Clearly, CASE-1, CASE-2 and CASE-3, CASE-4 are designed in order to capture the robust performance of proposed GFL and GFM control during maximal model uncertainty respectively. Fig. 2.12(a) and Fig. 2.12(b) show the current reference tracking capability of the proposed μ -synthesis-based optimal controller for GFL inverter (at Bus₁₄ of Fig. 2.11) in CHIL demonstration of CASE-1 and CASE-2 respectively. RMS instantaneous current tracking error (rICTE in %), defined as $100 \times \text{RMS}(i_{\text{ref}} - i_O) / [\sqrt{2} \text{RMS}(i_{\text{ref}})]$, is used for assessing the tracking performance of the current controller. It is observed in Fig. 2.12(a) that both the current reference and output current increases during 50% increase in P-Q set-points due to adopted reference generation of Appendix A.1. The proposed optimal controller has significantly small error in current reference tracking before (rICTE $\approx 1.7\%$) and after (rICTE $\approx 1.9\%$) the transition in CASE-1. Similarly, Fig. 2.12(b) shows that the proposed optimal controller has significantly small error in current reference tracking before (rICTE $\approx 1.9\%$) and after (rICTE $\approx 2.1\%$) the jump of equivalent Thevenin impedance in CASE-2. Fig. 2.13 shows the current response of GFL inverter (HUT-2 at Bus₁₀ of Fig. 2.11) as a part of PHIL demonstration of the

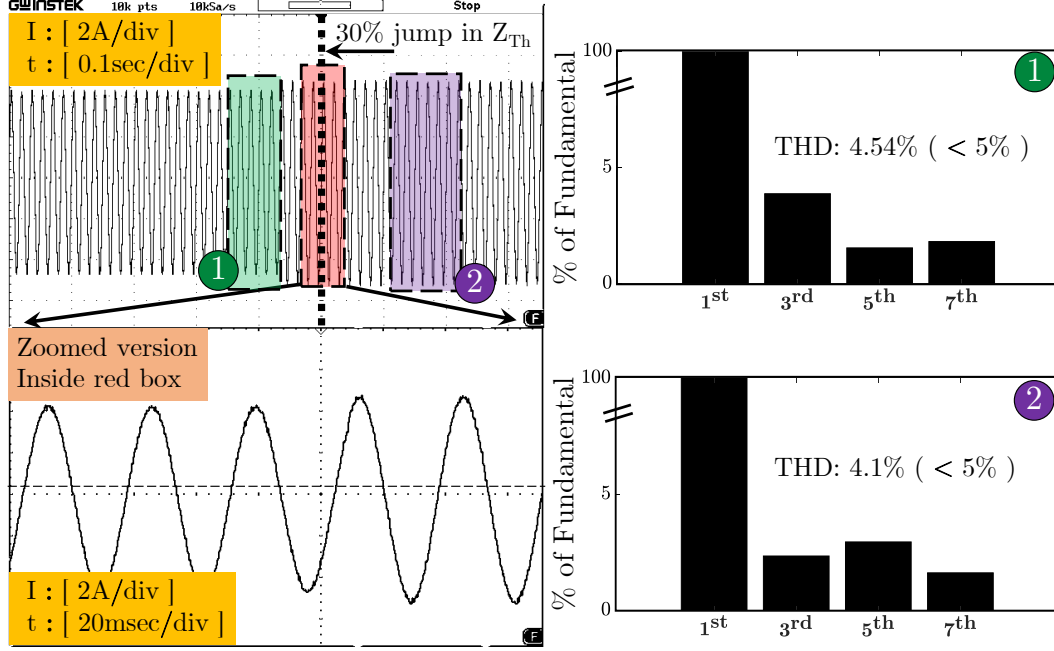


Figure 2.14: The current reference tracking capability of GFL inverter in PHIL setup at Bus₁₀ of Fig. 2.11 in CASE-2 with the proposed optimal controller.

same event. Here the result is focused on determining the harmonic compensation capability of the proposed optimal controller for GFL inverter during varying reference set-point. It is observed that the total demand distortion (TDD) of current waveform is $< 5\%$ before and after the transition as recommended in [73]. Thus, the proposed μ -synthesis-based controller for GFL inverter shows good reference tracking and harmonic compensation capability in CASE-1. Similarly, Fig. 2.14 shows the current response of GFL inverter (HUT-2 at Bus₁₀ of Fig. 2.11) as a part of PHIL demonstration of the same event. Here the result is focused on determining the harmonic compensation capability of the proposed μ -synthesis-based optimal controller for GFL inverter during model uncertainty. It is observed here that the TDD of current waveform is $< 5\%$ before and after the transition as recommended in [73]. Thus, the data corroborates the efficacy of the proposed μ -synthesis-based controller for GFL inverter for good reference tracking and harmonic compensation capability in CASE-2. The CHIL and PHIL results substantiate the fact that the proposed μ -synthesis-based optimal controller for GFL is showing robust performance by making sure to have good reference tracking, disturbance rejection and harmonic compensation capability under significant in the

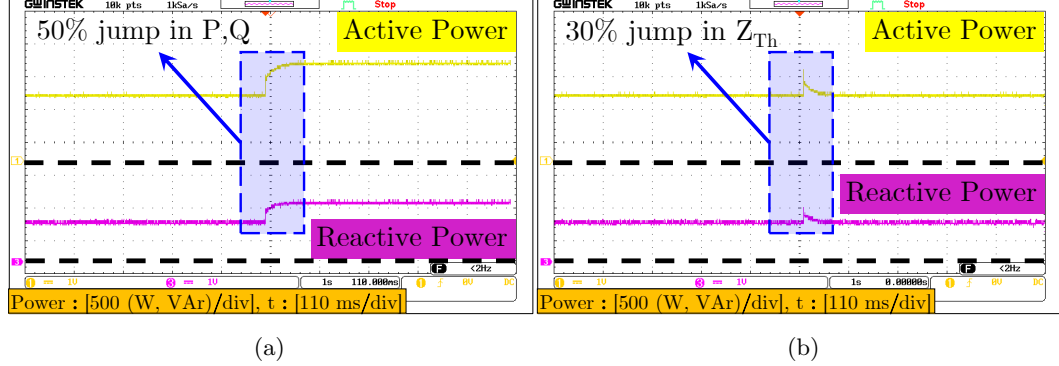


Figure 2.15: The output active and reactive power of the GFL inverter with proposed μ -synthesis-based optimal controller in PHIL setup at Bus₁₀ of Fig. 2.11 in (a) CASE-1 (b) CASE-2.

plant model uncertainty while supplying reference active and reactive power as shown in Fig. 2.15(a) and Fig. 2.15(b).

Fig. 2.16(a) and Fig. 2.16(b) show the voltage reference tracking capability of the proposed μ -synthesis-based optimal controller for GFM inverter (at Bus₁₀ of Fig. 2.11) in CHIL demonstration of CASE-3 and CASE-4 respectively. RMS instantaneous voltage tracking error (rIVTE in %), defined as $100 \times \text{RMS}(v_{\text{ref}} - v_o) / [\sqrt{2} \text{RMS}(v_{\text{ref}})]$, is used for assessing the tracking performance of the voltage controller. It is observed in Fig. 2.16(a) that both the voltage reference and output voltage drop during jump in loading (no-load to full-load) due to adopted droop-controlled reference generation of Appendix A.2. The proposed μ -synthesis-based optimal controller has significantly small error in voltage reference tracking as shown in Fig. 2.16(a) before (rIVTE $\approx 0.2\%$) and after (rIVTE $\approx 0.1\%$) the increase of equivalent loading. Similarly, it is observed in Fig. 2.16(b) that both the voltage reference and output voltage, increase during drop in loading (full-load to no-load). The proposed optimal controller has significantly small error in voltage reference tracking before (rIVTE $\approx 0.2\%$) and after (rIVTE $\approx 0.1\%$) the decrease in equivalent loading. Fig. 2.17 shows the voltage response of GFM inverter (HUT-1 at Bus₁₂ of Fig. 2.11) as a part of PHIL demonstration of the same event. Here the result is focused on determining the harmonic compensation capability of the proposed optimal controller for GFM inverter during model uncertainty change due to loading. It is observed that the total harmonic distortion (THD) of voltage waveform is

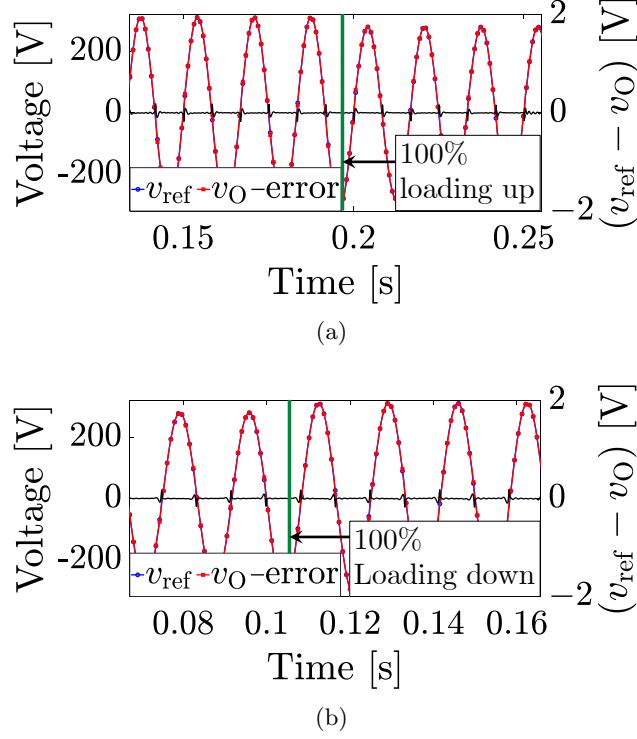


Figure 2.16: The voltage reference tracking capability of GFM inverter with the proposed μ -synthesis-based optimal controller in CHIL setup at Bus₁₀ of Fig. 2.11 in (a) CASE-3 and (b) CASE-4.

$< 3\%$ before and after the transition. This is significantly less than the voltage distortion limit ($< 8\%$) as recommended in [73]. Thus the data corroborates the advantage of the proposed μ -synthesis-based controller for GFM inverter shows robust performance during model uncertainty caused in CASE-3. Similarly, Fig. 2.18 shows the voltage response of GFM inverter (HUT-1 at Bus₁₂ of Fig. 2.11) as a part of PHIL demonstration of the same event. Here the result is focused on determining the harmonic compensation capability of the proposed optimal controller for GFM inverter during model uncertainty change due to loading. It is observed that the total harmonic distortion (THD) of voltage waveform is $< 3\%$ before and after the transition. This is significantly less than the voltage distortion limit ($< 8\%$) as recommended in [73]. Thus GFM inverter shows robust performance during model uncertainty caused in CASE-4. The CHIL and PHIL results substantiate the fact that the proposed optimal controller for GFM is showing

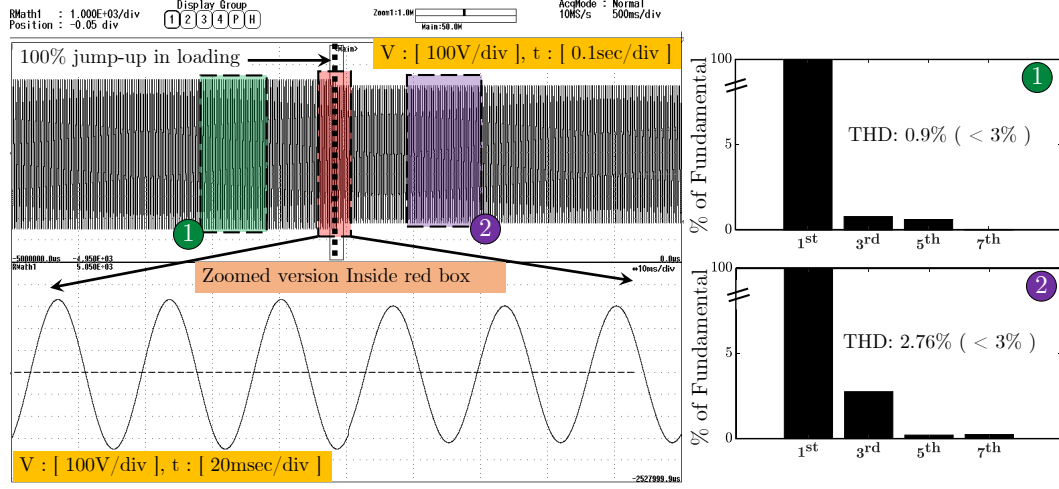


Figure 2.17: The voltage reference tracking capability of GFM inverter in PHIL setup at Bus₁₂ of Fig. 2.11 in CASE-3 with the proposed optimal controller.

robust performance by having good reference tracking, disturbance rejection and harmonic compensation capability under significant in the plant model uncertainty while sharing the active and reactive power demand as shown in Fig. 2.19(a) and Fig. 2.19(b).

An experimental study on the inverter systems are conducted to validate the plug-and-play (PnP) performance of the inverter while operating in GFM mode with the proposed μ -synthesis-based optimal voltage controller and GFL mode with the proposed μ -synthesis-based optimal current controller. Accredited standards like [71, 79, 80] provides a basic guidelines for testing PnP capabilities of GFM and GFL inverter in a microgrid scenario. In PnP testing, the following two test scenarios are validated in the same test system for the inverters of HUT-1 and HUT-2:

- the performance during plugging-in of a GFL inverter system at the point-of-connection with the microgrid network after a proper synchronization and during successive start by injecting power into the microgrid network as requested,
- the performance during starting and then plugging-in of the GFM inverter at the point-of-connection with the loads in the microgrid network after transitioning from on-grid to off-grid mode.

Figure 2.20 shows the voltage and current waveform of the GFL inverter (HUT-2) with the proposed μ -synthesis-based optimal current controller during the PnP operation. At time $t = t_1$, as highlighted in Figure 2.20, the GFL inverter starts synchronization process with respect to the voltage waveform (yellow colored waveform) at

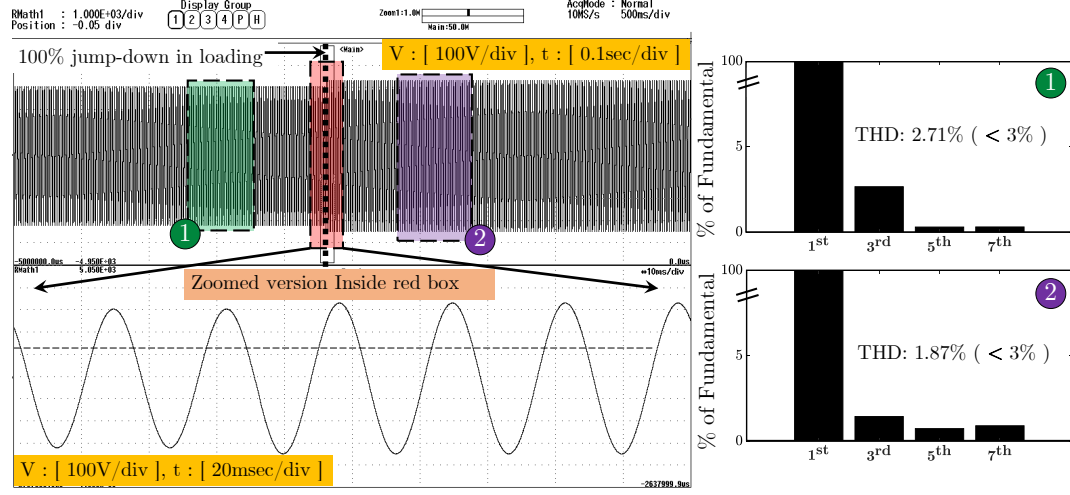


Figure 2.18: The voltage reference tracking capability of GFM inverter in PHIL setup at Bus₁₂ of Fig. 2.11 in CASE-4 with the proposed optimal controller.

point-of-connection and within 3-4 cycles the synchronization is achieved by aligning the inverter voltage (cyan colored waveform) with the grid voltage. At $t = t_2$ the GFL inverter is synchronized and plugged-in with the grid and starts injecting power into the network. Similarly, Figure 2.21 shows the voltage and current waveform of the GFM inverter (HUT-1) with the proposed μ -synthesis-based optimal voltage controller during the PnP operation. At time $t = t_1$, as highlighted in Figure 2.21, the GFM inverter starts from black-start and generates smooth and stable voltage waveform across its terminal (cyan colored waveform). It is observed that after having a ramp up time of 2-3 cycles, the GFM inverter generates a stable voltage across its terminal. At $t = t_2$ the GFM inverter is plugged-in with the loads in the microgrid and starts injecting power into the network. These PnP operations are achieved in the absence of any sort of communication among the inverters and the microgrid as recommended in the standards.

Moreover, it is important to validate the performance of the GFL inverter with the proposed μ -synthesis-based optimal current controller while the inverter is interfaced with a weak grid. An experimental study is conducted on the GFL inverter (HUT-2) when it is interfaced with a weak grid, having low short circuit ratio ($SCR < 2$) emulated by OPAL-RT and programmable voltage source. Figure 2.22 shows the voltage and current waveform of the GFL inverter. During this experiment, the point-of-connection of the emulated grid has a transition from strong grid to weak grid condition as observed

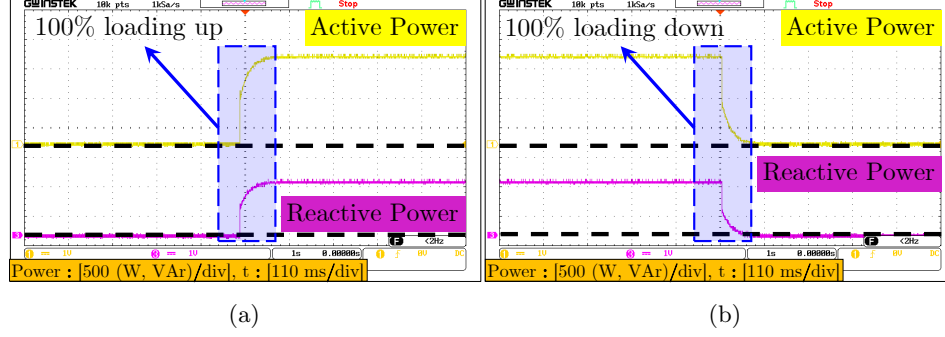


Figure 2.19: The output active and reactive power of the GFM inverter with proposed μ -synthesis-based optimal controller in PHIL setup at Bus₁₂ of Fig. 2.11 in (a) CASE-3 (b) CASE-4.

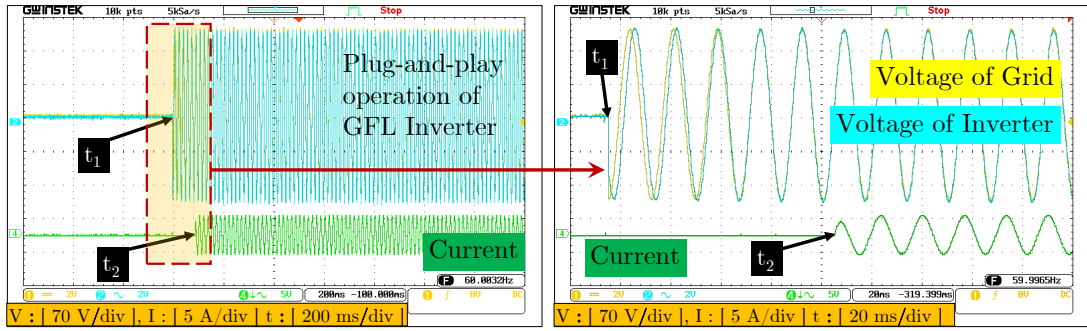


Figure 2.20: Experimental results showcasing the plug-and-play performance of the GFL inverter with the proposed μ -synthesis-based current controller.

in the voltage waveform of the grid (the cyan colored waveform). It is observed here that the current output of the GFL inverter with the proposed optimal controller have less impact on the controller performance during, before and after the transition from strong grid to weak grid condition. This corroborates the fact that the proposed μ -synthesis-based current controller is robust enough in its performance while experiencing a wide range of variation in grid impedance.

2.6.3 Performance Comparison

To showcase the advantages of the proposed μ -synthesis-based controller on robust performance, a CHIL-based performance comparison study is conducted with two controllers of the existing literature. The existing current controllers used in this study for

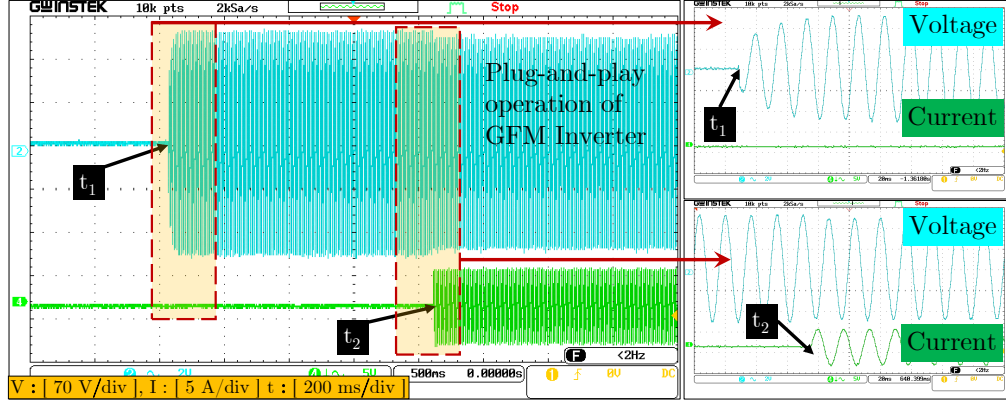


Figure 2.21: Experimental results showcasing the plug-and-play performance of the GFM inverter with the proposed μ -synthesis-based voltage controller.

comparison purposes are, 1) conventional classical proportional+resonant (PR)-based controller that can be found in many commercial grid-following inverter systems, and 2) an nominal $\mathcal{H}_\infty$ -based current controller considering a fixed grid impedance modeling (i.e. modeling of uncertainty imposed by the variation of the grid impedance is not considered). This method is adopted from the existing work of reference [42]. Similarly the existing voltage controllers used in this study for comparison purposes are, 1) conventional classical PR-based controller that can be found in many commercial grid-forming inverter systems, and 2) an nominal $\mathcal{H}_\infty$ -based voltage controller considering a fixed equivalent loading modeling (i.e. the modeling of the uncertainty imposed by variation of equivalent loading is not considered). This method is adopted from the existing work of reference [55]. Fig. 2.23(a) and Fig. 2.23(b) show the current reference tracking capability of the $\mathcal{H}_\infty$ -based and PR-based current controller for GFL inverter (at Bus₁₄ of Fig. 2.11) in CHIL demonstration of CASE-2 respectively. It is observed that the nominal $\mathcal{H}_\infty$ -based robust controller has significant error in current reference tracking (rICTE increases from $\approx 2.5\%$ to $\approx 10\%$) after 30% jump in equivalent Thevenin impedance from nominal value. It is observed that the PR-based controller has comparatively larger error both before (rICTE $\approx 6\%$) and after (rICTE $\approx 10\%$) the transition of CASE-2. In comparison, the proposed μ -synthesis-based optimal controller has significantly small error in current reference tracking before (rICTE $\approx 1.9\%$) and after (rICTE $\approx 2.1\%$) the jump of equivalent Thevenin impedance in CASE-2 as shown in Fig. 2.12(b). Similarly, Fig. 2.24(a) and Fig. 2.24(b) show the voltage reference tracking capability of the $\mathcal{H}_\infty$ -based and PR-based voltage controller for GFM inverter (at Bus₁₀ of Fig. 2.11) in

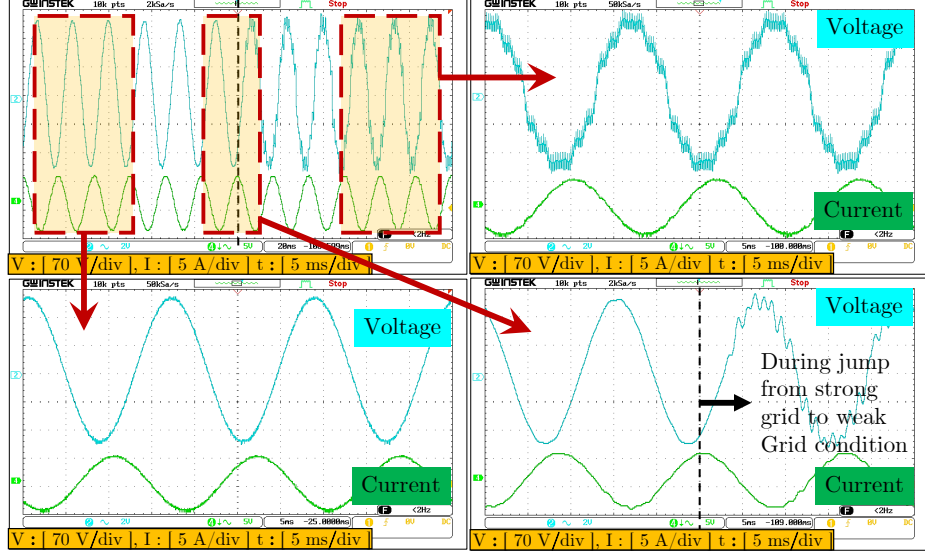


Figure 2.22: Experimental results showcasing the performance of the GFL inverter with the proposed μ -synthesis-based optimal current controller while connected with weak grid network.

CHIL demonstration of **CASE-3** respectively. It is observed that the nominal $\mathcal{H}_\infty$ -based robust controller has significant error in voltage reference tracking (rIVTE $\approx 1.2\%$) at no-load condition. The PR-based controller has comparatively larger error both before (rIVTE $\approx 1.4\%$) and after (rIVTE $\approx 0.8\%$) the transition of **CASE-3**. In comparison, the proposed μ -synthesis-based optimal controller has significantly small error in voltage reference tracking before (rIVTE $\approx 0.2\%$) and after (rIVTE $\approx 0.1\%$) the increase of equivalent loading as shown in Fig. 2.16(a).

Moreover, an experimental study is conducted to compare the dynamic and steady-state performance of the proposed μ -synthesis-based optimal voltage controller for GFM inverter under diode-based nonlinear load with respect to the existing nominal $\mathcal{H}_\infty$ -based robust controller and classical PR-based controller. The diode bridge rectifier-based nonlinear load of the form as shown in Figure 2.25 is considered in this study. Figure 2.26, Figure 2.27 and Figure 2.28 show the voltage and current waveform of the GFM inverter with the proposed μ -synthesis-based, nominal $\mathcal{H}_\infty$ -based, and classical PR-based voltage controller. It is observed that the GFM inverter is able to maintain a smooth and steady voltage waveform across the nonlinear load leveraged by the enhanced reference tracking and disturbance rejection capability of the proposed

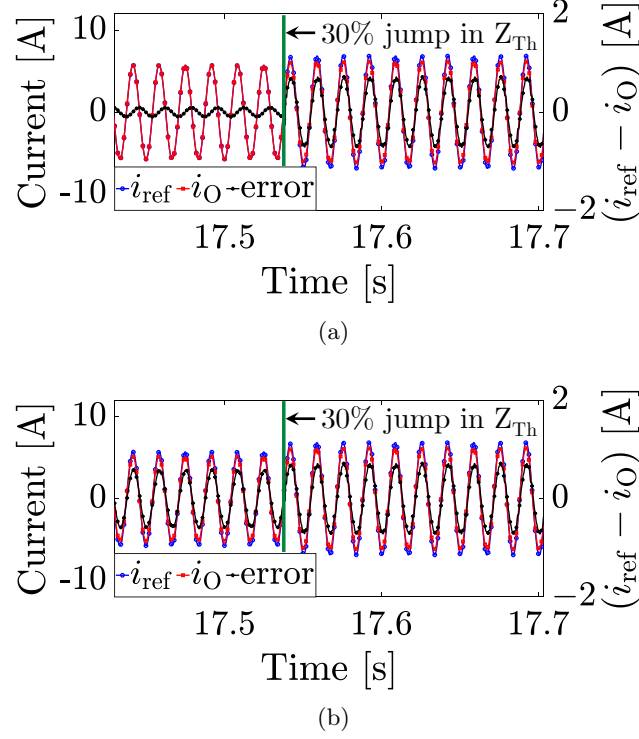


Figure 2.23: Performance comparison of the current reference tracking capability of GFL inverter in CHIL setup at Bus₁₄ of Fig. 2.11 with, (a) $\mathcal{H}_\infty$ -based controller without considering uncertainty of Thevenin impedance, (b) the classical PR controller.

voltage controller in comparison with the existing voltage controllers. The change in the plant model (plant uncertainty) due to the presence of non-linear load across the GFM inverter has comparatively less impact on the performance of the proposed voltage controller compared to the existing controllers as observed in the results. The total harmonic distortion (THD) of the voltage waveform are $\approx 2.65\%$, $\approx 9.1\%$ and $\approx 11.7\%$ for proposed μ -synthesis-based, nominal $\mathcal{H}_\infty$ -based and PR-based voltage controllers respectively after the nonlinear load transition.

Similarly, an experimental study is conducted to compare the steady-state performance of the proposed μ -synthesis-based optimal current controller for GFL inverter under distorted grid voltage condition with respect to the existing nominal $\mathcal{H}_\infty$ -based robust controller and classical PR-based controller. Figure 2.29, Figure 2.30 and Figure 2.31 show the voltage and current waveform of the GFL inverter with the proposed μ -synthesis-based, nominal $\mathcal{H}_\infty$ -based, and classical PR-based current controller. It is

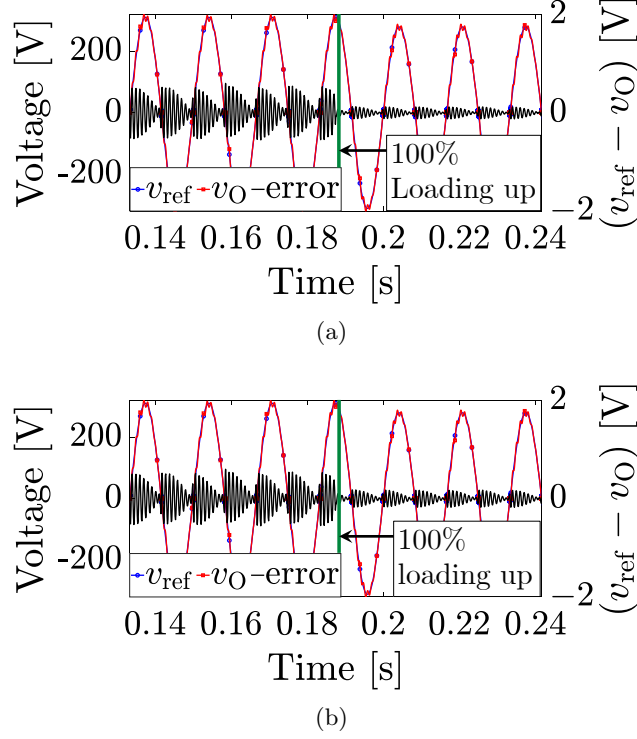


Figure 2.24: Performance comparison of the voltage reference tracking capability of GFM inverter in CHIL setup at Bus₁₀ of Fig. 2.11 with, (a) $\mathcal{H}_\infty$ -based controller without considering uncertainty of equivalent loading, (b) the classical PR controller.

observed that the GFL inverter is able to generate and inject a fairly smooth and steady sinusoidal current into the grid while interfaced with a distorted grid voltage. The proposed controller is able to nullify the impact of the distortion of the grid voltage on the current output, leveraged by the enhanced reference tracking and disturbance rejection capability of the proposed current controller under large variation of grid impedance and grid voltage distortion in comparison with the existing current controllers. The total demand distortion (TDD) of current waveform is $\approx 3.4\%$, $\approx 7.5\%$ and $\approx 9.4\%$ for proposed μ -synthesis-based, nominal $\mathcal{H}_\infty$ -based and PR-based current controller respectively with the distorted grid voltage.

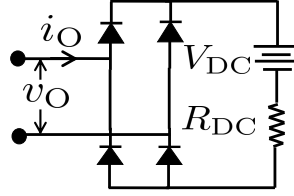


Figure 2.25: Schematic of the diode-bridge-based nonlinear load.

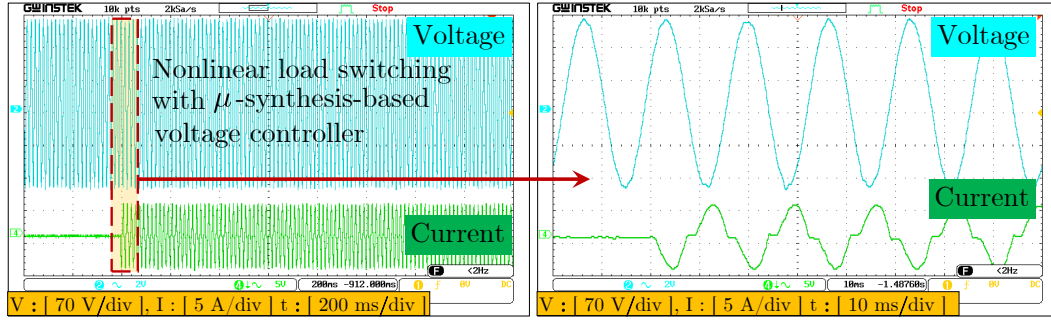


Figure 2.26: Experimental results showcasing the dynamic and steady-state performance of the proposed μ -synthesis-based optimal voltage controller employed in the grid-forming inverter terminated at diode-bridge-based nonlinear load.

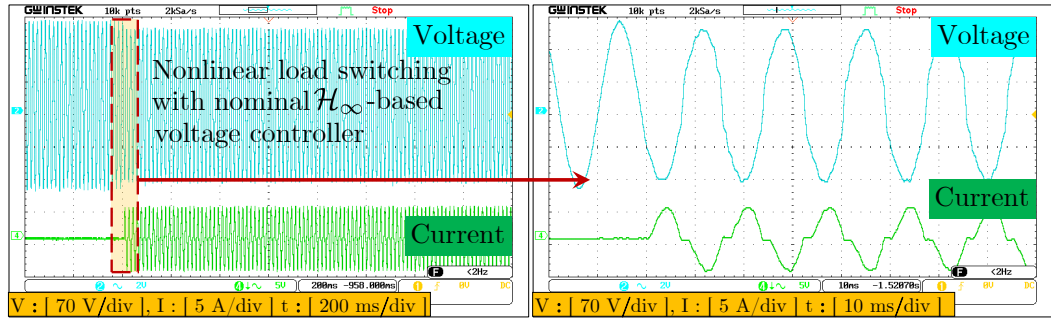


Figure 2.27: Experimental results showcasing the dynamic and steady-state performance of the existing nominal $\mathcal{H}_\infty$ -based voltage controller employed in the grid-forming inverter terminated at diode-bridge-based nonlinear load.

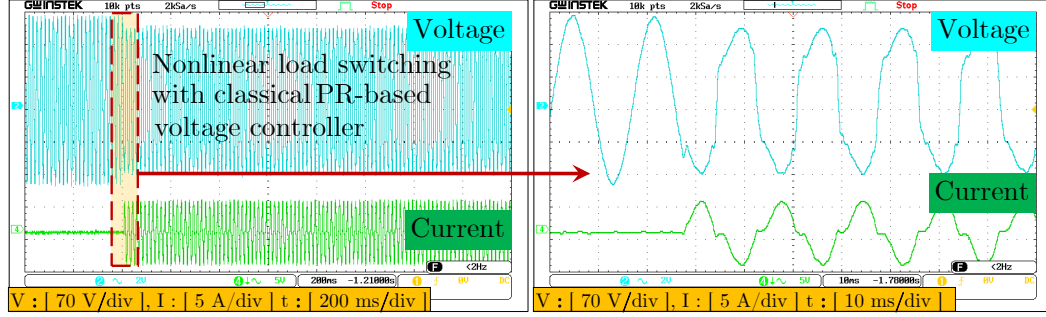


Figure 2.28: Experimental results showcasing the dynamic and steady-state performance of the existing PR-based voltage controller employed in the grid-forming inverter terminated at diode-bridge-based nonlinear load.

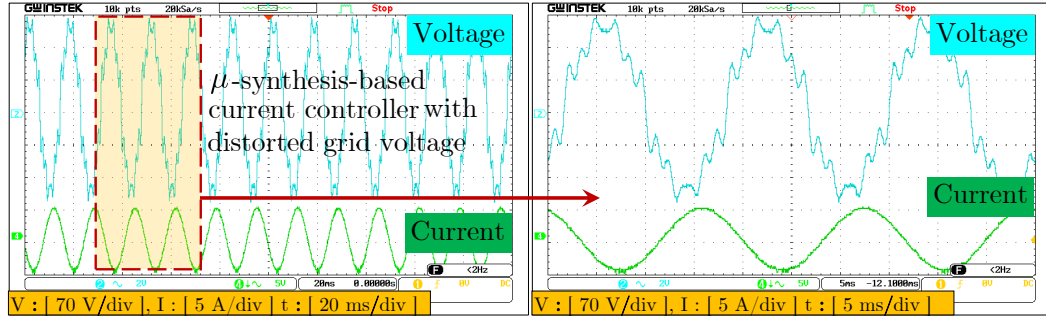


Figure 2.29: Experimental results showcasing the performance of the proposed μ -synthesis-based optimal current controller employed in the grid-following inverter terminated at grid with distorted voltage.

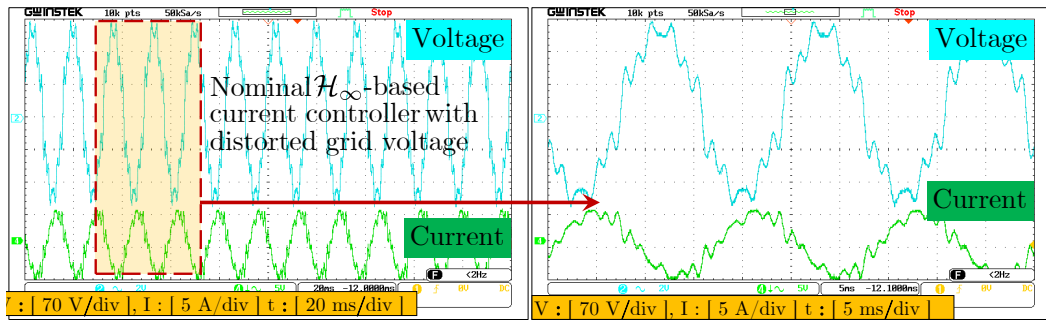


Figure 2.30: Experimental results showcasing the performance of the existing nominal $\mathcal{H}_\infty$ -based current controller employed in the grid-following inverter terminated at grid with distorted voltage.

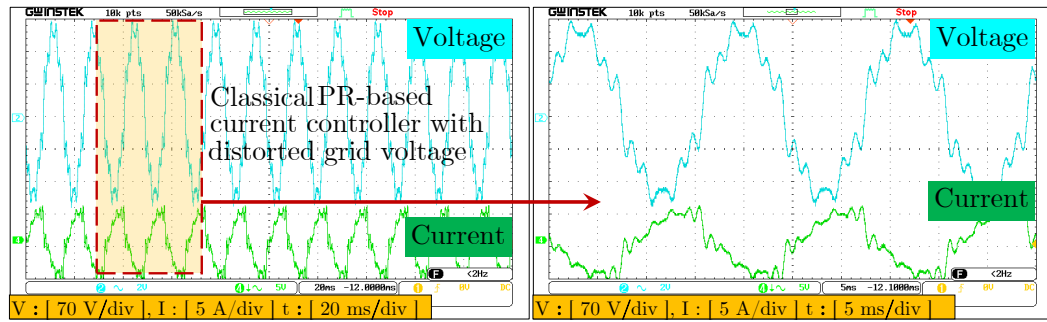


Figure 2.31: Experimental results showcasing the performance of the existing PR-based current controller employed in the grid-following inverter terminated at grid with distorted voltage.

Chapter 3

Dynamic Microgrid: Novel Seamless Transition Using Grid-forming Inverters

3.1 Gist of the Chapter

Seamless recovery of power to critical infrastructures, after grid failure, is a crucial need arising in scenarios that are increasingly becoming more frequent. This chapter describes a seamless transition strategy using a single and unified mode-dependent droop controlled grid-forming inverters. The control strategy achieves the following objectives: 1) regulates the output active and reactive power by the droop-controlled inverters to a desired value while operating in on-grid mode; 2) seamless transition and recovery of power injections into the load after grid failure by inverters that operates in grid-forming mode all the time; 3) requires only a single bit of information on the grid/network status for the mode transition. A framework for assessing the stability of the system and to guide the choice of parameters for controllers is developed using control-oriented modeling. A controller hardware-in-the-loop-based real-time simulation study on a test system based on the realistic electrical network of a commercial-scale medical center is conducted for initial prototyping of the control strategy. A hardware experiment is conducted with two 3- ϕ , 480-V, 125-kVA grid-forming inverters, a 3- ϕ , 480-V, 270-kVA grid simulator, a physical grid switch, and a physical load bank.

Note: More details on this work can be found in the published articles [81, 82]

3.2 State-of-the-art on Seamless Transition Schemes

Table 3.1: A summary of major works on seamless transition of inverter system.

Ref.	Pros and Cons
[83, 84, 85]	• Knowledge of magnitude, type of grid impedance and coupling impedances of inverters are prerequisite. • Inclusion of phase – locked loop dynamics in droop law impacts overall stability by grid disturbances.
[86, 87, 88, 89, 90]	• Distributed or centralized master – slave based mode – supervisory control suffers from loss of autonomy and independent nature of operation. • Single or multiple point failure and undesired delays in communication causes vulnerability in applications
[91, 92, 93, 94, 95, 96]	• Addition of multiple new control layers for seamless transition purpose only burdens the overall control and requires tuning of additional control parameters. • Emphasized specifically on re – synchronization method of single inverter system with incoming grid.
[97, 98, 99, 100, 101, 102] [103, 104, 105, 106, 107, 108]	• Single inverter is employed with dual mode control. where current control mode for on – grid and voltage control mode for off – grid mode of operation. • Large deviations in voltage/current of inverters due to switching of mode restricts smooth transition.
Proposed	✓ Single and unified voltage control architecture able to operate both in on – grid and off – grid mode. ✓ Guaranteed regulation over active, reactive power to desired value during on – grid mode of operation. ✓ Mode dependant droop with ease in implementation enables seamless transition with reduced transients.

Critical infrastructure (CI) describes essential assets of society that include medical centers and security service centers. Disruption of power to CIs often result in a debilitating impact on physical and economic security, public health and safety. Rapid and seamless recovery of power to restore CIs online, possibly after a power blackout caused by weather/climate disasters, is a crucial need arising in scenarios that are increasingly becoming more frequent [109]. IEEE 602 recommends Type-U emergency

power supply system (EPSS) by using local generation that enable CIs to have a seamless on-grid to off-grid and off-grid to on-grid transition with an uninterruptible power supply to its critical loads during such contingencies [110]. Gas/diesel generator-based traditional EPSS experiences difficulties in achieving such seamless transition due to its long startup time from standby. NFPA 111 recommends stored-energy EPSS (SEPSS) using battery storage-based grid-forming (GFM) inverter that provides an alternate solution with better response time of EPSS for CIs [111]. To achieve seamless transitions and uninterrupted supply to loads, GFM inverters are required to be synchronized and connected all the time with the network of the CI irrespective of the availability of main grid, unlike plug-and-play methods [112]. In addition, SEPSS needs to guarantee that the GFM inverters do not supply power in on-grid mode in order to maintain sufficient reserves of energy in battery storage units for emergency off-grid operation. Thus in on-grid mode, all of the power to the CI is to be supplied only by the grid while the GFM inverters remain on standby to enable a seamless transition to off-grid mode in case of grid interruption [113, 111]. Moreover, SEPSS is required to ensure that the GFM inverters maintain a stable voltage and frequency while meeting the power demand locally only in off-grid mode and realize a seamless off-grid to on-grid transition when grid returns. Droop-controlled GFM inverters provide an autonomous and communication-less solution for network in an off-grid mode operation. However, during the on-grid mode, the voltage and frequency of the network are governed by the stiff grid and as a result the control over output active and reactive power of the droop-controlled GFM inverter becomes challenging [10]. In addition, a control strategy needs to be employed in GFM inverter systems for seamless transition of the network between on-grid to off-grid mode and vice versa.

A comprehensive overview of various control existing schemes, enabling seamless transition and operation in on-grid and off-grid mode for microgrid, is provided in [114] as summarized in Table 6.1. Reference [83, 84] proposes an adaptive droop control suitable for both on-grid and off-grid modes of operation. However, knowledge of magnitude, type of grid impedance and coupling impedances of inverters are prerequisite for the control strategy [83] which may not be practical. Moreover, inclusion of phase-locked loop dynamics in the droop law has a serious impact on overall stability of the control influenced by the grid disturbances [84]. Reference [85] modifies the droop control law to achieve operation in on-grid mode; here, the prime focus is to inherit the advantages of the droop controller to limit the inverter current under both normal and faulty conditions. Master-slave-based architecture in multi-inverter

systems (electrically closest inverter to grid as master and rests as slaves) is proposed in [86, 87, 88, 89] both for on-grid and off-grid modes. However, here, a coordinated architecture is required which compromises autonomy and independent nature of operation. A distributed mode-supervisory control between multiple GFM inverters and the grid is proposed in [90] to avail both on-grid and off-grid operation and seamless transition capability. Even though the distributed architecture reduces communication burden, multiple-point failure and undesired delays in communication makes the solution vulnerable in applications. A traditional inverter control with additional layer to generate reference signals, enabling seamless mode transition for GFM inverters, is proposed in [91, 92, 93]. However, the additional control layer and its parameter tuning for restoring voltage deviations and synchronize phase to grid before connection/reconnection makes the solution difficult to realize in practical control boards. Reference [94] proposes a virtual impedance-based loop in the universal droop control layer to achieve seamless transfer from the off-grid to on-grid mode. However, [94] is more focused on the synchronization mechanism and doesn't provide explicit strategies for control over active and reactive power of GFM inverters in on-grid mode. Reference [95] proposes a virtual torque and virtual flux linkage-based synchronizing technique, suitable for seamless transition of synchronous inverters from off-grid to on-grid. [96] proposes a pre-synchronization unit-based transition method where inverters can transfer from grid current control to droop control in case of on-grid to off-grid transition and vice versa. However, a dedicated inverter-based unit is required to be employed in the system for the seamless transition purpose which is a costly solution. Dual mode operation capability (i.e. grid-following operation in on-grid and grid-forming operation in off-grid mode) is the typical strategy employed to meet seamless transition objectives [97]. Prior state-of-the-art on dual-mode control architecture provide methods to minimize fluctuations in phase, frequency, and voltage amplitude of the network during the transitions [98, 99, 100, 101, 102, 103, 104, 105, 106, 107, 108]. However, large deviations in the output voltage/current of inverters due to switching of its operating mode are drawbacks that restrict a smooth transition and can affects the stability.

In the current state-of-the-art, there is a lack of unified and single control architecture that enables inverter to; 1) operate in GFM mode always, irrespective of grid availability, with guaranteed regulation of output active and reactive power in on-grid mode and operate autonomously in off-grid mode, and 2) exhibit seamless transition capability between on-grid to off-grid and vice versa. This chapter proposes a novel

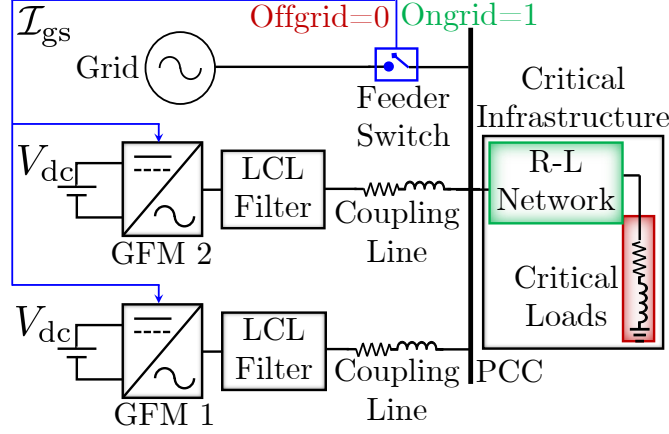


Figure 3.1: Diagram of the system of a critical infrastructure under study.

mode-dependent droop control framework for GFM inverters to achieve these functionalities with reduced communication burden and ease in implementability in practice. The key features of the control framework are; 1) guaranteed regulation over output active and reactive power of the GFM inverters to the desired value (for example, zero active and reactive power sourced by GFM inverters) while in on-grid mode; 2) a fast response time of recovery by GFM inverter with seamless transition when the main grid fails; 3) requirement of minimal information of grid status (as a single 0/1 bit) for the seamless mode transition.

3.3 Description of the System and Motivation

Here, the physical system, employed in this work, is described to instantiate the application scenario. It comprises a CI as shown in Fig. 3.1 with multiple critical loads that need continuous uninterrupted power supplied irrespective of the grid availability as defined in NFPA 111 [111]. At the emergency bus (PCC), multiple sources are connected in parallel. The grid, interfaced with feeder switch, is the main power supply in the on-grid mode of the CI. Two inverters (GFM-1 and GFM-2), operating in grid-forming mode with proposed mode-dependent droop control as described in Section 6.4, are always synchronized and connected to the PCC by coupling lines irrespective of grid availability. The flow of the proposed strategy for SEPSS of a CI is as follows:

- In on-grid mode, feeder switch is closed and status of the switch, $\mathcal{I}_{gs} = 1$, is transmitted to GFM-1 and GFM-2 of Fig. 3.1 that enables the SEPSS not to

de-energize the batteries. Hence, grid alone supplies the entire load.

- In case of a grid failure, feeder switch is open and status of the switch, $\mathcal{I}_{gs} = 0$, is transmitted to GFM-1 and GFM-2 of Fig. 3.1 that enables the SEPSS to ensure that supply to the critical load demand is met by GFM-1 and GFM-2.

Design objectives of the controller are:

- Provide seamless power recovery where inverters are always connected to the CI network while SEPSS ensures not to de-energize these energy sources in on-grid mode.
- During on-grid mode, CI is required to be supported only by the grid and the critical load demands need to be met.
- In case of a grid failure, SEPSS is required to ensure that the critical load demands are met by the local inverters.

3.4 Design of the Proposed Droop Controller

The power circuit of i^{th} , 3- ϕ H-bridge inverter consists of six switches distributed among three legs as shown in Fig. 3.2. It is connected to the network at PCC with voltage, $v_{\text{PCC}}^{\text{abc}}$, via an LCL filter ($L_{f,i}$, $C_{f,i}$, $L_{g,i}$) and associated equivalent series resistances ($R_{f,i}$ and $R_{g,i}$ of inductors) and a coupling line with line parameters, $L_{\text{line},i}$, $R_{\text{line},i}$. A dq-frame multi-loop controller is employed that generates modulated voltage vector signal, m_i^{abc} , to pulse-width modulation (PWM) controller to generate switching signals resulting in terminal voltages, $v_{t,i}^a$, $v_{t,i}^b$ and $v_{t,i}^c$. The control loop of Fig. 3.2 is described below.

3.4.1 Power Controller

$v_{o,i}^{\text{dq}}$, and $i_{o,i}^{\text{dq}}$ are used to determine p_i , and q_i . p_i and q_i are passed through low-pass filters with the time constant, $\tau_{S,i} \in \mathbb{R}_{>0}$, to obtain P_i and Q_i as described by

$$P_i = [1/(\tau_{S,i}s + 1)]p_i, \quad Q_i = [1/(\tau_{S,i}s + 1)]q_i, \quad (3.1)$$

where $p_i := 1.5[v_{o,i}^d i_{o,i}^d + v_{o,i}^q i_{o,i}^q]$, $q_i := 1.5[v_{o,i}^q i_{o,i}^d - v_{o,i}^d i_{o,i}^q]$.

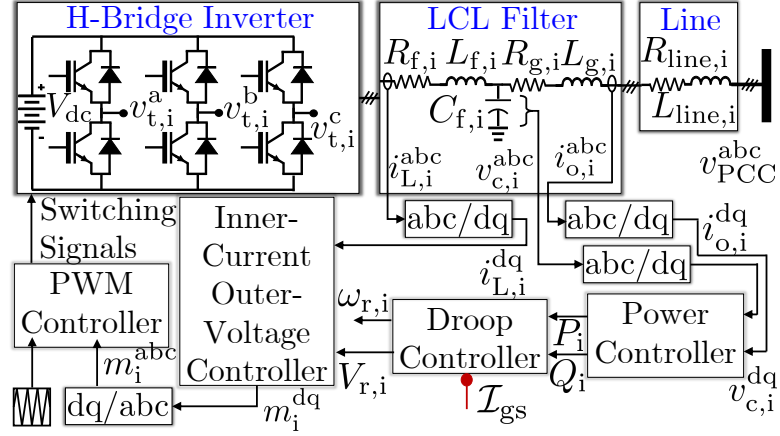


Figure 3.2: Power circuit and control loops of i^{th} GFM inverter.

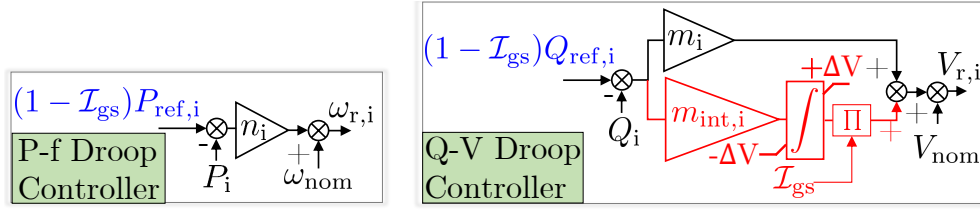


Figure 3.3: The proposed P - f and Q - V droop for seamless transition.

3.4.2 Droop Controller

Droop control is a proportional controller with active and reactive power as control variables where the control gains (also the droop gains) dictate the steady-state power sharing of the inverters. This chapter proposes a mode-dependent droop controller for i^{th} inverter for both on-grid and off-grid operation of the CI as shown in Fig. 3.3. The active power, frequency, P - f , droop control is considered here as a proportional controller (with proportional coefficient as n_i) with error signal $e_{P,i} := (1 - \mathcal{I}_{gs})P_{\text{ref},i} - P_i$ where P_i is the control variable and $(1 - \mathcal{I}_{gs})P_{\text{ref},i}$ is the reference. Whereas, the reactive power, voltage magnitude, Q - V , droop control is considered here as a proportional-integral controller (with proportional and integral coefficients as m_i and $m_{\text{int},i}$ respectively) with error signal $e_{Q,i} := (1 - \mathcal{I}_{gs})Q_{\text{ref},i} - Q_i$ where Q_i is the control variable and $(1 - \mathcal{I}_{gs})Q_{\text{ref},i}$ is the reference. The additional integral action in Q - V droop is effective only in on-grid mode which is ensured by multiplication of $\mathcal{I}_{gs}$ with the integral part. In summery, the

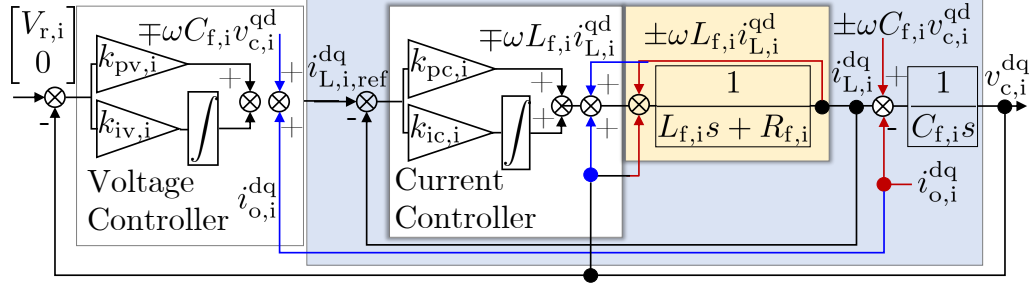


Figure 3.4: The conventional inner-current-outer-voltage controller.

proposed droop law is as follows:

$$\omega_{r,i} = \omega_{\text{nom}} - n_i [P_i - (1 - \mathcal{I}_{\text{gs}}) P_{\text{ref},i}], \quad (3.2)$$

$$V_{r,i} = V_{\text{nom}} - m_i [Q_i - (1 - \mathcal{I}_{\text{gs}}) Q_{\text{ref},i}] - \mathcal{I}_{\text{gs}} m_{\text{int},i} \psi_i^Q, \quad (3.3)$$

$$\psi_i^Q = \int [Q_i - (1 - \mathcal{I}_{\text{gs}}) Q_{\text{ref},i}] dt, \quad (3.4)$$

The proposed droop control law differs from conventional droop characteristics in the following way:

- Unlike *conventional* droop, an additional integral action is introduced in Q - V droop while on-grid. This results in a proportional controller for active and proportional-integral controller for reactive power with $(1 - \mathcal{I}_{\text{gs}}) P_{\text{ref},i}$ and $(1 - \mathcal{I}_{\text{gs}}) Q_{\text{ref},i}$ as reference signals respectively.
- Status of the feeder switch, $\mathcal{I}_{\text{gs}}$, is included in the droop law that makes it mode dependent. The function of $\mathcal{I}_{\text{gs}}$ is to modify the droop law based on the transition from on-grid ($\mathcal{I}_{\text{gs}} = 1$) to off-grid mode ($\mathcal{I}_{\text{gs}} = 0$).

ψ_i^Q guarantees that the inverters in on-grid mode to supply no reactive power in steady-state. This is further described in Section 3.6.A where it is established that in steady state the GFM inverters will not source any active or reactive power while on the on-grid operation. The addition of dependency on the variable, $\mathcal{I}_{\text{gs}}$, facilitates the inverters seamless functionality for CI during the transition of on-/off-grid and off-/on-grid modes. The values of n_i and m_i are typically chosen such that $\omega_{r,i}$ and $V_{r,i}$ are within the allowed specification, defined by IEEE 1547 Standard [71], for all $P_i \in [0, P_{\text{rated},i}]$ and $Q_i \in [-Q_{\text{rated},i}, Q_{\text{rated},i}]$ respectively. Here, $P_{\text{rated},i}$ and $Q_{\text{rated},i}$ are the rated active and reactive powers that can be delivered by the i^{th} inverter. The saturation limit

in the integrator of the Q - V droop law ensures that the entire system remains viable under compromised scenarios including significant delay in communicating $\mathcal{I}_{gs}$ which results in undesirable active integral operation during off-grid mode. A good candidate for selecting the magnitude of the upper and lower saturation limit of the integrator, $\pm\Delta V$, is $\pm 10\%$ of the nominal voltage rating as recommended by limits of IEEE 446 [115]. Although, these empirical upper bounds of droop co-efficient facilitate the initial design of droop law, system stability-constraint bounds of n_i , m_i and $m_{int,i}$ require special attention due to the system interconnection and its seamless transition between on-grid and off-grid mode of operation. Rigorous modeling and analysis of the system is required in order to assess the system stability and the power quality during transitions, and the relation of the performance goals with the control parameters.

3.4.3 Inner-current-outer-voltage Controller

The conventional inner-current-outer-voltage controller architecture is employed for the inverters [116] as shown in Fig. 3.4. For the inner-current controller, $i_{L,i,ref}^{dq}$ is provided as the reference signal to be tracked by the output signal, $i_{L,i}^{dq}$. A proportional-integral (PI) compensator is used for tracking the reference of the dq-axis inductor current. For a desired time constant, $\tau_{c,i}$, the parameters of the current controller are selected as $k_{pc,i} = L_{f,i}/\tau_{c,i}$ and $k_{ic,i} = R_{f,i}/\tau_{c,i}$. Depending on the switching frequency, $\tau_{c,i}$ is typically selected to be in the range of 0.5-2 ms [116]. Additional feed-forward signals, $v_{c,i}^{dq}$ and $\mp\omega L_{f,i}i_{L,i}^{qd}$ facilitate the disturbance rejection capability. For outer-voltage controller, $\begin{bmatrix} V_{r,i} & 0 \end{bmatrix}^T$ is the reference signal to be tracked by the voltage signal, $v_{c,i}^{dq}$. A PI compensator is used to enable reference tracking. For a desired phase margin and gain cross-over frequency, the parameters of the voltage controller ($k_{pv,i}$ and $k_{iv,i}$) can be designed based on *symmetrical optimum* method [116]. Similarly, additional feed-forward signals, $v_{o,i}^{dq}$ and $\mp\omega C_{f,i}v_{c,i}^{qd}$ facilitate the disturbance rejection capability for the outer voltage control loop.

3.5 Modeling of System Components

The system described in Section 6.3 is considered in this section for modeling and analysing. By defining $L_{l,i} := L_{g,i} + L_{line,i}$ and $R_{l,i} := R_{g,i} + R_{line,i}$ for $i = 1, 2$, model of individual inverter will be developed in this section. The grid is assumed to be stiff; commensurate with a real-world power system network. The grid parameters are $L_{l,g}$

and $R_{l,g}$. It is to be noted that $L_{g,i}$ and $R_{g,i}$ are design parameters, whereas $L_{line,i}$ and $R_{line,i}$ are uncertain line parameters. Here $L_{g,i}$ is chosen to be large enough such that the impedance between the capacitor of the LCL filter and the PCC is largely inductive. Moreover, the entire building network of CI along with loads are represented by an equivalent lumped load (R_L and L_L) terminated at PCC, for the sake of stability analysis. For the ease in analysis, the following assumptions are made:

- The inner-current-outer-voltage control loop is stable and has faster (> 10 times) dynamics when compared to the power and droop controller and tracks its voltage reference with minimal (assumed zero) tracking error as suggested in [116].
- Grid is modeled as a voltage source with voltage, V_{nom} , and frequency, ω_{nom} , and a Z-type load is considered [117].
- All abc-dq conversions are adopted w.r.t. a common reference with phase angle, θ_{ref} , and frequency, ω_{ref} .

It allows to model a GFM inverter as a voltage source with controllable phase, frequency and amplitude and the system in Fig. 3.1 can be represented equivalently by Fig. 3.5.

3.5.1 Modelling a Single Inverter in Off-grid Mode

The following non-linear state-space equations can be obtained for i^{th} inverter with $\mathcal{I}_{gs} = 0$ for off-grid mode:

$$\dot{\theta}_i = \omega_{r,i} - \omega_{ref}, \text{ where } \theta_i := \theta_{r,i} - \theta_{ref}, \quad (3.5)$$

$$\tau_{S,i} \dot{\omega}_{r,i} = \omega_{nom} - \omega_{r,i} - n_i[p_i - P_{ref,i}], \quad (3.6)$$

$$\tau_{S,i} \dot{V}_{r,i} = V_{nom} - V_{r,i} - m_i[q_i - Q_{ref,i}], \quad (3.7)$$

$$L_{l,i} \dot{i}_{o,i}^d = V_{r,i} \cos \theta_i - R_{l,i} i_{o,i}^d - v_{PCC}^d + \omega_{ref} L_{l,i} i_{o,i}^q, \quad (3.8)$$

$$L_{l,i} \dot{i}_{o,i}^q = V_{r,i} \sin \theta_i - R_{l,i} i_{o,i}^q - v_{PCC}^q - \omega_{ref} L_{l,i} i_{o,i}^d, \quad (3.9)$$

where, $p_i := 1.5[V_{r,i} i_{o,i}^d \cos \theta_i + V_{r,i} i_{o,i}^q \sin \theta_i]$, and $q_i := 1.5[-V_{r,i} i_{o,i}^q \cos \theta_i + V_{r,i} i_{o,i}^d \sin \theta_i]$. This results in a 5th-order non-linear electromagnetic model for the droop-controlled inverter operating in off-grid mode.

3.5.2 Modelling a Single Inverter in on-grid Mode

The following non-linear state-space equations can be obtained for i^{th} inverter with $\mathcal{I}_{\text{gs}} = 1$ for on-grid mode:

$$\dot{\theta}_i = \omega_{r,i} - \omega_{\text{ref}}, \text{ where } \theta_i := \theta_{r,i} - \theta_{\text{ref}}, \quad (3.10)$$

$$\tau_{S,i} \dot{\omega}_{r,i} = \omega_{\text{nom}} - \omega_{r,i} - n_i p_i, \quad (3.11)$$

$$m_i \dot{\psi}_i^Q = V_{\text{nom}} - V_{r,i} - m_{\text{int},i} \psi_i^Q. \quad (3.12)$$

$$\tau_{S,i} \dot{V}_{r,i} = k_{m,i} (V_{\text{nom}} - V_{r,i} - m_{\text{int},i} \psi_i^Q) - m_i q_i, \quad (3.13)$$

and (3.8) and (3.9), where, $k_{m,i} := 1 - \tau_{S,i} m_{\text{int},i} / m_i$. This results in a 6th-order non-linear electromagnetic model for the droop-controlled inverter operating in on-grid mode.

3.5.3 Modelling the Stiff Grid

The grid is modeled by the following non-linear equations:

$$\dot{\theta}_g = \omega_{\text{grid}} - \omega_{\text{ref}}, \text{ where } \theta_g := \theta_{\text{grid}} - \theta_{\text{ref}}, \quad (3.14)$$

$$L_{l,g} \dot{i}_g^d = V_g \cos \theta_g - R_{l,g} i_g^d - v_{\text{PCC}}^d + \omega_{\text{ref}} L_{l,g} i_g^q, \quad (3.15)$$

$$L_{l,g} \dot{i}_g^q = V_g \sin \theta_g - R_{l,g} i_g^q - v_{\text{PCC}}^q - \omega_{\text{PCC}} L_{l,g} i_g^d. \quad (3.16)$$

This results in a 3rd-order non-linear electromagnetic model for the stiff grid.

3.5.4 Modelling the Load at PCC

The following dynamics equations are derived for the load:

$$\begin{aligned} L_L \dot{i}_L^d &= v_{\text{PCC}}^d - R_L i_L^d + \omega_{\text{ref}} L_L i_L^q, i_L^d = \mathcal{I}_{\text{gs}} i_g^d + i_{o,1}^d + i_{o,2}^d \\ L_L \dot{i}_L^q &= v_{\text{PCC}}^q - R_L i_L^q - \omega_{\text{ref}} L_L i_L^d, i_L^q = \mathcal{I}_{\text{gs}} i_g^q + i_{o,1}^q + i_{o,2}^q \end{aligned}$$

Using (3.8), (3.9), (3.15) and (3.16) along with few algebraic manipulations, v_{PCC}^d and v_{PCC}^q can be represented by non-linear functions of θ_g , θ_i , V_g , $V_{r,i}$, i_g^d , i_g^q , $i_{o,i}^d$, $i_{o,i}^q$ for $i = 1, 2$.

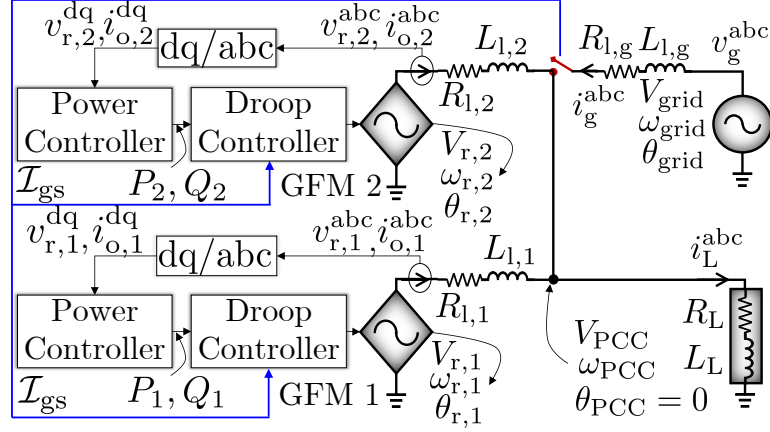


Figure 3.5: Equivalent circuit model of the system of Fig. 3.1.

3.6 Analysis of System Model and Stability

3.6.1 Analysis of System Operating in on-grid Mode

While the system of Fig. 3.5 is operating in on-grid mode, it can be modelled by a nonlinear system, $\dot{\underline{x}}_{\text{on}} = \underline{\mathcal{G}}_{\text{on}}(\underline{x}_{\text{on}})$, where, $\underline{\mathcal{G}}_{\text{on}}(\cdot)$ consists of (3.10), (3.11), (3.12), (3.13), (3.8), (3.9) for $i = 1, 2$ and (3.14), (3.15), (3.16). Here,

$$\underline{x}_{\text{on}}^{\top} = [\theta_1 \ \theta_2 \ \theta_g \ \omega_{r,1} \ \omega_{r,2} \ V_{r,1} \ V_{r,2} \ \psi_1^Q \ \psi_2^Q \ i_{o,1}^d \ i_{o,2}^d \ i_g^d \ i_{o,1}^q \ i_{o,2}^q \ i_g^q]. \quad (3.17)$$

System Model in on-grid condition with $\mathcal{I}_{\text{gs}} = 1$ is shown in Appendix B.1. As a result, the complete system in on-grid mode is a 15th order nonlinear autonomous system with state vector $\underline{x}_{\text{on}}$.

Theorem 1. *Consider the system as shown in Fig. 3.5 operating in on-grid mode where both the inverters are following droop laws given in (3.2), (3.3) and (3.4). In steady-state, both the inverters will supply zero active and reactive output power with the grid supplying the entire demand of the system.*

Proof. While the system, shown in Fig. 3.5, is operating in on-grid mode, i.e. $\mathcal{I}_{\text{gs}}=1$ and grid is assumed to be at nominal, the complete system model is formulated by $\dot{\underline{x}}_{\text{on}} = \underline{\mathcal{G}}_{\text{on}}(\underline{x}_{\text{on}})$. Therefore, in steady-state, $\underline{\mathcal{G}}_{\text{on}}(\underline{x}_{\text{on}})|_{\underline{x}_{\text{on}}^{\text{eq}}} = 0$. Hence, (3.5) with $i=1, 2$ and (3.14) yield $\omega_{r,1} = \omega_{r,2} = \omega_{\text{grid}} = \omega_{\text{nom}}$. However, from (3.11) for $i=1, 2$, $\omega_{r,1} = \omega_{\text{nom}} - n_1 p_1$ and $\omega_{r,2} = \omega_{\text{nom}} - n_2 p_2$. Hence, it follows $n_1 p_1 = n_2 p_2 = 0$ which implies $p_1 = p_2 =$

0. Similarly, (3.12) for $i=1, 2$ yields $V_{r,1} = V_{\text{nom}} - m_{\text{int},1}\psi_1^Q$ and $V_{r,2} = V_{\text{nom}} - m_{\text{int},2}\psi_2^Q$. Moreover, from (3.13) for $i=1, 2$ it follows that $m_1q_1 = k_{m,1}(V_{\text{nom}} - V_{r,1} - m_{\text{int},1}\psi_1^Q)$ and $m_2q_2 = k_{m,2}(V_{\text{nom}} - V_{r,2} - m_{\text{int},2}\psi_2^Q)$. As a result, $m_1q_1 = m_2q_2 = 0$ which implies $q_1 = q_2 = 0$. Hence, active and reactive power of both the inverters are zero and grid alone supplies the total load demand of the CI. $\square$

Based on the Theorem 1, following remarks can be made:

- Inverters with droop law stated in (3.2), (3.3) and (3.4) do not need to de-energize and can remain connected to the system, while the grid is available to supply all the electrical loads.
- Seamless power recovery can be achieved to CI while grid is disconnected as inverters are always synchronized to the system and no transfer-switching process is required.
- While the same system is operating in on-grid mode without the integrator-loop in Q - V droop, it can be shown mathematically that active power sourced by both the inverters will be zero. However, reactive power can be non-zero and the value will depend on line parameters and droop coefficients. Due to limited space, the derivation is not shown here.

Remark 1. *The capability, where the GFM inverter imports power from grid to recharge the battery during on-grid mode after supplying power to the loads in off-grid mode, can be achieved by minor change in the droop laws of (3.2) as follows:*

$$\omega_{r,i} = \omega_{\text{nom}} - n_i[P_i - (1 - \mathcal{I}_{\text{gs}})P_{\text{ref},i} + \mathcal{I}_{\text{gs}}\mathcal{I}_{\text{soc}}P_{\text{chr},i}],$$

where, $P_{\text{chr},i}$ is the power rating at which the i^{th} battery can be recharged. $\mathcal{I}_{\text{soc}}$ is an indicator that determines whether battery connected to the GFM inverter needs to be charged or not. $\mathcal{I}_{\text{soc}} = 1$ signifies that the battery needs charging and $\mathcal{I}_{\text{soc}} = 0$ signifies that the battery does not need charging. The modification is only in the P - f droop law and the Q - V droop law is same. Strategies for generating $\mathcal{I}_{\text{soc}}$ can be employed locally or globally and kept for further research, as it is out of scope of the current work.

Clearly, the proposed technique adds another control parameter i.e. the integrator coefficient, that needs to be tuned. Considering the system linearized around an equilibrium

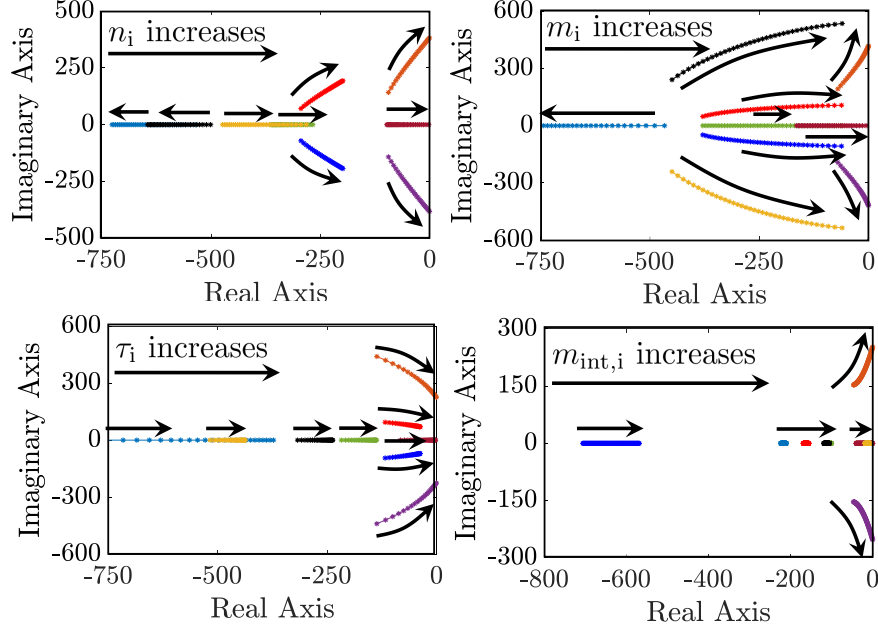


Figure 3.6: Eigenvalue spectrum of $\mathbf{A}_{\text{on}}^s$ for range of parameters in Table 3.2.

Table 3.2: Domain of Control Parameters for $i = 1, 2$

Parameters	Range	Selected Value
n_i (rad/s/kW)	$[1.04 \ 3.64] \times 10^{-2}$	2.5×10^{-2}
m_i (volt/kVAr)	$[41.67 \ 416.6] \times 10^{-3}$	192×10^{-3}
$m_{\text{int},i}$ (volt/s/kVAr)	$[0 \ 0.04]$	0.01
τ_i (ms)	$[24.7 \ 41.2]$	33

point, $\underline{x}_{\text{on}}^{\text{eq}}$ such that $\Delta \dot{\underline{x}}_{\text{on}} = \mathbf{A}_{\text{on}} \Delta \underline{x}_{\text{on}}$, where $\mathbf{A}_{\text{on}} = \mathcal{F}_{\text{on}}(\underline{x}_{\text{on}}^{\text{eq}})$ and $\mathcal{F}_{\text{on}}(\cdot)$ is the vector field of $\underline{\mathcal{G}}_{\text{on}}$, a systematic framework is developed for tuning the control parameters i.e. $n_1, n_2, m_1, m_2, m_{\text{int},1}, m_{\text{int},2}, \tau_1, \tau_2$, so that asymptotic stability is guaranteed. The linearized model of the system is further reduced by suppressing the EM dynamics of the lines and loads. The assumption made here is that the characteristic EM timescale for lines and loads are slower ($\sim 3\text{-}5$ ms [117]) and are below the characteristic timescale of droop controls ($\sim 30\text{-}40$ ms [118]), dictated by the choice of the parameter, $\tau_{\text{S},i}$. This strong time-scale separation in the system between the slow states denoted as $\underline{x}_{\text{on}}^{\text{s}} := [\theta_1 \ \theta_2 \ \theta_g \ \omega_{r,1} \ \omega_{r,2} \ V_{r,1} \ V_{r,2} \ \psi_1^{\text{Q}} \ \psi_2^{\text{Q}}]^{\top}$, and the fast states denoted as $\underline{x}_{\text{on}}^{\text{f}} := [i_{\text{o},1}^{\text{d}} \ i_{\text{o},2}^{\text{d}} \ i_{\text{g}}^{\text{d}} \ i_{\text{o},1}^{\text{q}} \ i_{\text{o},2}^{\text{q}} \ i_{\text{g}}^{\text{q}}]^{\top}$, allows the assumption that the fast states reach

close to their quasi-steady-state values (i.e. $\underline{x}_{\text{on,ss}}^f$), much faster than the slow states [119]. Therefore, only remaining slow states are of interest from the stability point of view of the reduced-order model of the system of given as $\Delta \dot{\underline{x}}_{\text{on}}^s = \mathbf{A}_{\text{on}}^s \Delta \underline{x}_{\text{on}}^s$. With the equilibrium point of this slow model and the parameters indicated in Table 3.2 and 5.4, the root locus-based analysis is conducted in order to obtain a range of various control parameters. With the assumption of having homogeneous control parameters, i.e. $n_1 = n_2$, $m_1 = m_2$, $m_{\text{int},1} = m_{\text{int},2}$ and $\tau_1 = \tau_2$, the complete eigenvalue spectrum of the system state matrix $\mathbf{A}_{\text{on}}^s$ are shown in Fig. 3.6. It is observed here that with the increase of these control parameters individually, some of the low-frequency eigenvalues gradually move away from the real axis, meanwhile being close to the imaginary axis until the eigenvalues reach unstable right-half zone. On the contrary, some of the stable eigenvalues gradually move on the real axis only away from the imaginary axis. It is evident that the low-frequency eigenvalues near the imaginary axis are the most dominant and crucial for system stability, while the eigenvalues close to the real axis will affect the dynamic performance and damping properties of the system [120]. While the analysis is provided for the system in Fig. 3.1, the methodology can be extended to the general case of n GFM inverters, but not included here due to space constraints.

3.6.2 Analysis of System Operating in off-grid Mode

While the system of Fig. 3.5 is operating in off-grid mode, i.e. $\mathcal{I}_{\text{gs}} = 0$, it can be modelled by a non-linear system, $\dot{\underline{x}}_{\text{off}} = \underline{\mathcal{G}}_{\text{off}}(\underline{x}_{\text{off}})$, where, $\underline{\mathcal{G}}_{\text{off}}(\cdot)$ consists of (3.5), (3.6), (3.7), (3.8), (3.9) for $i=1, 2$. Here,

$$\underline{x}_{\text{off}}^\top = [\theta_1 \ \theta_2 \ \omega_{r,1} \ \omega_{r,2} \ V_{r,1} \ V_{r,2} \ i_{o,1}^d \ i_{o,2}^d \ i_{o,1}^q \ i_{o,2}^q] \quad (3.18)$$

System Model in off-grid condition with $\mathcal{I}_{\text{gs}} = 0$ is shown in Appendix B.2. As a result, the complete system in off-grid mode is a 10th order nonlinear autonomous system with state vector $\underline{x}_{\text{off}}$. The system in off-grid mode is similar to islanded microgrid with two *conventional*-droop controlled inverters in order to share the common load and the followings can be concluded:

- In steady-state, (3.5) and (3.6) for $i = 1, 2$ yield $\omega_{r,1} = \omega_{r,2} = \omega_{\text{ref}}$ which implies $n_1 P_1 = n_2 P_2$ iff $n_1 P_{\text{ref},1} = n_2 P_{\text{ref},2}$.
- It is known that the reactive power sharing accuracy with Q - V droop control is strongly affected by line parameters [120]. Therefore, if line parameters are

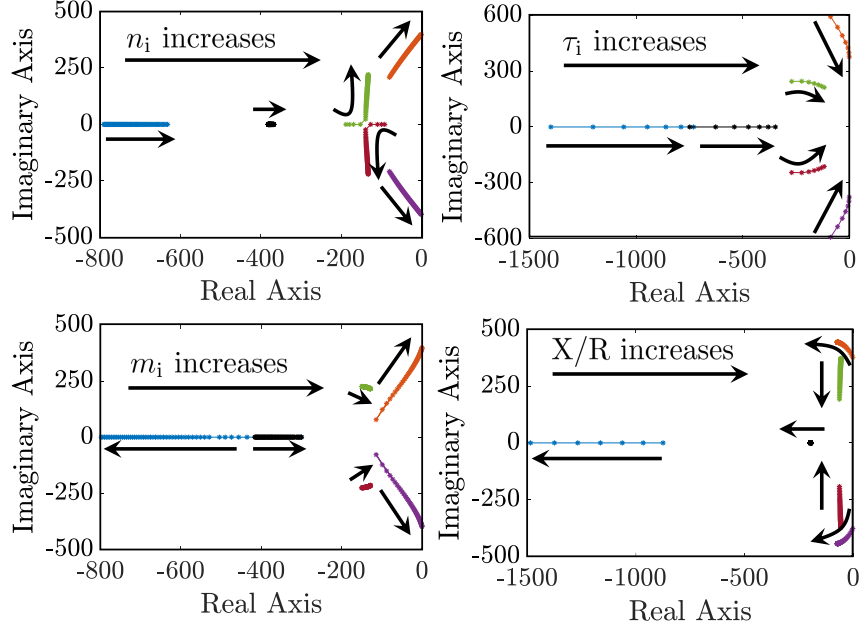


Figure 3.7: Eigenvalue spectrum of $\mathbf{A}_{\text{off}}^s$ for range of parameters in Table 3.2.

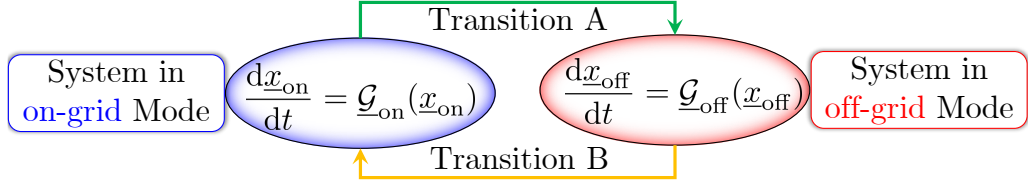


Figure 3.8: Switching of the system during on- and off-grid mode transitions.

neglected then $V_{r,1} = V_{r,2}$ that implies $m_1 Q_1 = m_2 Q_2$ iff $m_1 Q_{\text{ref},1} = m_2 Q_{\text{ref},2}$.

Similarly, considering the system linearized around an equilibrium point, $\underline{x}_{\text{off}}^{\text{eq}}$, such that $\Delta \dot{\underline{x}}_{\text{off}} = \mathbf{A}_{\text{off}} \Delta \underline{x}_{\text{off}}$, where $\mathbf{A}_{\text{off}} = \mathcal{F}_{\text{off}}(\underline{x}_{\text{off}}^{\text{eq}})$ and $\mathcal{F}_{\text{off}}(\cdot)$ is the vector field of $\mathcal{G}_{\text{off}}(\underline{x}_{\text{off}})$, a systematic framework is developed for tuning the control parameters i.e. $n_1, n_2, m_1, m_2, \tau_1, \tau_2$, so that asymptotic stability is guaranteed. Similarly by suppressing the EM dynamics of the lines and loads with the assumption of reaching the fast states i.e. $\underline{x}_{\text{off}}^f := [i_{o,1}^d \ i_{o,2}^d \ i_{o,1}^q \ i_{o,2}^q]^\top$ to a quasi-steady-state (i.e. $\underline{x}_{\text{off,ss}}^f$), the reduced-order model of the system with slow states, $\underline{x}_{\text{off}}^s := [\theta_1 \ \theta_2 \ \omega_{r,1} \ \omega_{r,2} \ V_{r,1} \ V_{r,2}]^\top$ can be formulated as $\Delta \dot{\underline{x}}_{\text{off}}^s = \mathbf{A}_{\text{off}}^s \Delta \underline{x}_{\text{off}}^s$. Similar root locus-based analysis is conducted in order to obtain a range of various control parameters as shown in Fig. 3.7. The sensitivity of droop

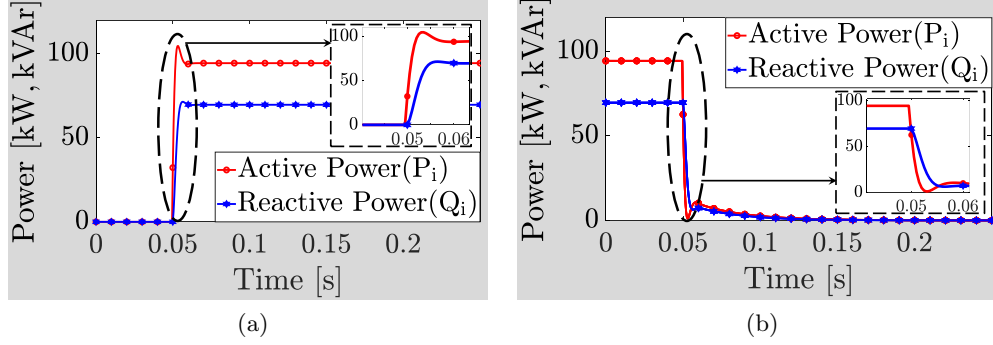


Figure 3.9: Dynamic response of the output active power and reactive power during, (a) Transition A, (b) Transition B of Fig. 3.8 of i^{th} inverter system.

control in off-grid mode w.r.t. X/R ratio of lines is also added in this stability study. It is observed here also that with the increase of these control parameters individually, some of the low-frequency eigen values gradually move away from the real axis, meanwhile close to the imaginary axis until it reach to unstable right-half zone. Similarly, some of the stable eigenvalues gradually move on the real axis only away from the imaginary axis. Moreover, the same range of common control parameters (i.e. n_i 's, m_i 's, τ_i 's are obtained.

3.6.3 Analysis of System During on-/off-grid Mode Transition

Island detection and grid re-synchronization algorithms are necessary for on-grid to off-grid and off-grid to on-grid transition respectively. While the system of Fig. 3.5 experiences transition from on- to off-grid mode and vice versa, the transient behaviors are important to analyze. The transitions are triggered exogenously (due to some aspect of grid failure determined via $\mathcal{I}_{gs}$) which ensures that the stability and steady state analysis of off-grid and on-grid hold for real scenarios. Fig. 3.8 shows the switching of the non-linear system during on-grid to off-grid mode jump (“Transition A”) and off-grid to on-grid mode jump (“Transition B”). During “Transition A”, the system of Section 3.6.2 will have a trajectory starting from initial conditions of the states, $\underline{x}_{\text{off}}^{\text{initial}} \subset \underline{x}_{\text{on}}^{\text{eq}}$, (where $\subset$ denotes the common states between $\underline{x}_{\text{off}}$ and $\underline{x}_{\text{on}}$), to the steady-state values, $\underline{x}_{\text{off}}^{\text{eq}}$. Similarly, during “Transition B”, the system of Section 3.6.1 will have a trajectory starting from initial conditions of the states, $\underline{x}_{\text{on}}^{\text{initial}}$ with $\underline{x}_{\text{off}}^{\text{eq}}$ and added initial states of θ_g , i_g^d , i_g^q , ψ_1^Q , ψ_2^Q , to the steady-state values, $\underline{x}_{\text{off}}^{\text{eq}}$. As evidenced, the

Table 3.3: Inverter and Network Parameters Under Study

Inverter	Value
Ratings (3- ϕ)	480 V (L-L), 60 Hz, 120 kVA, 0.85 pf
Parameters	$V_{dc} = 850$ V, $f_{sw} = 10$ kHz
Filter	$L_f = 150$ μ H, $L_g = 15$ μ H, $C_f = 110$ μ F
Line	$R_{Line,i} = 0.55$ m Ω , $L_{Line,i} = 0.2$ mH, $i = 1, 2$
Grid	480 V (L-L), 60 Hz, $R_{l,g} = 5$ m Ω , $L_{l,g} = 30$ μ H

systems are highly non-linear in nature and as a result $\underline{x}_{on}^{eq}$ and $\underline{x}_{off}^{eq}$ are determined numerically by solving $\underline{\mathcal{G}}_{on}(\underline{x}_{on}) = 0$ and $\underline{\mathcal{G}}_{off}(\underline{x}_{off}) = 0$ using Newton–Raphson method respectively. Forward Euler method is used for simulating the respective transitions with selecting the control parameters from Table 3.2. The response of i^{th} inverter output active power, P_i , and reactive power, Q_i are shown in Fig. 3.9(a) and Fig. 3.9(b) respectively. It is observed that with selected control parameters, both the transitions are smooth enough after minor transients in output power dynamics. Moreover, the response time of reaching new steady-states from initial conditions during the transitions are within few power cycles that are sufficiently fast.

3.7 Controller Hardware-in-the-loop Based Validation

3.7.1 Controller Hardware-in-the-loop Configuration

The CHIL-based real-time (RT) simulation studies are conducted for rapid prototyping of the proposed strategy of seamless transition on a test-system under study that is developed based on electrical network of a commercial scale medical center [121]. The entire system of Fig. 3.1 with R-L network, critical loads (180kW and 80kVAr), and the power circuits of two GFM inverters (parameters tabulated in Table 5.4) are emulated using eMEGASIM platform inside the OP5700 RT-simulator (RTS) manufactured by OPAL-RT. The control loops of both the inverters are realized on two Texas-Instruments TMS320F28379D, 16/12-bit floating-point 200 MHz Delfino micro-controller boards. The control parameters used for validation are tabulated in Table 3.2.

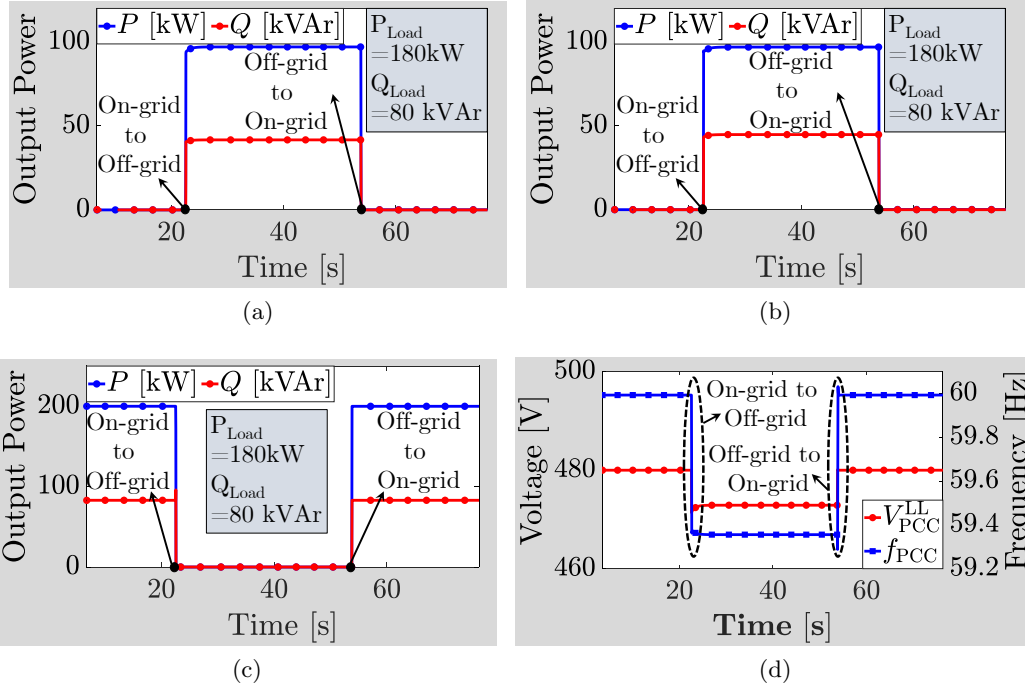


Figure 3.10: Controller Hardware-in-the-loop based results of active and reactive power share of (a) GFM inverter-1, (b) GFM inverter-2, (c) the grid, and (d) RMS voltage and frequency of the system during Case-1.

3.7.2 Validation and Results

To validate the proposed control architecture, two test cases are examined in the RT platform. In **Case-1**, both the GFM inverters are configured for equal power sharing functionality by selecting equal droop co-efficient. In **Case-2**, inverters are configured for unequal power sharing functionality by selecting 2:3 droop co-efficient. In all cases, the CI is initially operating in on-grid mode with a subsequent grid-disconnection and eventually the grid returns. Fig. 3.10(a), Fig. 3.10(b), Fig. 3.10(c) and Fig. 3.11(a), Fig. 3.11(b), Fig. 3.11(c) illustrate the active and reactive power sharing of the two GFM inverters and the grid for **Case-1** and **Case-2** respectively. It is observed in all cases that even if both the GFM inverters are connected and synchronized to the network, during on-grid mode, the output active and reactive power of both the GFM inverters are zero. Thus, the proposed droop controller in all cases makes sure not to utilize GFM inverters while grid is available and supply the entire critical load by the grid alone. Once the

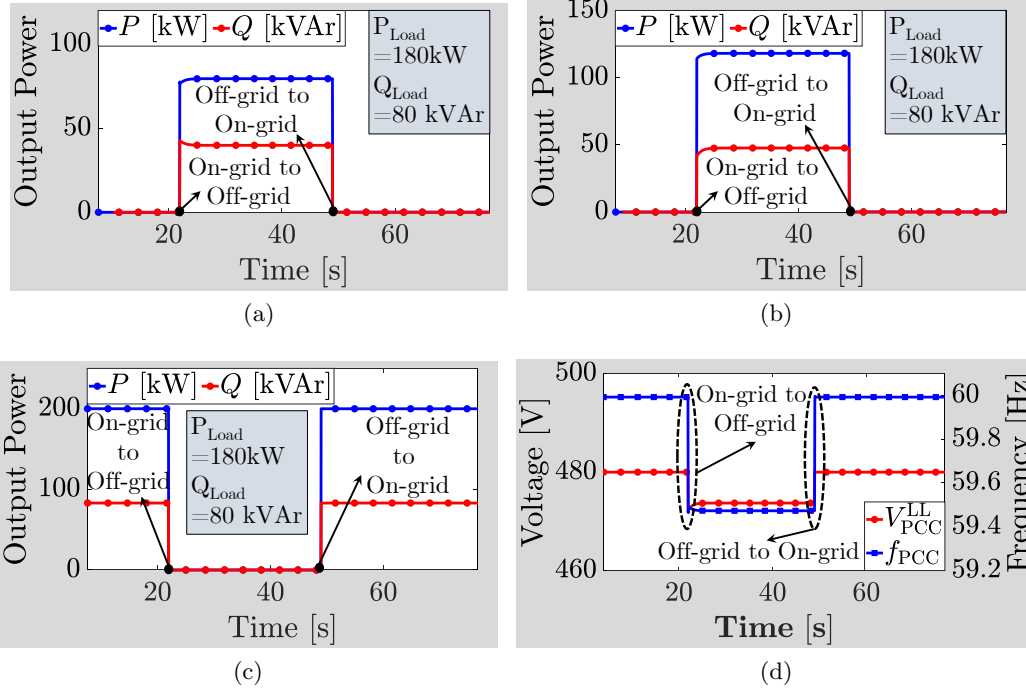


Figure 3.11: Controller Hardware-in-the-loop based results of active and reactive power share of (a) GFM inverter-1, (b) GFM inverter-2, (c) the grid, and (d) RMS voltage and frequency of the system during **Case-2**.

CI is disconnected from grid and operating in off-grid mode at $t \approx 22$ s for **Case-1** and $t \approx 21$ s for **Case-2**, the critical loads are served by both GFM inverters based on droop sharing capability. When grid returns at $t \approx 54$ s for **Case-1** and $t \approx 50$ s for **Case-2**, the entire load is again served by the grid alone and output active and reactive power of the GFM inverters become zero. Fig. 3.10(d) and Fig. 3.11(d) illustrate the voltage RMS and frequency at PCC of the CI for **Case-1**, **Case-2** respectively. It is observed that the voltage RMS and the frequency during the transients and steady-state are within the recommended limits IEEE 446 [115].

During off-grid to on-grid transition, $i_{L,i,ref}^{dq}$ drops down to $[0 \ 0]^T$ from some steady-state value of $i_{L,i,ref}^{dq,ss}$ and stays there as long as it operated in on-grid condition. The $i_{L,i,ref}^{dq,ss}$ is governed by the share of the total load demand of the system in off-grid mode. Similarly, during on-grid mode to off-grid transition, $i_{L,i,ref}^{dq}$ jumps from $[0 \ 0]^T$ to some steady-state value of $i_{L,i,ref}^{dq,ss}$. Figure 3.12 and Figure 3.13 show the response of the inner

current controller (both $i_{L,d}$ and $i_{L,q}$ current loop) of the commercial scale 3- ϕ , 480-V, 125-kVA, four-quadrant GFM inverter manufactured by Dynapower employed with the proposed droop controller during on-grid to off-grid, off-grid to on-grid and during on-grid and off-grid condition.

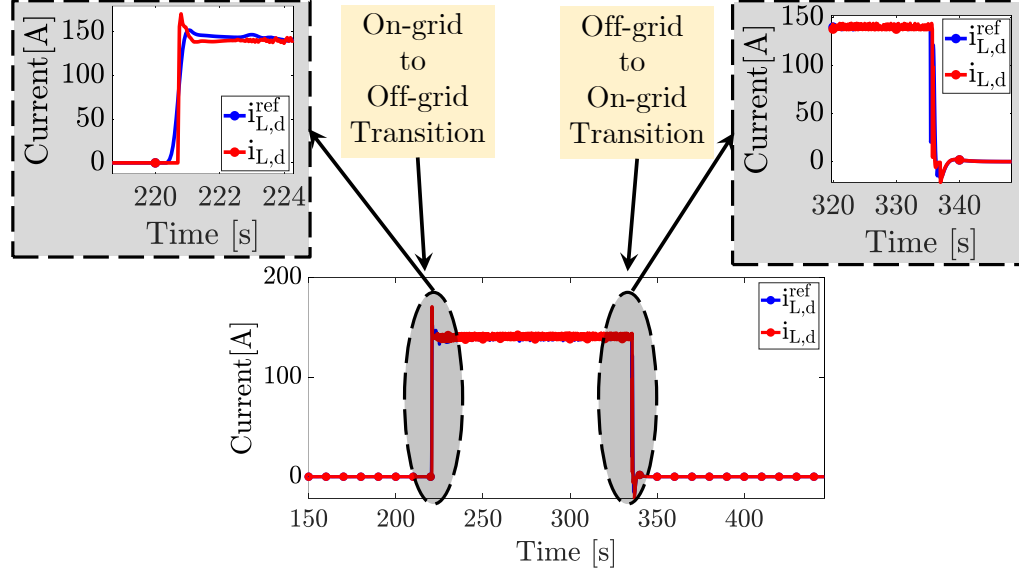


Figure 3.12: The response of $i_{L,d}$ during on-grid to off-grid transition and vice versa.

3.7.3 Performance Comparison with Existing Methods

To evaluate the performance of the proposed seamless transition method, a CHIL-based study is conducted. Here a comparison of the always grid-forming strategy presented in this chapter with the following two methods is presented: Method-1 : the seamless transition method of reference [92] by dual-mode control architecture with pre-determined sinusoidal waveform detection and fast commutation current compensation, and Method-2 : the seamless transition method of reference [107] where a separate smooth transition compensator is added in the outer-voltage control loop of the GFM inverter system. Fig. 3.14(a), Fig. 3.14(b) and Fig. 3.14(c) show the voltage waveform measured at the PCC during on-grid to off-grid transition at $t \approx 22$ s that results employing GFM inverters with proposed seamless control, Method-1, and Method-2 respectively. It is observed that the proposed seamless transition method has significantly less transients in the voltage waveform where the existing transition methods, though

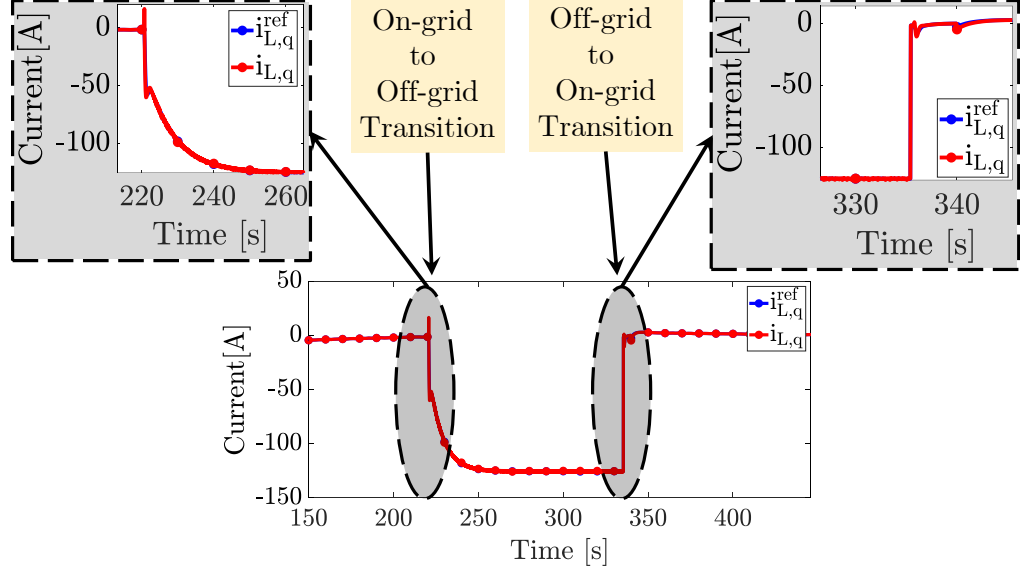


Figure 3.13: The response of $i_{L,q}$ during on-grid to off-grid transition and vice versa.

seamless without any electrical interruptions, has distorted behavior in the waveforms. Similarly, Fig. 3.14(d), Fig. 3.14(e) and Fig. 3.14(f) show the voltage waveform measured at the PCC during off-grid to on-grid transition at $t \approx 42$ s that results while employing with proposed seamless control, Method-1, and Method-2 respectively. It is observed that the proposed transition method with modified droop control has better transient behavior compared to the existing methods. Critical loads in a CI are usually sensitive to the nature of voltage waveform and therefore the proposed seamless transition method is seemed to be a better fit for EPSS applications.

3.8 Hardware-based Experiment and Results

3.8.1 High Power Hardware Configuration

A comprehensive hardware system is developed as shown in Fig. 3.15. The proposed controller with parameters of Table 3.2 is implemented on two commercial scale 3- ϕ , 480-V, 125-kVA, four-quadrant GFM inverter manufactured by Dynapower with Aerovironment AV-900 and NH Research 9300 battery energy storage system, terminated at PCC with a 3- ϕ , 480-V, 270-kVA Ametek RS-series grid simulator, LoadTec OSW4C-based physical load banks with 180 kW, 80 kVAR. The inverters are equipped with

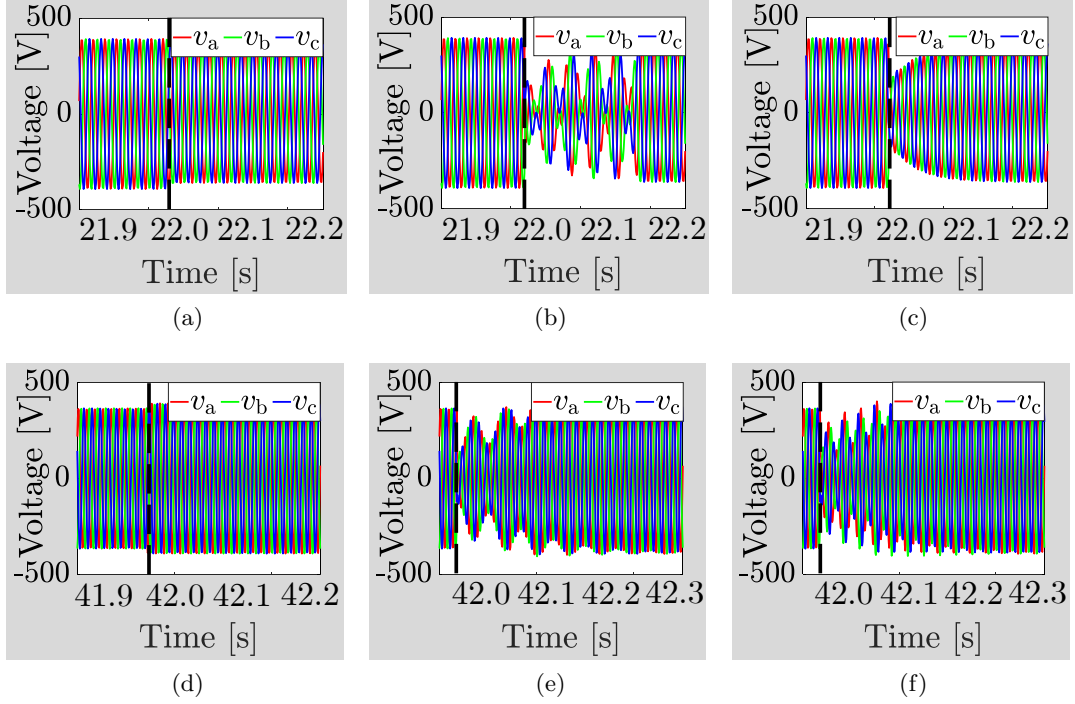


Figure 3.14: Controller Hardware-in-the-loop based performance comparison of (a) proposed method, (b) Method-1 (c) Method-2 during on-grid to off-grid transition and (d) proposed method, (e) Method-1 (f) Method-2 during off-grid to on-grid transition.

AC voltage and frequency protective functionalities for under-voltage (UV) of $< 50\%$ to over-voltage (OV) of $> 120\%$ of the nominal voltage and under-frequency (UF) of $< 57\text{Hz}$ to over-frequency (OF) of $> 62\text{Hz}$ [122]. In compliance with the IEEE Std. 1547, the minimum and maximum ride-through setting times (adjustable by the end user) are 0.16s-300s. Moreover, both instantaneous saturation limit and latched limit-based over-current protection schemes are employed. A supervisory remote island detection is used by means of modbus-based low-bandwidth communication to the inverters. IEEE Std. 1547.4-2011 [71] recommends that to initiate a reconnection of an islanded microgrid and the utility grid, the voltage of the utility grid should be within the “Range B of ANSI/NEMA C84.1-2006, Table 1” and the frequency range should be between 59.3 Hz and 60.5 Hz (for a system of 60 Hz nominal frequency), phase rotation must be correct, and phase angle difference between the islanded microgrid and the utility grid should be within specified acceptable limits. A passive synchronization method with stringent condition (absolute value of voltage, frequency and phase deviations are within ± 0.03

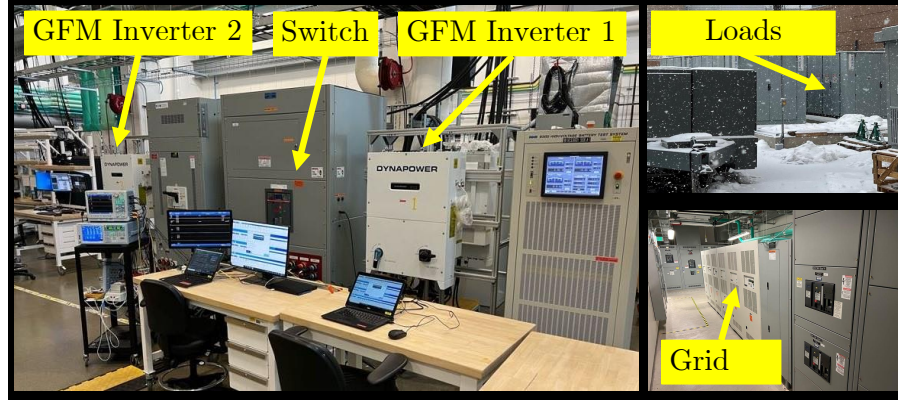


Figure 3.15: Laboratory-based hardware setup for system of Fig. 3.1.

pu, ± 0.1 Hz, and $\pm 1^\circ$ respectively) are employed using a SACE Emax 2 ABB switch, as shown in Figure 3.16.

3.8.2 Validation and Results

The purpose of the hardware validation is to showcase the response of physical GFM inverters with the proposed controller especially during on- to off-grid transition and vice versa as shown in Fig. 3.17 and Fig. 3.18 respectively. It is observed from Fig. 3.17(a) and Fig. 3.17(b) that during on-grid to off-grid transition at $t \approx 0.5$ s, the output active and reactive power of both the GFM inverters jumps from zero to 90kW and 40kVAR respectively. Fig. 3.17(c) and Fig. 3.17(d) show the output current waveforms of the inverters during this transition. Fig. 3.17(e) and Fig. 3.17(f) show the instantaneous voltage waveform, RMS value of the voltage and the frequency of the system respectively. Similarly, Fig. 3.18(a) and Fig. 3.18(b) show that during off-grid to on-grid transition at $t \approx 0.3$ s, the output active and reactive power of both the GFM inverters jumps from 90kW and 40kVAR to zero respectively. Fig. 3.18(c) and Fig. 3.18(d) show the output current waveforms of the inverters during this transition. Fig. 3.18(e) and Fig. 3.18(f) show the instantaneous voltage waveform, RMS value of the voltage and the frequency of the system respectively. The instantaneous waveforms substantiate the fact that both the transitions are smooth and seamless with the proposed controller. Moreover, the voltage RMS and the frequency during the transition and steady-state are within the recommended limits of IEEE 446 [115].

Moreover, a hardware experiment is conducted to evaluate the performance of the

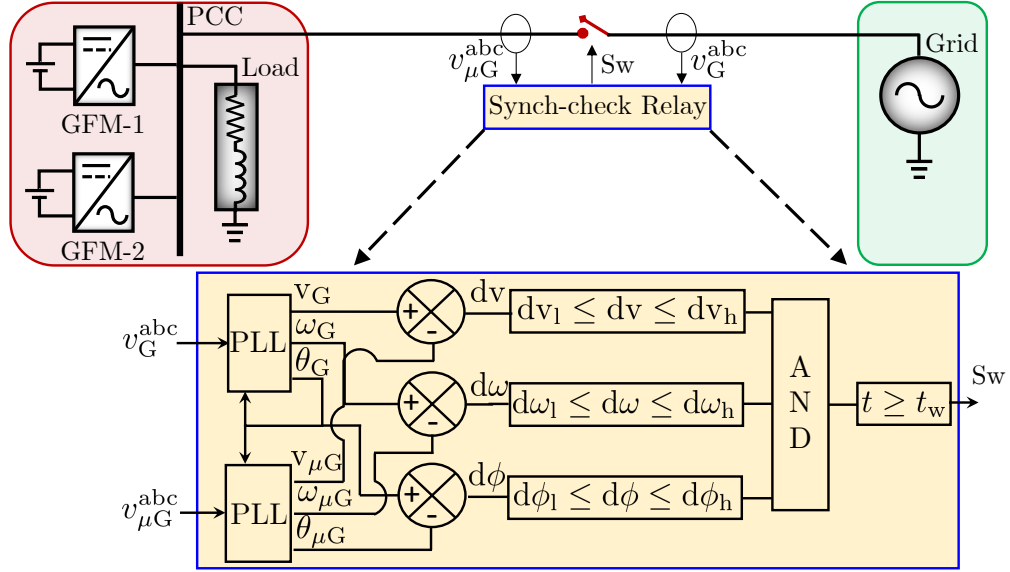


Figure 3.16: Passive synchronization-based reconnection method of islanded microgrid with utility grid in accordance with the IEEE Std. 1547.

proposed seamless transition strategy in presence of non-linear load and communication delay. **Case-3** is defined where the load in the system is modified by adding a three-phase diode-bridge rectifier based non-linear load. The system experiences an on-grid to off-grid transition at $t \approx 5s$ and an off-grid to on-grid transition at $t \approx 26s$. Moreover, a delay of $\approx 2s$ is introduced for the signal $\mathcal{I}_{gs}$ to be transmitted to GFM-2. Fig. 3.19(a) and Fig. 3.19(b) show the active and reactive power share between the GFM inverters and the grid. It is observed that the reactive power of GFM-2 after $t \approx 5s$ is gradually decreasing until Q - V integrator hits the limit caused by the integrator loop of GFM-2 remaining active during off-grid mode due to communication delay. As a result, in order to maintain generation-demand balance, GFM-1 supplies more reactive power during the 2s time span. However at $t \approx 7s$, when the integrators' are inactive in both the GFMs', the system returns with normal droop-controlled GFM inverters and reactive power demand is shared accordingly. It is observed that due to the communication delay, there is minimal impact on the active power sharing and control as shown in Fig. 3.19(a). The similar situation arises during $t \approx 26s$ also where integrator is not active in GFM-2 after off-grid to on-grid transition. As a result GFM-2 supplies reactive power during on-grid mode for some time and as a result the grid is supplying less reactive power until the time instant when the integrator is active again after the communication delay. Fig. 3.19(c)

and Fig. 3.19(d) show the current waveform of the inverters where the total demand distortion of the current waveforms of GFM-1 and GFM-2 are $\approx 2.83\%$ and 2.77% of fundamental current that signifies that the harmonic load is almost equally shared by the GFM inverters. Moreover, total harmonic distortion of the voltage waveform, as shown in Fig. 3.20(a), is 1.16% due to harmonic voltage drop across the lines that is less than limits recommends by [123]. Fig. 3.20(b) shows that the frequency and RMS voltage during the transition and steady-state are within the recommended limits [115].

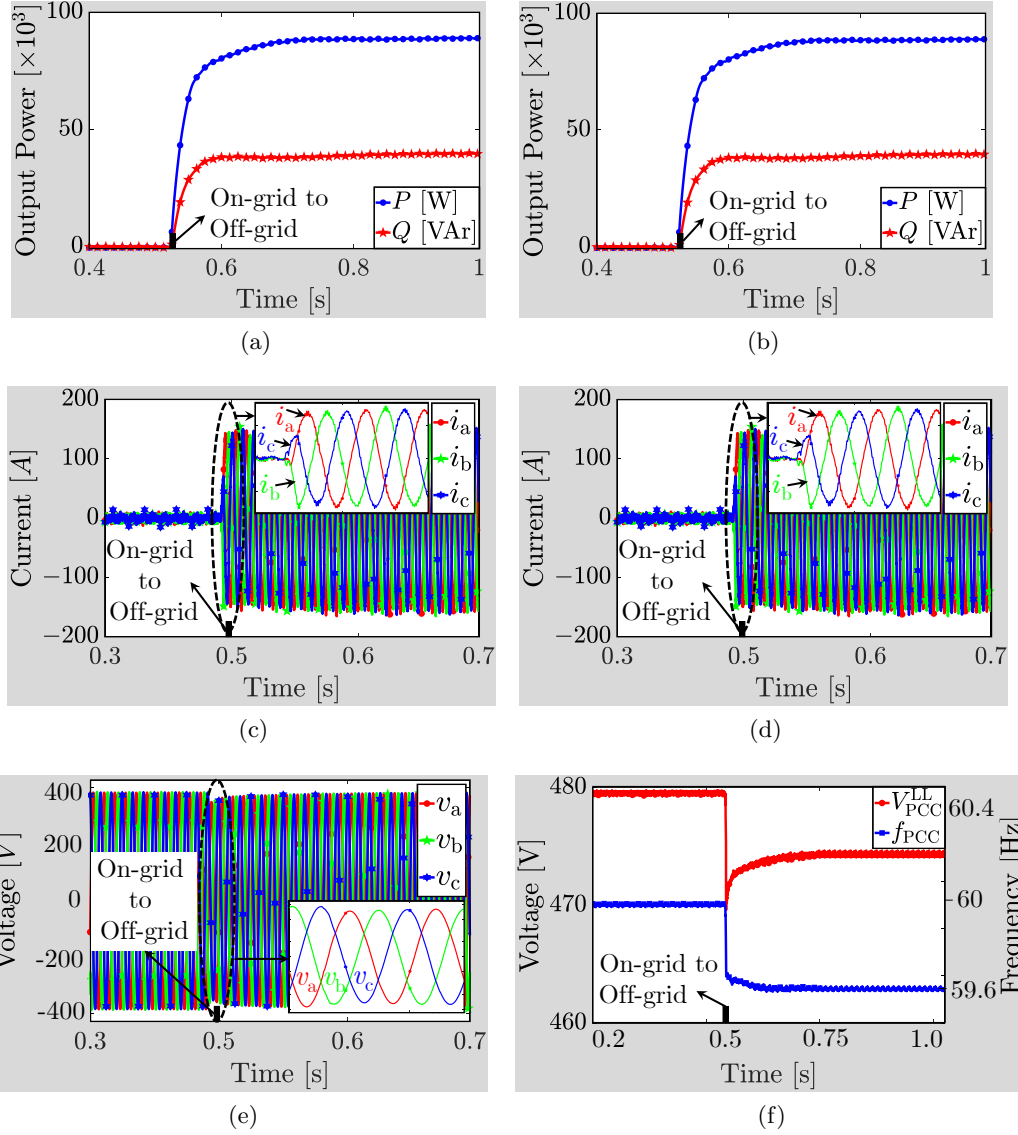


Figure 3.17: Hardware results of output active and reactive power of (a) GFM inverter-1, (b) GFM inverter-2, instantaneous output current waveform of (c) GFM inverter-1, (d) GFM inverter-2, (e) instantaneous voltage waveform, and (f) RMS voltage and frequency of the system measured at PCC during on-grid to off-grid transition of the hardware system.

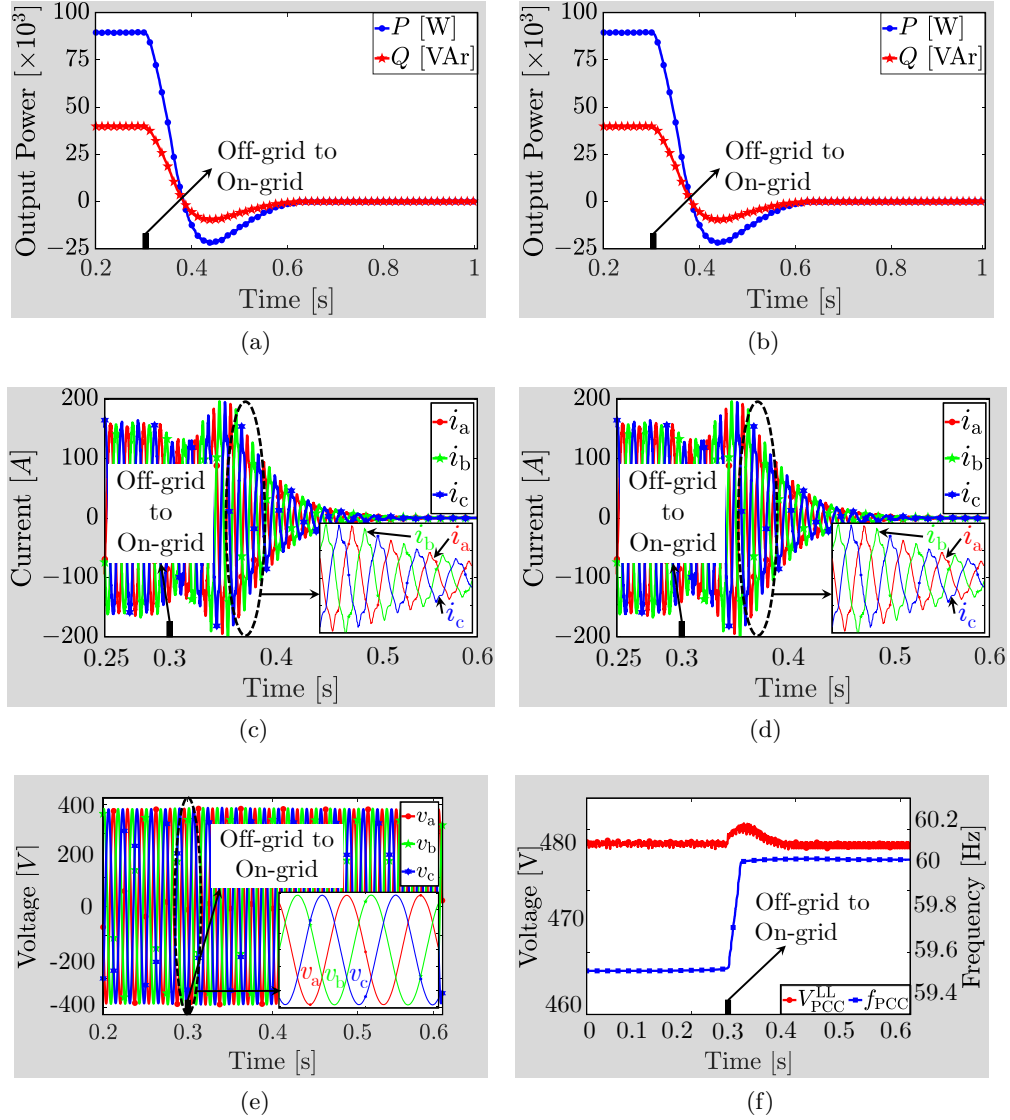


Figure 3.18: Hardware results of output active and reactive power of (a) GFM inverter-1, (b) GFM inverter-2, instantaneous output current waveform of (c) GFM inverter-1, (d) GFM inverter-2, (e) instantaneous voltage waveform, and (f) RMS voltage and frequency of the system measured at PCC during off-grid to on-grid transition of the hardware system.

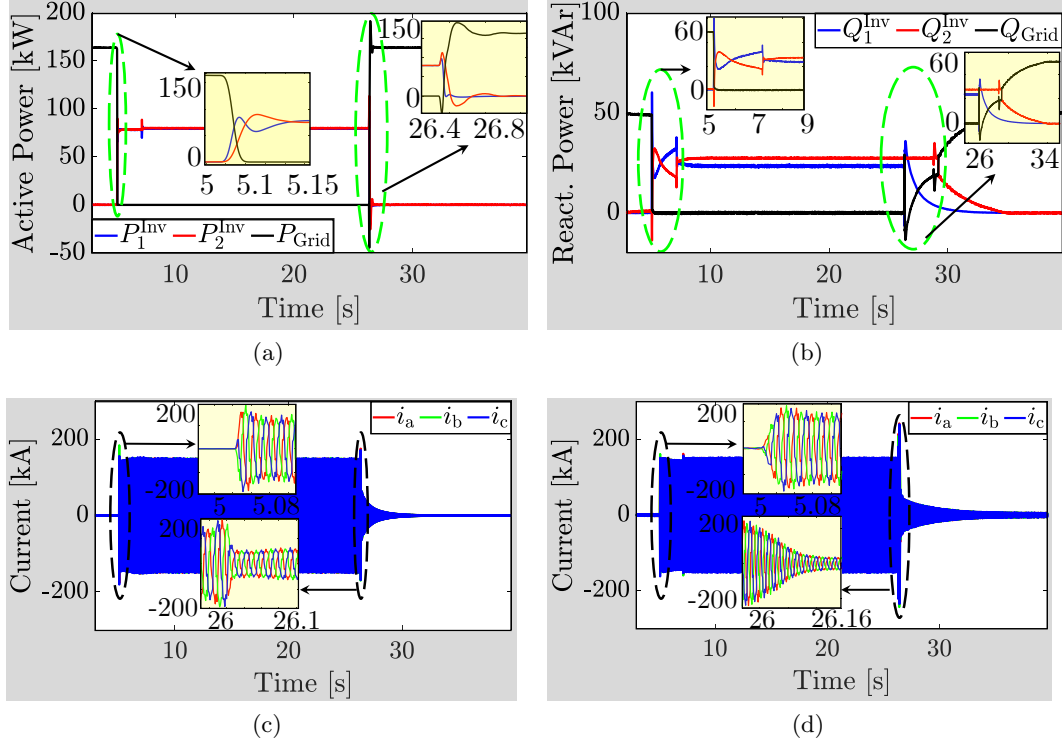


Figure 3.19: Hardware results of (a) active, (b) reactive power share of GFM inverter-1, GFM inverter-2, and the grid, and instantaneous output current waveform of (c) GFM inverter-1, (d) GFM inverter-2, during Case-3.

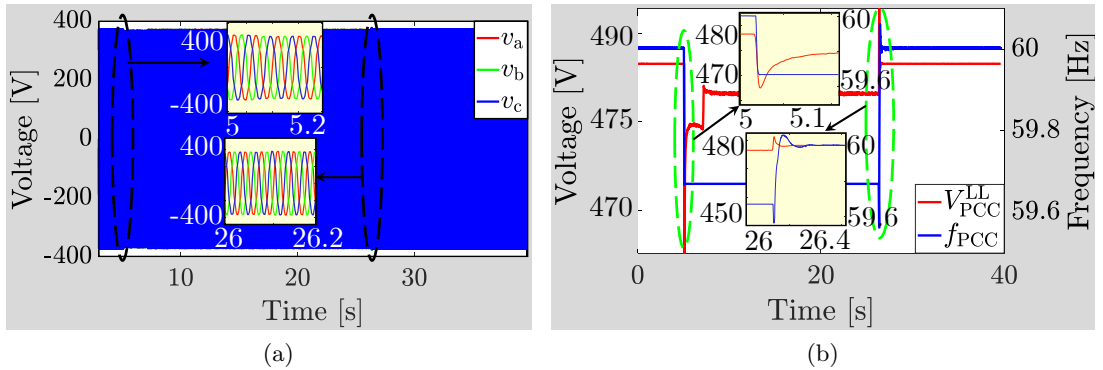


Figure 3.20: Hardware result of (a) instantaneous voltage waveform, (b) RMS voltage and frequency measured at PCC during Case-3.

Chapter 4

Dynamic Microgrid: Novel Active Synchronization Using Grid-forming Inverters

4.1 Gist of the Chapter

For smooth transition from islanded mode to grid-tied mode, synchronization to the incoming grid is required for any islanded microgrid. This chapter describes a novel active synchronization method for such islanded microgrids. In this method, the proposed synchronization controller compensates the phase angle and voltage magnitude difference between islanded microgrid side and the incoming grid side to zero by adjusting the droop laws of all grid-forming (GFM) inverter systems of the microgrid dynamically via secondary control layer of microgrid hierarchical control architecture. It is analytically shown that compensation on only phase angle and voltage magnitude is sufficient for synchronization as frequency is compensated by this proposed method indirectly by the phase angle compensation. A systematic approach for tuning the proposed synchronization controllers are provided. For validation, a controller hardware-in-the-loop-based real-time simulation is conducted using OP5700 RT-simulator manufactured by OPAL-RT on a 3-phase, 480V, 500kVA, 55-bus urban microgrid system with 6 GFM inverters where control of 2 GFM inverters are realized on low-cost Texas-Instruments TMS28379D Delfino controller boards.

Note: More details on this work can be found in the published article [\[124\]](#).

4.2 State-of-the-art on Synchronization Schemes

The emerging concept of microgrid has gained increased popularity in recent decades. It is usually self sustained owing to its ability to integrate distributed generation units such as photo-voltaics, wind generators, diesel generators, battery energy storage systems etc. By virtue of such abilities, it can maintain reliable power supply to the local loads in presence of a grid or in absence of a grid via grid-forming (GFM) inverter systems. Moreover, it is capable of systematically connecting to or disconnecting from the grid based on certain operating conditions. But to facilitate such operating capabilities, the microgrids must have some well defined methods that will allow to connect or disconnect from the larger utility grid.

One such important method is synchronization of an islanded microgrid with the incoming grid before re-connection with it. During such synchronization process, the islanded microgrid is required to fulfill some standard operating criteria. Accredited standard IEEE 1547.4-2011 recommends that for synchronization of an islanded microgrid, its voltage should be within the “Range B of ANSI/NEMA C84.1-2006” and its frequency should be between 59.3-60.5 Hz, and phase rotation must be correct [71]. Further, the voltage, frequency, and phase angle between the islanded microgrid and the grid should be within acceptable limits specified in IEEE 1547-2003 [125].

In literature many synchronization methods have been proposed and briefly mentioned in Table 4.2. Synchronization methods can broadly be classified into three categories [126]. The most basic of which is called “Open Transition Transfer” method that simply de-energizes the whole islanded microgrid before connecting it to the utility grid. Even though the simplicity of such methods reduces the capital cost and design complexity, the interruption of power supply during re-connection process makes it less reliable and less attractive in modern applications. A simple alternative is “Passive” method [127, 128]. As summarized in Table 4.2, this method simply monitors the difference between voltage, frequency and phase angle of the microgrid and the utility grid and waits for a favourable condition to occur. Even though these methods are simple and cost effective, it might come short in terms of reliability and speed. A more advanced alternative is “Active” method [129, 130, 131, 132, 133, 134, 135, 136]. As described in Table 4.2, these methods actively control voltage, frequency, or phase angle of microgrid side by utilizing various control methods to make the microgrid side favourable for the re-connection and thus actively synchronize the microgrid with the grid. But the existing active methods have some challenges and disadvantages. As pointed out in

Table 4.2, some of these methods makes certain assumptions that are rarely satisfied in real-world scenarios; some need costly and bulky equipment and high bandwidth communication facilities; few methods use complicated controllers with challenging designs; many of these methods create unwanted transients in the system.

The proposed method in this chapter alleviates such difficulties by implementing a novel synchronization control method with droop-controlled GFM inverters in the microgrid. It compensates the phase angle and voltage magnitude difference between islanded microgrid side and the incoming grid side to zero by adjusting the droop laws of all GFM inverter systems of the microgrid dynamically via secondary control layer of microgrid hierarchical control architecture and as a result the frequency is also synchronized indirectly. This modification allows for better synchronization conditions and performs relatively fast and smooth synchronization process.

Table 4.1: Major Works on Microgrid Re-synchronization Methods

Ref.	Types	Description of the Method	Potential Drawbacks and Advantages
[129, 130] [131, 132] [133, 134] [135, 136]	Active Method	• Voltage and phase difference between microgrid side and the grid side is regulated to zero by utilizing ramp/virtual inertia/PI based control methods • Determines the effects of a connecting impedance between grid side and the microgrid side and emulates virtual impedance inside inverter control. • Emulates the behaviour of induction machines to extract grid side frequency for synchronization. • Calculates grid side phase angle w.r.t. microgrid side and compensates the angle difference directly. • The error between grid side frequency and the inverter side frequency is calculated and then depending on its value, transitions are made between modes until it stabilizes at the final.	• Underlying assumption that the frequency on both grid side and microgrid are same. • Need high bandwidth remote measurements. • Complicated controllers, challenging tuning. • Abrupt change in phase and voltage during synchronization causes large transients. • Methods require bulky circuit elements. • Difficulties in rapid transition between modes because of slow response in synchronization. • Performance of the synchronizations depend on initial phase, frequency, voltage difference. • The most notable advantage of these methods are that these are more reliable and faster compared to other synchronization methods.
		• Checks favourable conditions of synchronization and facilitates the interconnection mechanism only when all the conditions are met and the connecting switches are activated automatically.	• Lacks reliability and slow, uncontrolled in nature. • Large transients during synchronization. • These methods are less costly due to the lack of active synchronization control method.
		• In this method the microgrid is de-energized completely before connecting it to the grid. • The passive network are connected to grid first and local generating resources are brought online.	• Intermittent power supply in the microgrid during the re-connection process makes it unreliable and not acceptable in most modern applications. • The method incurs very low capital cost.
		✓ Synchronization controller compensates the phase angle and voltage difference between two sides to zero by adjusting the droop laws of GFM inverter systems of the microgrid dynamically and quickly.	✓ The proposed method is relatively fast and the synchronization process enables relatively smooth re-connection with the incoming grid with less transients while maintaining the capital cost at the same level of existing active methods.
		✓ Uses secondary microgrid control architecture layer.	
[127, 128]	Passive Method		
[126, 71]	Open Transition Transfer Method		
	Active Method		
Proposed method			

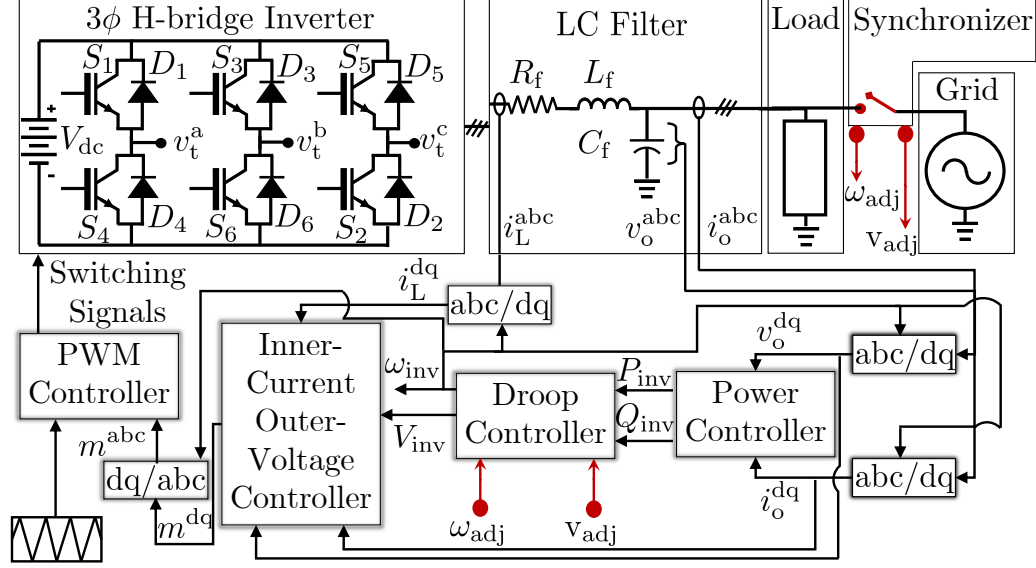


Figure 4.1: An islanded microgrid with a single grid-forming inverter supplying load while grid is disconnected by the microgrid switch.

4.3 Proposed Synchronization Mechanism

The proposed synchronization mechanism will be discussed in detail on a single GFM inverter-based microgrid system first and then adopted for a generic microgrid system.

4.3.1 Synchronization for Single GFM Inverter-based Microgrid

A microgrid system while operating in islanded mode with single GFM inverter supplying load and the grid is disconnected by a switch is shown in Fig. 4.1. The power circuit of the 3- ϕ H-bridge GFM inverter is connected across the load via an LC filter with L_f , C_f and R_f as filter inductance, capacitance and parasitic resistance respectively. A dq-frame multi-loop controller is employed that generates modulated voltage vector signal, m^{abc} , to pulse-width modulation (PWM) controller to generate switching signals resulting in terminal voltages, v_t^a , v_t^b and v_t^c . This multi-loop control consists of conventional inner-current-outer-voltage controller as discussed in [116]. By referring Fig. 4.1, the power controller, the modified droop controller for the purpose of proposed synchronization method and the synchronizing controller will be discussed next.

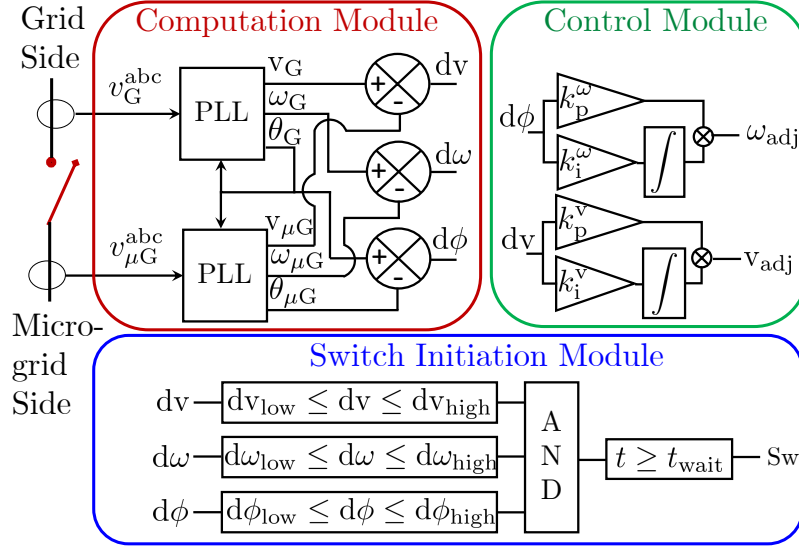


Figure 4.2: Control architecture of the proposed active synchronization controller.

Power Controller

The dq component of output voltage, v_o^{dq} , and current, i_o^{dq} , are used to determine the instantaneous active power, p_i , and reactive power, q_i , of the GFM inverter and processed through low-pass filters with the time constant, $\tau_S \in \mathbb{R}_{>0}$, to obtain the average active power, $P_i = [1/(\tau_S s + 1)]p_i$, and reactive power, $Q_i = [1/(\tau_S s + 1)]q_i$, where $p_i := 1.5 \times [v_{o,i}^d i_{o,i}^d + v_{o,i}^q i_{o,i}^q]$ and $q_i := 1.5 \times [v_{o,i}^q i_{o,i}^d - v_{o,i}^d i_{o,i}^q]$.

Modified Droop Controller

The proposed modified active power/frequency, P - ω , and reactive power/voltage, Q - V , droop control law is described as follows:

$$\omega_{inv} = \omega_N - n_P P_i + \omega_{adj}, \quad V_{inv} = V_N - m_Q Q_i + v_{adj}, \quad (4.1)$$

where, ω_N , V_N are the nominal frequency (in rad/s) and voltage (in volt) of the system respectively. n_P and m_Q are the P - ω and Q - V droop coefficients respectively. The values of n_P and m_Q are typically chosen such that ω_i and V_i are within the allowed specification, defined by IEEE 1547 Standard [71], while GFM is at rated loading condition. ω_{adj} (in rad/s) and v_{adj} (in volt) are the additional adjustments in droop laws taken care by the proposed synchronizing controller.

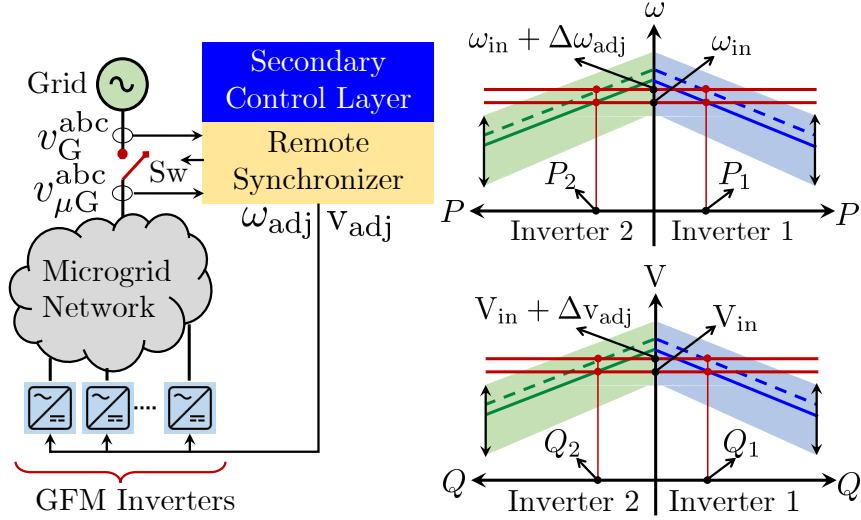


Figure 4.3: Proposed synchronization method adopted for generic microgrid.

Synchronization Controller

Unlike the power and droop controller that are employed inside GFM inverter system, synchronization controller is situated inside the microgrid switch. The synchronizer of Fig. 4.1 is shown in more details in Fig. 4.2. It measures the instantaneous voltage waveforms of both sides of the microgrid switch, i.e. grid side and microgrid side. It consists of three modules as follows:

- **Computational Module:** In this module, two conventional synchronous reference frame-based phase-locked loops (SRF-PLL), that can be found in [137], are used to compute the respective magnitude (v_G , $v_{\mu G}$), frequency (ω_G , $\omega_{\mu G}$) and phase angles (θ_G , $\theta_{\mu G}$) of the main grid and the microgrid respectively w.r.t. the grid reference considered as common reference frame. The outputs of this module are $dv := v_G - v_{\mu G}$, $d\omega := \omega_G - \omega_{\mu G}$ and $d\phi := \theta_G - \theta_{\mu G}$.
- **Control Module:** This module consists of two proportional and integral (PI) controllers. The purpose of the PI controller with parameters, k_p^ω and k_i^ω , is to regulate the input signal, $d\phi$, to zero. Whereas, the purpose of the PI controller with parameters, k_p^v and k_i^v , is to regulate the input signal, dv , to zero. It is important to mention that it is sufficient to regulate $d\phi = (\omega_G - \omega_{\mu G}) \times t = d\omega \times t$ to zero that results in regulating $d\omega$ to zero dynamically. The regulators are generating the required adjustment signals, ω_{adj} and v_{adj} as output and these are sent to the GFM inverter droop controller to adjust

the droop laws as given in (4.1). It is to be noted that both the adjustment signals, ω_{adj} and v_{adj} , can be transmitted to the local GFM inverter controller via conventional secondary control architecture of a microgrid [138].

- **Switch Initiation Module:** This module is designed in order to find the suitable time instant for re-connection of the islanded microgrid with the main grid by the microgrid switch. This module continuously tracks the dv , $d\omega$ and $d\phi$ across the switch and checks whether these signals are within the pre-defined bound as shown in Fig. 4.2. If all three signals are within their respective bounds continuously for t_{wait} , then it generates a signal, Sw , to make the microgrid switch on. The bound check and the waiting conditions are making sure that the voltage waveform of the islanded microgrid is sufficiently aligned with the voltage waveform of the grid and hence the microgrid can be connected with the main grid.

Remark 2. *Once the microgrid switch is closed, the same signal, Sw , is used to reset the integrator in the Control Module. This makes the values of the adjustments zero once the microgrid is connected to main grid. Hence, normal droop equations returns on on-grid conditions.*

4.3.2 Synchronization for a Generic Microgrid

For generic microgrid system as shown in left of Fig. 4.3 with multiple droop controlled GFM inverters operating in parallel during islanded mode, the proposed synchronization method, employed near the grid switch, can be adopted as part of secondary control layer of microgrid control architecture [138]. In other words, the remote synchronization switch and the control method generates the adjusted signals, ω_{adj} and v_{adj} , and transmit these to all GFM inverters as conventional secondary frequency and voltage adjustments without employing additional communication medium. The very basic idea of running each GFM inverters distributively but not in master-slave way during synchronization, is the cause of transmitting the adjusted signals to all GFM inverters to modify droop laws equally. For an example, in an islanded microgrid with two droop-controlled GFM inverters, the steady-state frequency and voltage, before the initiation of synchronization, was ω_{in} and V_{in} and the total load was shared such that $n_{P,1}P_1 = n_{P,2}P_2$ and $m_{Q,1}Q_1 = m_{Q,2}Q_2$ as shown in right top and bottom of Fig. 4.3. After the synchronization starts and adjustment signals are added to each GFM inverter equally, the new steady-state frequency becomes $\omega_{\text{in}} + \Delta\omega_{\text{adj}}$ and $V_{\text{in}} + \Delta v_{\text{adj}}$.

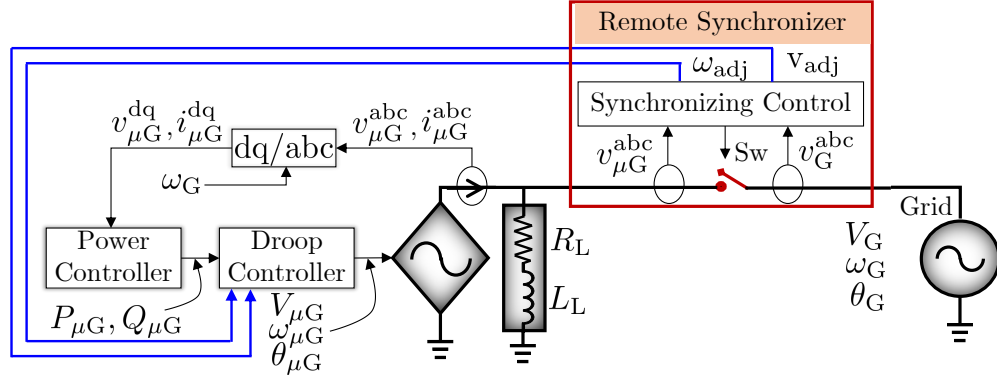


Figure 4.4: Equivalent representation of the microgrid system of Fig. 4.1.

The following can be stated in steady-state:

$$\begin{aligned}\omega_{in} + \Delta\omega_{adj} &= \omega_N - n_{P,1}P_1 + \omega_{adj} = \omega_N - n_{P,2}P_2 + \omega_{adj}, \\ V_{in} + \Delta v_{adj} &= V_N - m_{Q,1}Q_1 + v_{adj} = V_N - m_{Q,2}Q_2 + v_{adj}.\end{aligned}$$

Therefore, the total load sharing will be intact such that $n_{P,1}P_1 = n_{P,2}P_2$ and $m_{Q,1}Q_1 = m_{Q,2}Q_2$ as shown in right top and bottom of Fig. 4.3. Therefore, this way of compensation impacts less on the normal operating conditions of the islanded microgrid during synchronization process.

4.4 Synchronization Control-loop Analysis

The goal of this section is to analyse the working principle and to guide the selection of the parameters of the proposed synchronization control by referring the system in Fig. 4.1. For the sake of ease in analysis, the followings are considered:

- The outer-voltage-inner-current control loop is stable and has faster (> 10 times) dynamics when compared to the power and droop controller and tracks its voltage reference with minimal (assumed zero) tracking error as suggested in [116].
- Load is assumed to be a Z-type with R_L and L_L in series.
- abc-dq conversions are adopted w.r.t. the main grid reference with fixed phase angle, θ_G , frequency, ω_G and voltage, V_G .

These allow to model a GFM inverter as a voltage source with controllable phase, frequency and amplitude. As a result, circuit in Fig. 4.1 can be represented by Fig. 4.4. To model the system with the remote synchronizer, the following non-linear state-space equations can be obtained for modeling:

$$\dot{\theta}_{\text{ig}} = \omega_{\mu\text{G}} - \omega_{\text{G}}, \text{ where } \theta_{\text{ig}} := \theta_{\mu\text{G}} - \theta_{\text{G}}, \quad (4.2)$$

$$\tau_{\text{S}}\dot{\omega}_{\mu\text{G}} = \omega_{\text{N}} - \omega_{\mu\text{G}} - n_{\text{P}}p_{\mu\text{G}} + \omega_{\text{adj}} + \tau_{\text{S}}\dot{\omega}_{\text{adj}}, \quad (4.3)$$

$$\tau_{\text{S}}\dot{V}_{\mu\text{G}} = V_{\text{N}} - V_{\mu\text{G}} - m_{\text{Q}}q_{\mu\text{G}} + v_{\text{adj}} + \tau_{\text{S}}\dot{v}_{\text{adj}}, \quad (4.4)$$

$$L_{\text{L}}\dot{i}_{\mu\text{G}}^{\text{d}} = V_{\mu\text{G}} \cos \theta_{\text{ig}} - R_{\text{L}}i_{\mu\text{G}}^{\text{d}} + \omega_{\text{G}}L_{\text{L}}i_{\mu\text{G}}^{\text{q}}, \quad (4.5)$$

$$L_{\text{L}}\dot{i}_{\mu\text{G}}^{\text{q}} = V_{\mu\text{G}} \sin \theta_{\text{ig}} - R_{\text{L}}i_{\mu\text{G}}^{\text{q}} - \omega_{\text{G}}L_{\text{L}}i_{\mu\text{G}}^{\text{d}}, \quad (4.6)$$

where, $p_{\mu\text{G}} := 1.5 \times [V_{\mu\text{G}}i_{\mu\text{G}}^{\text{d}} \cos \theta_{\text{ig}} + V_{\mu\text{G}}i_{\mu\text{G}}^{\text{q}} \sin \theta_{\text{ig}}]$ and $q_{\mu\text{G}} := 1.5 \times [-V_{\mu\text{G}}i_{\mu\text{G}}^{\text{q}} \cos \theta_{\text{ig}} + V_{\mu\text{G}}i_{\mu\text{G}}^{\text{d}} \sin \theta_{\text{ig}}]$. Therefore, the islanded microgrid system can be modeled as:

$$\dot{\underline{x}}_{\mu\text{G}} = \mathcal{F}(\underline{x}_{\mu\text{G}}, \underline{u}_{\mu\text{G}}), \quad \underline{y}_{\mu\text{G}} = \mathcal{H}(\underline{x}_{\mu\text{G}}, \underline{u}_{\mu\text{G}}), \quad (4.7)$$

$$\text{where, } \underline{x}_{\mu\text{G}} = \begin{bmatrix} \theta_{\text{ig}} & \omega_{\mu\text{G}} & V_{\mu\text{G}} & i_{\mu\text{G}}^{\text{d}} & i_{\mu\text{G}}^{\text{q}} \end{bmatrix}^{\top},$$

$$\underline{u}_{\mu\text{G}} = [\omega_{\text{adj}} \quad \dot{\omega}_{\text{adj}} \quad v_{\text{adj}} \quad \dot{v}_{\text{adj}}]^{\top}, \quad \underline{y}_{\mu\text{G}} = [\theta_{\text{ig}} \quad V_{\mu\text{G}}]^{\top}.$$

PI-based synchronizing controller equation can be written as:

$$\omega_{\text{adj}} = k_{\text{p}}^{\omega}\theta_{\text{ig}} + k_{\text{i}}^{\omega} \int \theta_{\text{ig}} dt; \quad \dot{\omega}_{\text{adj}} = k_{\text{p}}^{\omega}\dot{\theta}_{\text{ig}} + k_{\text{i}}^{\omega}\theta_{\text{ig}}, \quad (4.8)$$

$$v_{\text{adj}} = k_{\text{p}}^{\text{v}}V_{\text{ig}} + k_{\text{i}}^{\text{v}} \int V_{\text{ig}} dt; \quad \dot{v}_{\text{adj}} = k_{\text{p}}^{\text{v}}\dot{V}_{\mu\text{G}} + k_{\text{i}}^{\text{v}}V_{\mu\text{G}}; \quad (4.9)$$

where, $V_{\text{ig}} = V_{\mu\text{G}} - V_{\text{G}}$. Based on the system and the controller models given in (4.2)-(4.9), following remark can be made:

Remark 3. *Considering a stable synchronization operation with properly tuned controller parameters, in steady-state of the system (4.2), (4.3), (4.4), (4.8) and (4.9) result in $\theta_{\text{ig}}^{\text{ss}} = 0$, $\omega_{\mu\text{G}}^{\text{ss}} = \omega_{\text{G}}$ and $V_{\mu\text{G}}^{\text{ss}} = V_{\text{G}}$. In other words, voltage of the islanded microgrid, $v_{\mu\text{G}}^{\text{abc}}$, is synchronized with the grid voltage, $v_{\text{G}}^{\text{abc}}$.*

Now, in order to get a guidance for selecting a properly tuned parameters of the synchronization controller (i.e. k_{p}^{ω} , k_{i}^{ω} , k_{p}^{v} and k_{i}^{v}), the system in (4.7) is linearized

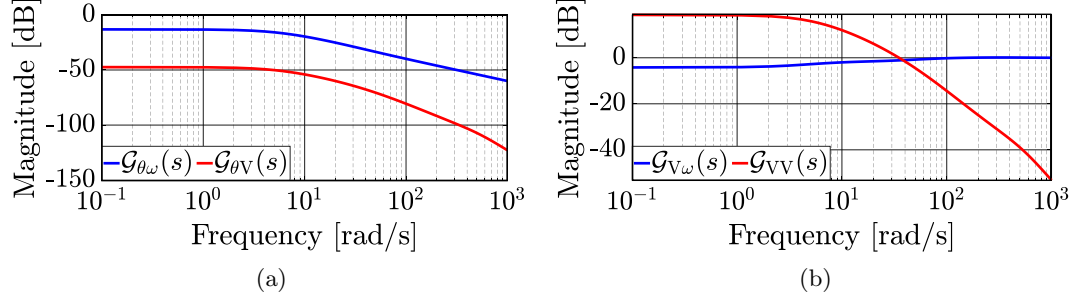


Figure 4.5: Bode magnitude plot of, (a) $\mathcal{G}_{\theta\omega}(s)$, $\mathcal{G}_{\theta V}(s)$, (b) $\mathcal{G}_{V\omega}(s)$, $\mathcal{G}_{VV}(s)$.

around the steady-state equilibrium point and can be written as follows:

$$\Delta \dot{x}_{\mu G} = \underbrace{\frac{\partial \mathcal{F}}{\partial x_{\mu G}} \bigg|_{\text{eq}}}_{\mathbf{A}_{\mu G}} \Delta x_{\mu G} + \underbrace{\frac{\partial \mathcal{F}}{\partial u_{\mu G}} \bigg|_{\text{eq}}}_{\mathbf{B}_{\mu G}} \Delta u_{\mu G}, \quad (4.10)$$

$$\Delta y_{\mu G} = \underbrace{\frac{\partial \mathcal{H}}{\partial x_{\mu G}} \bigg|_{\text{eq}}}_{\mathbf{C}_{\mu G}} \Delta x_{\mu G} + \underbrace{\frac{\partial \mathcal{H}}{\partial u_{\mu G}} \bigg|_{\text{eq}}}_{\mathbf{D}_{\mu G}} \Delta u_{\mu G}. \quad (4.11)$$

Taking Laplace transformation of (4.10), (4.11) and having some algebraic manipulation, the following multi-input-multi-output (MIMO) transfer function model can be derived for the system:

$$\begin{bmatrix} \Delta \theta_{\text{ig}}(s) \\ \Delta V_{\mu G}(s) \end{bmatrix} = \begin{bmatrix} \mathcal{G}_{\theta\omega}(s) & \mathcal{G}_{\theta V}(s) \\ \mathcal{G}_{V\omega}(s) & \mathcal{G}_{VV}(s) \end{bmatrix} \begin{bmatrix} \Delta \omega_{\text{adj}}(s) \\ \Delta v_{\text{adj}}(s) \end{bmatrix}, \quad (4.12)$$

where, $\mathcal{G}_{\theta\omega}(s) = \mathcal{G}_{11}(s) + s\mathcal{G}_{12}(s)$, $\mathcal{G}_{\theta V}(s) = \mathcal{G}_{13}(s) + s\mathcal{G}_{14}(s)$, $\mathcal{G}_{V\omega}(s) = \mathcal{G}_{21}(s) + s\mathcal{G}_{22}(s)$, $\mathcal{G}_{VV}(s) = \mathcal{G}_{23}(s) + s\mathcal{G}_{24}(s)$ with the the state-space to transfer function conversion of (4.10) and (4.11) of the linearized system as follows:

$$\begin{bmatrix} \mathcal{G}_{11} & \mathcal{G}_{12} & \mathcal{G}_{13} & \mathcal{G}_{14} \\ \mathcal{G}_{21} & \mathcal{G}_{22} & \mathcal{G}_{23} & \mathcal{G}_{24} \end{bmatrix} = \mathbf{C}_{\mu G} [s\mathbf{I} - \mathbf{A}_{\mu G}]^{-1} \mathbf{B}_{\mu G} + \mathbf{D}_{\mu G}.$$

The impact of Δv_{adj} w.r.t. the impact of $\Delta \omega_{\text{adj}}$ on $\Delta \theta_{\text{ig}}$ is significantly less as shown in Bode magnitude plot of $\mathcal{G}_{\theta\omega}(s)$ and $\mathcal{G}_{\theta V}(s)$ in Fig. 4.5 (a). Similarly, the impact of $\Delta \omega_{\text{adj}}$ w.r.t. the impact of Δv_{adj} on $\Delta V_{\mu G}$ is significantly less as shown in Bode magnitude

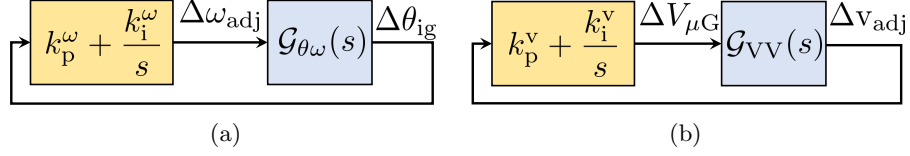


Figure 4.6: The closed loop control systems of the proposed synchronization with, (a) phase and frequency, (b) voltage synchronization control loop.

Table 4.2: Parameters of the Synchronization Controller

Types	Value
Control Module	$k_p^\omega = k_p^v = 0.4, k_i^\omega = k_i^v = 0.01$
Switch Initiation Module	$[dv_{\text{low}} \ dv_{\text{high}}] = \pm 1V, [d\phi_{\text{low}} \ d\phi_{\text{high}}] = \pm 1^\circ$ $[d\omega_{\text{low}} \ d\omega_{\text{high}}] = \pm 3.77\text{rad/sec}, t_{\text{wait}} = 1 \text{ sec}$

plot of $\mathcal{G}_{V\omega}(s)$ and $\mathcal{G}_{VV}(s)$ in Fig. 4.5 (b). In other words, $\|\mathcal{G}_{V\omega}(s)\|_\infty / \|\mathcal{G}_{\theta\omega}(s)\|_\infty \ll 1$ and $\|\mathcal{G}_{V\omega}(s)\|_\infty / \|\mathcal{G}_{VV}(s)\|_\infty \ll 1$. Therefore, the cross-coupling transfer functions of (4.12) can be neglected and as a result Fig. 4.6 (a) and Fig. 4.6 (b) show the closed loop control systems with the linearized plant model and the proposed synchronizing PI controllers. The controller parameters are tuned based on classical rule of thumb (i.e. $\pm 6\text{dB}$ of gain margin and $\pm 45^\circ$ of phase margin) followed in designing a stabilizing controller with enough robustness. The resulting controller parameters are tabulated in Table. 4.2.

4.5 HIL Demonstrations and Results

To substantiate the proposed synchronization method, a controller hardware-in-the-loop (CHIL)-based simulation study is conducted on OP5700 RT-simulator manufactured by OPAL-RT as shown in Fig. 4.7. The microgrid test system under study is developed based on a 55 bus urban distribution network [139]. For demonstration, the synchronizing switch is employed near the grid connection and 6 GFM inverters are connected at various buses in the network. The proposed active synchronization method is employed inside “Synch” that transmits the required adjustments to all the GFM inverters as shown in Fig. 4.7. Parameters of the network and the GFM inverters are tabulated in Table 5.4. Control of 2 GFM inverters are realized on low-cost Texas-Instruments TMS28379D Delfino controller boards. Two case-studies are demonstrated as follows.

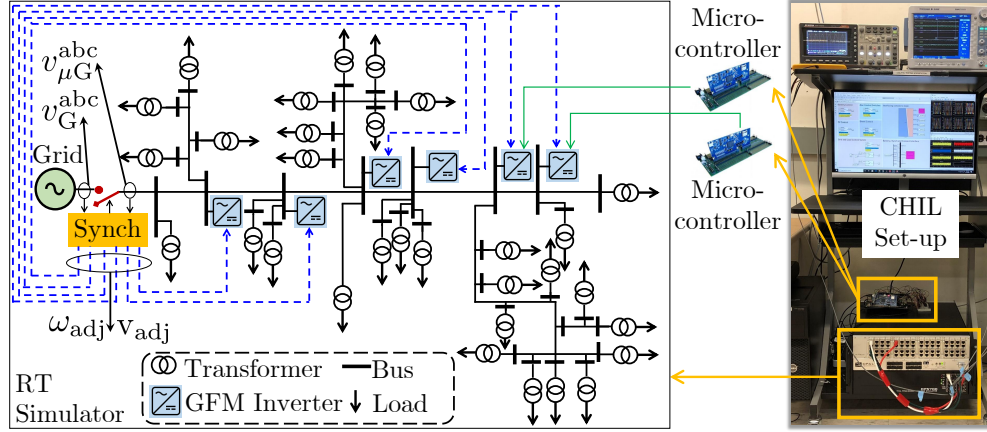


Figure 4.7: Hardware-in-the-loop-based simulation platform with 55-bus microgrid system and Texas Instruments Delfino TMS320F28379D controller board.

Table 4.3: Parameters of the Test System Under Study	
Inverter	Value
Ratings (3- ϕ)	480 V, 60 Hz, 100 kVA
Filter Parameters	$L_f = 150$ mH, $R_f = 50$ m Ω , $C_f = 110$ μ F
Other Parameters	$V_{dc} = 1000$ V, $f_{sw} = 20$ kHz, $\tau_S = 0.025$ s
Droop	$n_P = 5$ mHz/kVA, $m_Q = 0.48$ V/kVA
Network	Value
Ratings (3- ϕ)	480 V, 60 Hz, 500 kVA
Transformer	3- ϕ , 60 Hz, 5 kVA, 480/208 V

- **Case 1** : To showcase the instantaneous voltage waveforms that are synchronized by the proposed method.
- **Case 2** : To showcase the dynamic performance and impacts on system operation during the synchronization.

In the demonstration of **Case 1**, the microgrid was initially running in islanded mode by supplying all the loads by 6 GFM inverters. At $t \approx 5.3$ s, the synchronization process is initialized and at $t \approx 5.8$ s, the synchronization is completed and the synchronizing switch connects the microgrid with the grid after a waiting time, t_{wait} , of 0.2 s. The instantaneous waveform of v_{Grid}^a and $v_{\mu Grid}^a$, v_{Grid}^b and $v_{\mu Grid}^b$, v_{Grid}^c and $v_{\mu Grid}^c$ and instantaneous waveform of current through synchronization switch during connection, i_{Switch}^{abc} , are shown in Fig. 4.8(a), Fig. 4.8(b), Fig. 4.8(c) and Fig. 4.8(d) respectively.

The zoomed waveforms are shown during the synchronization process and it shows that $v_{\mu\text{Grid}}^a$, $v_{\mu\text{Grid}}^b$ and $v_{\mu\text{Grid}}^c$ are dynamically being aligned quickly with v_{Grid}^a , v_{Grid}^b and v_{Grid}^c respectively by the synchronization method with the help of adjustments in the droop curves of all GFM inverters simultaneously. It is observed here that the voltage waveforms of microgrid side during the synchronization process have minimal distortions and the transients during final re-connection of the microgrid with the main grid is smooth and seamless. The current waveform through the synchronization switch during connection shows that the magnitude jump is under a reasonable upper limit during transients.

Similarly in **Case 2**, the microgrid was initially running in islanded mode by supplying all the loads by 6 GFM inverters. At $t = t_{\text{Start}} \approx 33.1$ s, the synchronization process is initialized and at $t = t_{\text{Syn}} \approx 35.2$ s, the synchronization is completed and the synchronizing switch connects the microgrid with the grid after a waiting time, t_{wait} , of 1 s. Fig. 4.9(a) shows the dynamic response of the frequency in Hz of the microgrid, $f_{\mu\text{Grid}}$, of the main grid, f_{Grid} , and the difference between them, Δf , respectively. Similarly, Fig. 4.9(b) shows the dynamic response of the RMS line-to-line voltage in PU of the microgrid, $V_{\mu\text{Grid}}^{\text{LL}}$, of the main grid, $V_{\text{Grid}}^{\text{LL}}$, and the difference between them, ΔV , respectively. Moreover, Fig. 4.9(c) shows the dynamic response of the phase angle difference, $\Delta\phi$, in degree between the microgrid and the main grid. As it is observed that before $t = t_{\text{Start}}$, the frequency and voltage magnitude of the islanded microgrid is less than frequency and voltage magnitude of the main grid. This results in negative values of Δf and ΔV that need to be compensated by the synchronization method. Moreover, due to non-zero value of Δf , $\Delta\phi$ will change continuously from -180° to 180° . Once the proposed synchronization method starts at $t = t_{\text{Start}}$, both Δf , ΔV and $\Delta\phi$ are regulated inside their respective limits within 1 s with the help of adjustments in the droop curves of all GFM inverters simultaneously. After a waiting period of 1 s, the microgrid is connected with the main grid at t_{Syn} . It is observed here also that the voltage waveforms of microgrid side during the synchronization process have minimal distortions and the transients during final re-connection of the microgrid with the main grid is smooth and seamless. Moreover, Fig. 4.10 and Fig. 4.11 show the active and reactive power flow before, during and after the synchronization process. It is observed that the dynamic adjustments of droop curves in all GFM inverters do not alter the system operation significantly and hence the active and reactive power supplied by the GFM inverters are not changing significantly during the synchronization process.

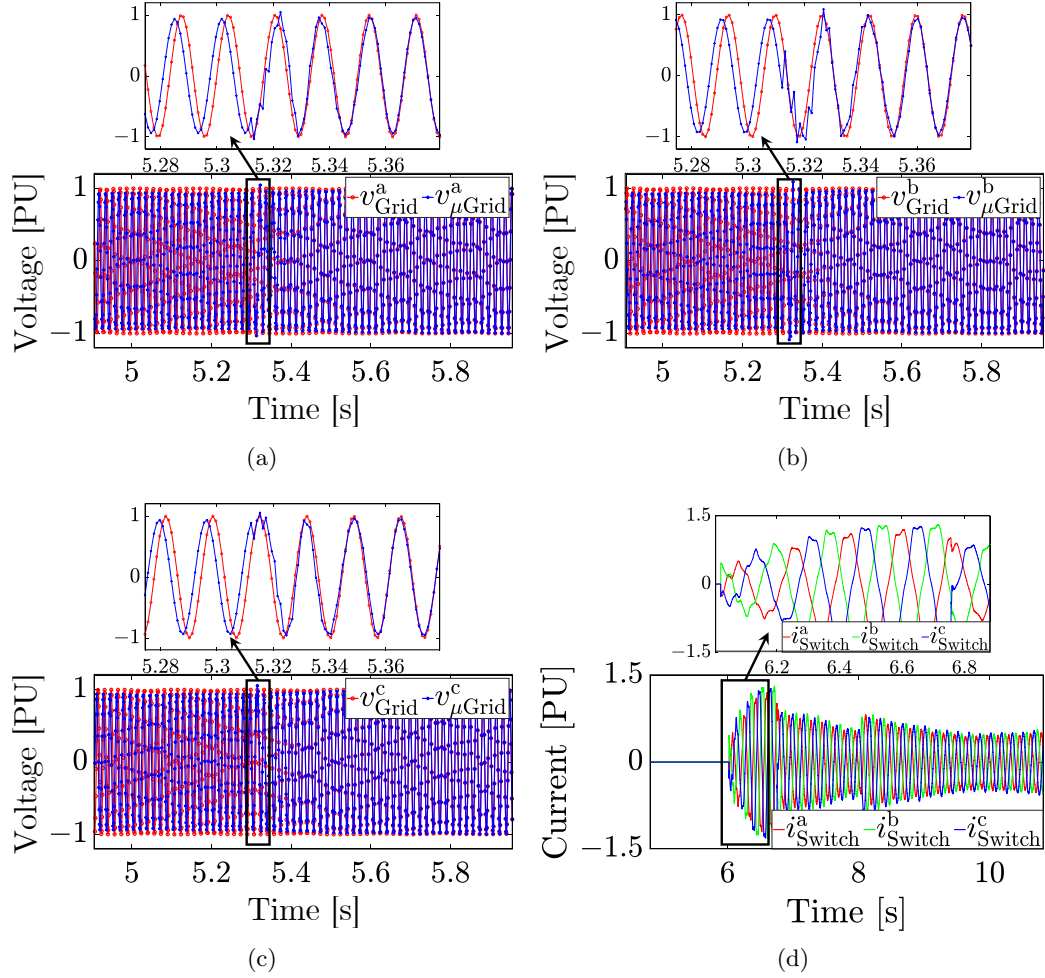
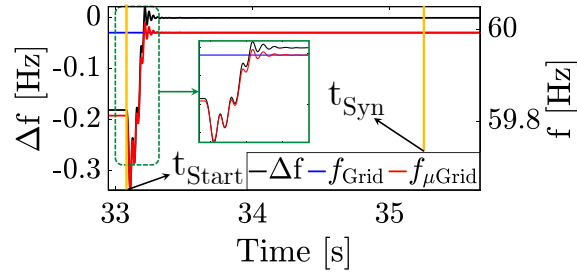
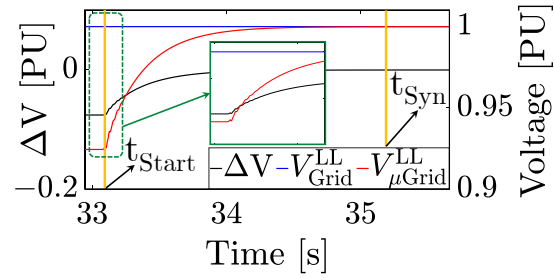


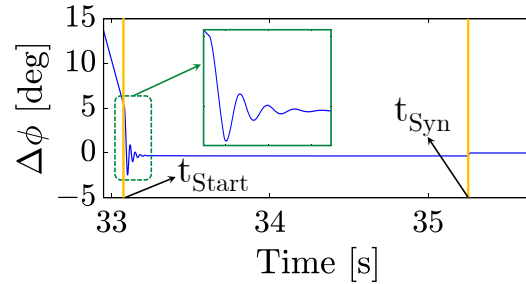
Figure 4.8: Instantaneous voltage waveform of grid side and microgrid side during synchronization process, (a) A-phase, (b) B-phase, (c) C-phase, and (d) instantaneous waveform of current through synchronization switch during connection after synchronization.



(a)



(b)



(c)

Figure 4.9: Dynamic behavior of, (a) frequency of the islanded microgrid and the main grid and the difference, (b) line-to-line voltage of the islanded microgrid and the main grid and the difference, (c) instantaneous phase angle difference between the islanded microgrid and the main grid, during synchronization.

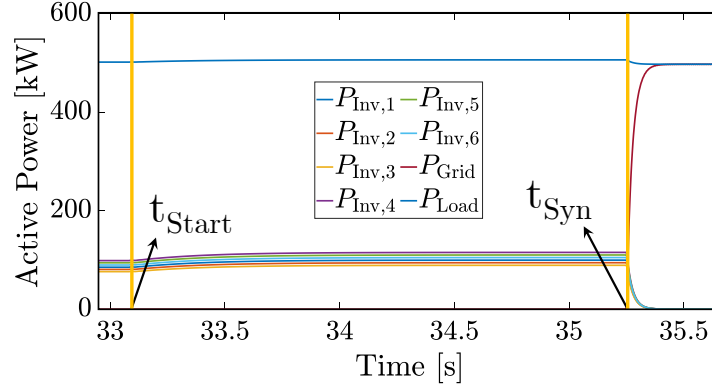


Figure 4.10: Active power output of 6 GFM inverters in the islanded microgrid, the main grid and the load during the proposed synchronization process.

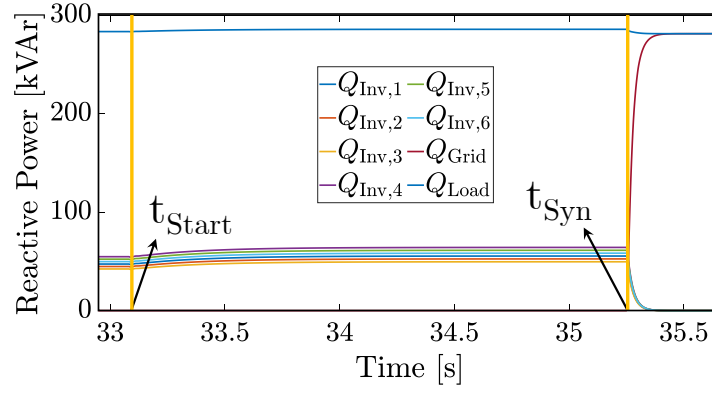


Figure 4.11: Reactive power output of 6 GFM inverters in the islanded microgrid, the main grid and the load during the proposed synchronization process.

Chapter 5

Resilient Microgrid: Design and Large Scale Hardware Demonstration

5.1 Gist of the Chapter

Seamless recovery and sustained power to critical infrastructures (CIs), after grid failure, is a crucial need arising in disaster scenarios that are increasingly becoming more frequent. Accreditation standards recommend CIs to have emergency power supply system (EPSS) in order to form a local microgrid network with backup resources (generation units/renewable resources) in case of sudden power blackouts of main grid supply. The standards also recommend enhancements such as seamless transition capability during on-grid to off-grid jump and vice versa, efficient and optimal operation while maintaining the standard regulation and power quality limits during power contingencies, and operation with sustenance of reliable power for long periods of time. This chapter is proposing a comprehensive design of the EPSS for uninterrupted operation of CIs by employing novel techniques, such as 1) mode-dependent droop controlled grid-forming inverters for seamless transition capability; 2) fast-acting optimal net-load management engine for efficient and optimal operations maintaining regulation and power quality limits; 3) optimization-based horizon of viability engine for longer, sustained and viable operation in the aftermath of grid failure as recommended in standards. Two stage validations are conducted using defined test cases while considering various types of

contingencies on a 3- ϕ , 60 Hz, 480 V, 450 kVA test system based on a realistic electrical network of a commercial scale medical center. The comprehensive controller-hardware-in-loop and power-hardware-in-the-loop based experiments corroborate the efficacy and robustness of the proposed strategy for EPSS employed in the CI. The mode-dependent droop controller enables seamless transition of the critical infrastructure between the off-grid and the on-grid modes. Inverters operate in the grid-forming mode when the CI is off-grid and also when the CI is on-grid while keeping frequency and voltage regulated between ± 0.5 Hz and $\pm 10\%$ of the nominal respectively. The fast-acting net-load management engine maintains the balance in total available generation and the power demand by determining the optimal decisions within 100 ms during various contingencies while regulating the voltage, frequency, and total harmonic distortion (THD) of the system under the strict limit recommended by the grid codes. The horizon of viability optimization engine showcases advantages over conventional scheduling engines with higher utilization of critical loads and maintaining higher energy at the end of an 1-hour experiment.

5.2 State-of-the-art on Emergency Power Supply Systems

Critical infrastructure (CI) describe essential assets of society, including medical centers, security service centers, food production and distribution centers, and communication facilities. Disruption of power to CI often results in a debilitating impacts on physical and economic security as well as public health and safety. The rapid recovery of power, possibly after a power blackout caused by increasingly frequent weather/climate disasters, for CI is a crucial need [109].

Accreditation standards, such as IEEE 602, require CI to have an emergency power supply system (EPSS) to form a local microgrid network with local generation and automatic transfer switches, in case of sudden power blackouts of the main grid supply [110]. Depending on the level of criticality and the urgency of the electrical loads, the EPSS is required to form the local microgrid within a specified amount of time to restore the operation of the CI. Various types of EPSS include Type-U characterizing uninterruptible EPSS, and Type-10 where the EPSS is allowed 10 s for recovery [140, 113]). Gas/diesel generator sets are traditional choices for most CI as local energy units for EPSS due to their sustained and robust power supply capability; however, the long startup time from standby mode makes the task of a seamless and rapid power restoration of CI difficult to achieve with these sources [115]. Battery energy storage units interfaced with

power electronic inverters provide uninterrupted power supply (UPS) system that are an alternate solution that enhances the ease in operation and reduces the response time of EPSS for CI. Accreditation standards, such as NFPA 111, recommend stored-energy EPSS (SEPSS), which employ batteries/fuelcells/ultracapacitors as the main energy harvesting units along with inverter topology to assist in restoring of power to CI in the case of the grid failure [111]. Briefly, these standards for CI recommend the following:

- To circumvent any short/temporary electrical interruptions to the critical and sensitive loads of the CI, Type-O/Type-U SEPSS are recommended, which possess the capability of seamless transition during the on-grid to off-grid jump and vice versa.
- The EPSS is recommended to efficiently operate the CI, by maintaining standard regulation and power quality limits and to operate optimally enabling smart and dynamic prioritizing of critical loads based on real-time situational awareness of the overall system.
- After the initial recovery from a power blackout, the EPSS is recommended to operate the CI while sustaining reliable power for long periods of time in such emergency situations.

In [141], a comprehensive literature review is conducted on service restoration through microgrid formation techniques for distribution networks. Planning and service restoration through microgrid formation, virtual power plants, grid-forming inverters and smart circuit breakers are some promising solutions for emergency power systems [142, 143, 144]. However, a thorough design and procedure for restoration of a microgrid network after power blackout are not discussed in these works. Moreover, being compliant with the grid regulation code during the restoration process is critical for the health of equipment in critical infrastructures. The role of power electronic converters in climate emergency is highlighted in [145]. References [146, 147, 148, 149, 150] develop emergency power systems that address prolonged power blackouts for various facilities such as hybrid advanced traction power supply system and medical centers. To the best of the authors' knowledge, there are no works present in the current literature that provide a comprehensive framework/strategy for EPSS to operate CI with the recommendations of NFPA 111 under emergency situations [151, 152, 153, 154]. This chapter proposes a comprehensive design of an EPSS for the uninterrupted operation of CI by

employing novel techniques to enhance the capabilities of the EPSS as recommended in the standards.

1. The seamless transition capability of the EPSS is achieved by using novel mode-dependent droop-controlled grid-forming (GFM) inverters [155]. The control strategy achieves the following objectives: 1) the seamless transition and recovery of power by the GFM inverters after the grid failure and 2) the seamless transition and re-synchronization by the GFM inverters after the grid returns. Further, the control strategy only requires information on the grid switch status for the mode transition. The advantage of the seamless transition capability, as proposed in [155], is that it does not require an additional control layer and parameter tuning to restore the voltage, frequency, and phase during transitions and synchronizations. Moreover, it ensures significantly less deviations in the output voltage/current of the inverters during the switching of its operating mode, unlike some existing works [91, 94, 102, 106]. The seamless control strategy of [155] is selected as a good candidate for this EPSS application.
2. To enhance the quality, resiliency, and reliability of the EPSS in the operation of the CI, this chapter proposes a centralized, high-speed, and cost-optimal coordination and allocation of loads during operational contingencies as an essential improvement, especially during off-grid mode of operation after the grid failure or during a complete black start operation of the CI [11, 12]. Unlike some existing works [156, 157, 158], the “net-load management” (NLM) engine of [11, 12] is superior in the optimal coordination process in applications with much larger numbers of net-load resources and fast-changing generation resources. It is also capable of accurately meeting a specific aggregate power set point to generation resources while achieving dispatch on a faster timescale and maintaining the standard regulation and power quality limits.
3. To guarantee the viability of a sustained, reliable supply of power over a substantial time horizon for the CI operation in the aftermath of the grid failure, an intelligent optimization scheme, the “horizon of viability” (HoV) engine, based on the cost-optimal portfolio of the locally available distributed energy resources (DERs) and the loads, is proposed and employed in the EPSS [13]. The advantage of the proposed HoV engine over existing works [159, 160, 161, 162], is that it ensures sustained viable operation of the CI for a longer time horizon by

supplying critical, sub-critical, and non-critical loads via using the backup DERs based on the historical load and solar irradiance forecast profiles and the real-time awareness of the run-time capabilities of the DERs.

This chapter assimilates three novel techniques for enhancing the capability of the EPSS of CI. These are as follows:

1. mode-dependent integrator-based droop-controlled grid forming inverters for seamless transition capability of the EPSS during on-grid to off-grid jump and vice versa [155],
2. fast-acting optimal net-load management engine for efficient and optimal operations of the EPSS, maintaining regulation and power quality limits throughout the operation [11],
3. optimization-based horizon of viability engine for longer, sustained and viable operation of the EPSS via local and emergency energy resources in the aftermath of grid failure [13].

The main challenge here, is to employ all three techniques and operate all of these simultaneously so that the combined functionalities of these novel techniques serve the purpose of enhancing the capability of the EPSS of the CI. Therefore, the main contribution of this chapter is to provide a systematic guidelines for implementing such novel techniques for EPSS of the CI and highlighting the key challenges in the implementation and operation of such large system. The proposed advancements for EPSS of a CI are validated in two stages using defined test cases while considering various contingencies. In the first stage, a controller hardware-in-the-loop (CHIL)-based real-time simulation study is conducted on a test system based on the realistic electrical network of a commercial-scale medical center. This stage of validation establishes the preliminary evidences showcasing the advantages of these novel techniques in the EPSS of a CI. In the second stage, a comprehensive power hardware-in-the-loop (PHIL)-based real-time study is conducted on the same electrical network. The test system under study is an R-L network of 3- ϕ , 60Hz, 480V and it has critical loads of 150kW, 72kVAr; sub-critical loads of 60kW, 30kVAr; and non-critical loads of 240kW, 115kVAr. The test system is equipped with two 3- ϕ , 125-kVA, 4-quadrant GFM inverters, two 3- ϕ , 125-kVA, 4-quadrant GFL inverters, and, one 3- ϕ , 135-kW diesel genset. The large-scale PHIL-based validation corroborates the fact that the employment of these novel techniques enhances the capabilities of the EPSS of a CI.

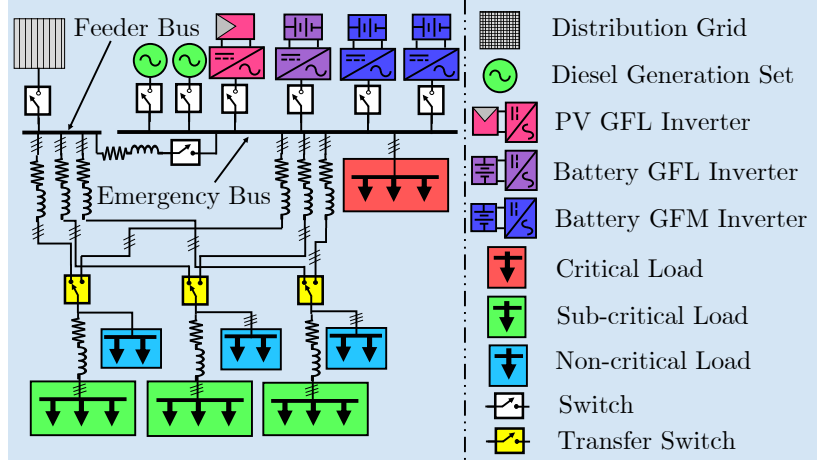


Figure 5.1: The power circuit of the critical infrastructure under study.

5.3 Description of the Existing EPSS of the CI

5.3.1 Existing Power Circuit and Operational Engine of the CI

The one-line diagram of the 3- ϕ power circuit of the CI is shown in Fig. 5.1. The considered CI has three classifications of loads, as defined in NFPA 99 [140], namely: i) critical load: loads that need continuous uninterrupted power supply irrespective of the grid availability (e.g., critical operation room, life-saving equipment in a medical center); ii) sub-critical load: loads that allow the temporary shutdown of services in case of grid failures (e.g., general patient room, HVAC system in a medical center); and iii) non-critical load: loads that allows temporary/permanent shutdown of services in case of grid failures (e.g., public lighting and elevators in a medical center). The building electrical network of the CI has a main feeder bus and an emergency bus and are connected to each other via a switch. Loads are connected to the buses via transfer switches. The grid is connected to the feeder bus via a feeder switch. At the emergency bus, multiple emergency DERs are connected via switches in parallel. They include two diesel generation sets, two GFM battery inverter systems, one solar inverter system, and one grid-following (GFL) mobile battery inverter system. The diesel generation sets are operated with load-frequency-based governor control and voltage-reactive power-based automatic voltage regulator control [117]. The solar and mobile battery inverter system are operated in GFL mode with a d-q axis-based conventional inner-current controller and globally/locally provided P-Q set points [15]. The battery inverter systems are

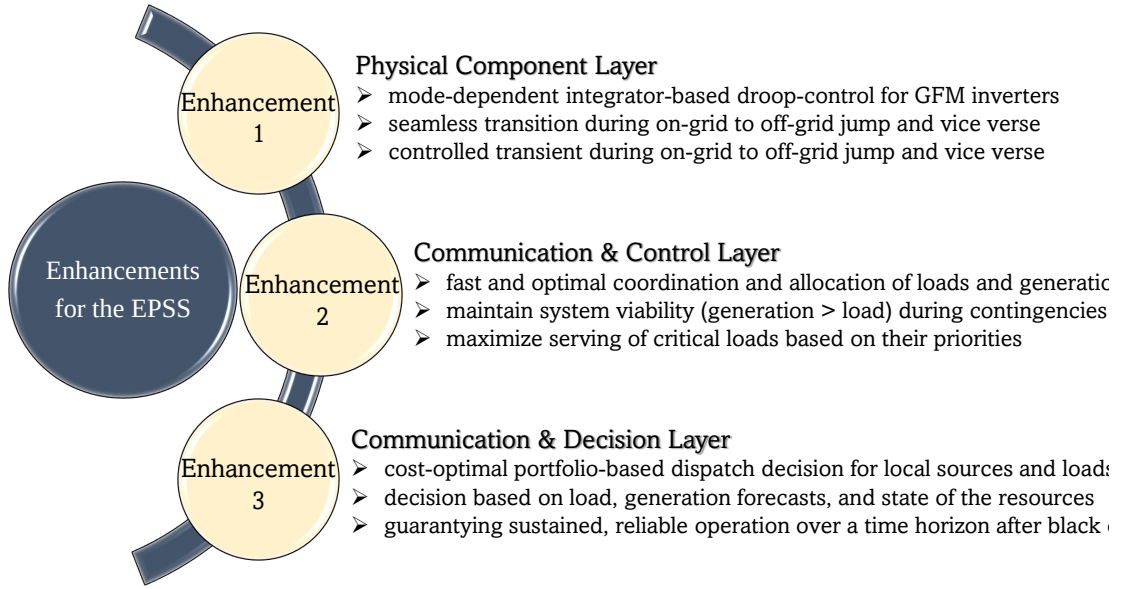


Figure 5.2: The methodology flowchart of the proposed enhancements for the emergency power supply system of a critical infrastructure.

operated in conventional GFM mode with d-q axis-based conventional outer-voltage-inner-current controller and P-f/Q-V-based droop control laws [163]. The existing EPSS of the CI enables the following:

- In on-grid mode, the grid at the feeder bus is the main power supply of the CI. At the emergency bus, apart from the solar GFL inverter, all other DERs are disconnected. While the main grid is available, the total demand of the CI (i.e., entire critical, sub-critical, and non-critical loads) is met. The EPSS enables the entire sets of load to be activated due to the availability of the grid.
- In off-grid mode, the DERs in the emergency bus are the main power supply of the CI. In case of a grid failure, the EPSS activates its emergency protocol for operation, and the operational engine starts priority-based load shedding, which is decided by the criticality of the loads. As a result, the EPSS supplies only the critical loads for sustained and viable emergency operation depending on the available rating of the DERs.

5.3.2 Challenges in the Existing EPSS for CI

The existing EPSS for CI has the following challenges:

- The existing EPSS is unable to provide a rapid and seamless power recovery (on-grid to off-grid transition and vice versa without electrical interruption) in case of a grid failure. The fundamental cause is the types of DERs and their control architectures. For example, traditional gas/diesel generator sets, while having sustained and robust power supply capability, have inherently long startup times from standby mode [113]. The GFM battery inverter system (not connected to the emergency bus during on-grid mode), while having a quick startup time, can supply power only after successful synchronization and connection of the switches to the emergency buses [111]; therefore, with the existing EPSS management for the CI, Type-U-based emergency recovery (seamless power restoration) of the entire CI is challenging.
- The existing EPSS load management engine is not designed for the sustained and viable operation of the CI in emergency situations. Currently, the operational engine follows a mere brute-force method to allocate loads based on their criticality for CI operation, depending on the available DER capability. As a result, it cannot operate the CI optimally (in the sense of the allocation of critical loads and longer viable operation of CI) during emergency situations. A centralized and intelligent control and monitoring method of the system, including sources and loads, to obtain the situational awareness and to take automatic corrective actions is lacking.

5.4 Proposed Enhancements of the EPSS for CI

To overcome the challenges mentioned in Section 5.3.2, this chapter proposes various enhancements to be employed in the EPSS for the CI. The proposed enhancements in the EPSS aims to achieve (i) seamless transitions of the CI from on-grid to off-grid transition and vice versa, (ii) centralized load management control to optimally manage the CI operation, (iii) intelligent load management for longer viable operation during emergency situations. The methodology flowchart of the proposed enhancements for EPSS of a CI is shown in Figure 5.2. The proposed enhancements are further described as follows:

- **Enhancement 1:** Ensuring seamless transitions (i.e., without any electrical interruption to the loads of the CI) from on-grid to off-grid and from off-grid to on-grid is an essential improvement for the uninterrupted operation and health of sensitive loads of the CI during emergency situations. As mentioned in Section 5.3.2, the existing circuitry and control architectures of the DERs are unable to provide seamless transition capability. The modification in the droop control law of the GFM battery inverters, as mentioned in Section 5.5.1, is proposed to achieve seamless transition capability [155].
- **Enhancement 2:** Ensuring fast and optimal coordination and allocation of loads during operational contingencies in the CI based on its criticality is another essential improvement for the EPSS, especially during off-grid mode of operation after the grid failure or during a complete black-start operation of the CI. The employment of the proposed centralized NLM engine, as mentioned in Section 5.5.2, ensures enhancing the quality, resiliency, and reliability of the emergency power supply operation of the CI [11, 12].
- **Enhancement 3:** Guarantying the viability of a sustained, reliable supply of power over a substantial time horizon is crucial for CI operation in the aftermath of a grid failure caused by increasingly frequent catastrophic events. An intelligent optimization scheme based on the cost-optimal portfolio of the locally available DERs and the loads, leveraged by the increasing expertise in accurate load and generation forecasts, is employed in the EPSS of the CI to achieve the target [13], as mentioned in Section 5.5.3.

5.5 Description of the Proposed Modifications

5.5.1 Modified Droop Law for Seamless Transition Capability

The proposed control strategy of the inverters aims to achieve the following objectives: 1) regulate the output active and reactive power by the droop-based battery inverters to a desired value while operating in on-grid mode, 2) perform seamless transition and recovery of power injections into the critical loads after a grid failure by the inverters operating in GFM mode all the time, 3) require minimal, single-bit information on the grid status for the mode transition of droop control. The design guidelines and the control architecture are briefly described below. Reader can find detailed descriptions in

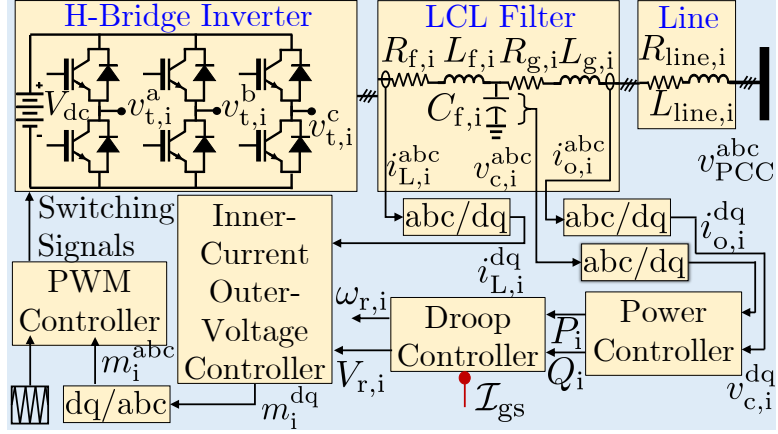


Figure 5.3: The power circuit and various control loops for the i^{th} GFM inverter.

reference [155]. The power circuit of the i^{th} , 3- ϕ H-bridge inverter is shown in Fig. 5.3. It is connected to the network at point-of-common-coupling via an LCL filter and a coupling line. The control architecture, as shown in Fig. 5.4, is described here.

1) Power controller: It determines the averaged active, P_i , and reactive power, Q_i , of the i^{th} GFM inverter as follows:

$$P_i = \frac{1}{\tau_{S,i}s + 1} p_i, \quad (5.1)$$

$$Q_i = \frac{1}{\tau_{S,i}s + 1} q_i, \quad (5.2)$$

where,

$$p_i := \frac{2}{3} [v_{o,i}^d i_{o,i}^d + v_{o,i}^q i_{o,i}^q], \quad (5.3)$$

$$q_i := \frac{2}{3} [v_{o,i}^q i_{o,i}^d - v_{o,i}^d i_{o,i}^q], \quad (5.4)$$

where, $v_{o,i}^d$, $v_{o,i}^q$, $i_{o,i}^d$, $i_{o,i}^q$ are the dq-components of output voltage and current and $\tau_{S,i} \in \mathbb{R}_{>0}$ is the time constant of the low-pass filters.

2) Droop controller: The mode-dependent droop controller for the GFM inverter for both the on-grid and off-grid operation of the CI is shown in Fig. 5.4. The proposed

droop law is as follows:

$$\omega_{r,i} = \omega_N - n_i P_i, \quad (5.5)$$

$$V_{r,i} = V_N - m_i Q_i - \mathcal{I}_{gs} k_i \int Q_i dt, \quad (5.6)$$

where, ω_{nom} , V_{nom} are the nominal frequency (in rad/s) and voltage set-point (in volt) of the system, respectively; n_i , m_i are the coefficients of the P - f and Q - V droop law, k_i is the integral coefficient of the Q - V droop law. $\mathcal{I}_{gs} = 1$ and $\mathcal{I}_{gs} = 0$ in on-grid and off-grid mode respectively. $\mathcal{I}_{gs}$ ensures that the integral action in the Q - V droop is effective only in on-grid mode. This droop law ensures that the GFM inverters in on-grid mode supply no reactive power in steady state. The addition of dependency on the variable, $\mathcal{I}_{gs}$, facilitates the GFM inverters seamless functionality for CI during the transition of on-grid to off-grid and off-grid to on-grid modes. The values of n_i and m_i are typically chosen such that $\omega_{r,i}$ and $V_{r,i}$ are within the allowed specification, defined by IEEE 1547 Standard [71], for all $P_i \in [0, P_{rated,i}]$ and $Q_i \in [-Q_{rated,i}, Q_{rated,i}]$ respectively. Here, $P_{rated,i}$ and $Q_{rated,i}$ are the rated active and reactive powers that can be delivered by the i^{th} GFM inverter. The saturation limit in the integrator of the Q - V droop law ensures that the entire system remains viable under compromised scenarios including significant delay in communicating $\mathcal{I}_{gs}$ which results in undesirable active integral operation during off-grid mode. A good candidate for selecting the magnitude of the upper and lower saturation limit of the integrator, $\pm\Delta V$, is $\pm 10\%$ of the nominal voltage rating as recommended by limits of IEEE 446 [115].

3) Inner-current-outer-voltage controller: The conventional inner-current-outer-voltage controller is employed here, as shown in Fig. 5.4. For the inner-current controller, a proportional-integral (PI) compensator with parameters $k_{pc,i}$ and $k_{ic,i}$, is used for tracking the reference of the dq-axis inductor current with the desired time constant, $\tau_{c,i}$ (typical range from 0.5 – 2 ms) [116]. Similarly, for the outer-voltage controller, a PI compensator, with parameters $k_{pv,i}$ and $k_{iv,i}$, is used to enable reference tracking with the desired phase margin and gain crossover frequency, based on the *symmetrical optimum* method [116]. Additional feed-forward signals facilitate the disturbance rejection capability of each layers.

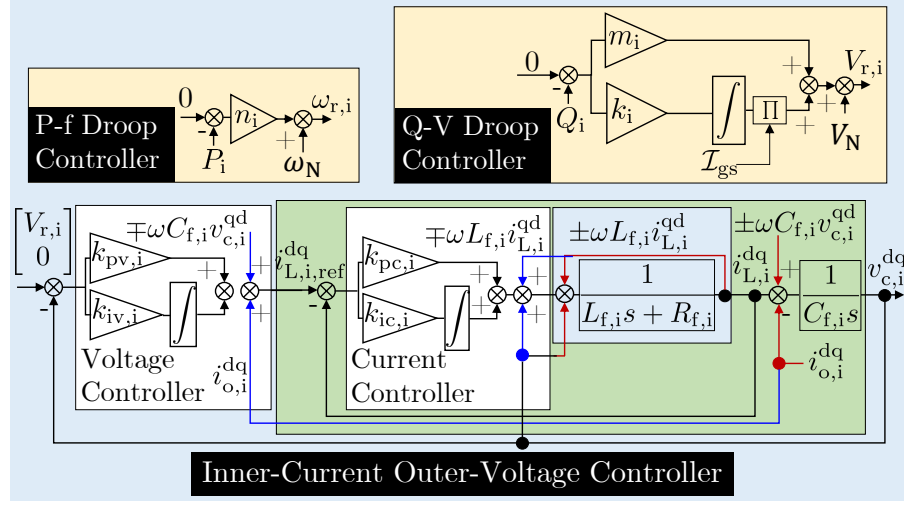


Figure 5.4: Inner-current-outer-voltage control loop with the proposed modified P-f, Q-V droop controller for the i^{th} GFM inverter.

5.5.2 NLM Engine for Reliable Operation

The NLM engine is a central dispatch control system that provides high-speed, cost-optimal coordination of a set of net loads (a combination of generation and controllable loads). The reader can find a more detailed description of the NLM engine in references [11, 12]. The high-level objectives of the NLM engine are to:

1. Maintain system viability (generation capacity $>$ load), including a specified generation reserve margin, by triggering dispatch actions each time the generation reserve margin falls below or exceeds the target range.
2. Minimize the overall cost (the sum of the unit cost change in the active power of the net load resource for all resources) of each dispatch action.
3. Maximize the amount of controllable load that is served using the available generation based on their priorities.

The system, called as the net-load-unit (NLU), contains an aggregated single uncontrollable (critical) load; N_d number of controllable (non-critical) loads; N_G number of solar and mobile battery-based GFL battery inverter system(s) and an aggregated single entity called the *slack bus* that comprises single/multiple GFM battery inverter systems and diesel generation sets. To meet these objectives, the NLM engine at any time instant, t_0 , performs the following functionalities:

1. The NLM engine receives measurements from the off-grid NLU and tracks the latest system state.
2. At each measurement update, the NLM engine evaluates the current power system operating margin (a comparison between the current operating power of the *slack bus*, $P_{\text{slk}}(t_0)$, and the largest *slack bus* capacity that can be provided, $|P_{\text{slk}}^{\text{max}}|$) and determines if a dispatch is needed to maintain:

$$|P_{\text{slk}}^{\text{max}}| - 2R_m \leq P_{\text{slk}}(t_0) \leq |P_{\text{slk}}^{\text{max}}| - R_m, \quad (5.7)$$

where R_m is the specified reserve margin. Specifically, if $P_{\text{slk}}(t_0) > |P_{\text{slk}}^{\text{max}}| - R_m$, then it commands a decrease of $P_{\text{slk}}(t_0)$. If $P_{\text{slk}}(t_0) \leq |P_{\text{slk}}^{\text{max}}| - 2R_m$, then it commands an increase in $P_{\text{slk}}(t_0)$.

3. If a set point is required to maintain the viability of the NLU with the margin (5.7), the NLM engine finds the optimal set of dispatch commands (on/off decisions for the controllable loads, dispatch set points to GFL inverters) that accomplishes these objectives using an optimization problem.

The optimization problem is given below:

$$\underset{x_j, \Delta P_{G,j}, \Delta P_{\text{slk}}}{\text{minimize}} \quad \sum_{j=1}^{N_d} c_{L,j} x_j P_{L,j} + \sum_{j=1}^{N_G} c_{G,j} |\Delta P_{G,j}| \quad (5.8)$$

subject to

$$\sum_{j=1}^{N_d} \bar{x}_j P_{L,j} + P_C - \sum_{j=1}^{N_G} (P_{G,j} + \Delta P_{G,j}) = P_{\text{slk}} + \Delta P_{\text{slk}} \quad (5.9)$$

$$\Delta P_{G,j}^{\text{down}} \leq \Delta P_{G,j} \leq \Delta P_{G,j}^{\text{up}}, \quad \forall j, \quad (5.10)$$

$$\Delta P_{G,j} = P_{G,j}(t) - P_{G,j}(t_0), \quad \forall j, \quad (5.11)$$

$$\Delta P_{\text{slk}} = P_{\text{slk}}(t) - P_{\text{slk}}(t_0), \quad (5.12)$$

$$x_j \in \{0, 1\}, \quad \forall j \quad (5.13)$$

$$\bar{x}_j = 1 - x_j, \quad \forall j, \quad (5.14)$$

where $c_{L,j}$ and $c_{G,j}$ are the participation unit costs for the j^{th} controllable load and the j^{th} GFL inverter in the NLU, respectively. x_j is the switch ON/OFF signal of the j^{th} controllable load. $P_{L,j}$ and $P_{G,j}$ are the active powers of the j^{th} controllable load and

the base active power rating of the j^{th} GFL inverter, respectively. $\Delta P_{G,j}$, $\Delta P_{G,j}^{\text{up}}$ and $\Delta P_{G,j}^{\text{down}}$ are the computed optimal headroom, and upward and downward headroom limits of the j^{th} GFL inverter, respectively. t and t_0 are the current time instant and the time instant when the contingency happens, respectively. $P_{\text{slk}}(t_0)$ and $P_C(t_0)$ are the active power capacity of the combined *slack bus* and uncontrollable load, respectively, at time instant t_0 . ΔP_{slk} is the total active power change of the *slack bus*.

5.5.3 HoV Engine for Sustainable Operation

Considering a grid interruption event and the CI is working in off-grid mode while being supported by its own available backup sources. The objective of the HoV engine is to sustain a viable operation of the CI for a longer time horizon by supplying critical, sub-critical, and non-critical loads by using only the available backup DERs. A trivial solution for the microgrid is to shut down all the sub- and non-critical loads of the CI and use the available DERs to supply only the critical loads (existing brute-force method of the EPSS of the CI); however, this might lead to a reduced quality of service in the CI due to no sub- and non-critical loads being served. Hence, there is a need to devise an optimal scheduling policy for the net-load resources in the CI. To achieve this objective, the optimization algorithm of the HoV engine determines a portfolio of generation and loads to guarantee the viability of all the critical loads while maximizing the level of service of the sub- and non-critical loads in the CI over a given time horizon [13].

A CI contains a collection of net-load resources: n_d controllable loads (i.e. sub- and non-critical loads), n_c non-controllable loads (i.e., critical loads); n_g diesel generator sets; n_p GFL solar inverter systems; n_b GFM battery inverter systems and a slack bus (combination of all the GFM inverter systems in the CI) to maintain stable voltage and frequency. Let $P_{r_j}^t$ denote the power output of the j^{th} grid-following solar inverter system at time t . The power delivered by the j^{th} diesel generator set and j^{th} GFL battery inverter system at time t are denoted by $P_{g_j}^t$ and $P_{b_j}^t$, respectively. Power delivered by the slack bus at time t is denoted by P_s^t . Here, each GFL battery inverter system, j , is characterized by maximum available energy capacity, E_{b_j} , and maximum discharging, W_{b_j} , and charging, $-W_{b_j}$, power transfer rates. The maximum power dispatch capacity of the j^{th} diesel generator set and the slack bus are denoted by W_{g_j} and W_s , respectively. The maximum available energy in the diesel generator, j , is denoted by E_{g_j} .

The HoV engine solves a convex optimization problem at each discrete time-instant,

t_0 , to determine the net-load scheduling decisions at some future time instant, $t_0 + t$, where $t \in \{1, 2, \dots, T\}$, with T being the horizon length. Denote by $x^t := [x_1^t, \dots, x_{n_d}^t] \in \{0, 1\}^{n_d}$, $P_g^t = [P_{g_1}^t, \dots, P_{g_{n_g}}^t] \in \mathbb{R}^{n_g}$, $P_b^t = [P_{b_1}^t, \dots, P_{b_{n_b}}^t] \in \mathbb{R}^{n_b}$, and $P_s^t \in \mathbb{R}$, $\mathbb{T}_{t_0} := \{t_0, \dots, T + t_0 - 1\}$. $s_{g_j}^{t_0}$ and $s_{b_j}^{t_0}$ are the state-of-energy (SOE) and state-of-charge (SOC) of the diesel generator set and the GFL battery inverter systems respectively, at time t_0 . Let $\tilde{s}_{b_j}^{t_0} := -(1 - s_{b_j}^{t_0})$. The optimization problem is given by:

$$\begin{aligned}
& \underset{\pi}{\text{minimize}} \quad \mathcal{C}(x, p_g, p_b, s_b^{t_0}, s_g^{t_0}), \quad \text{subject to :} \\
& 0 \leq \sum_{t=t_0}^{T+t_0-1} P_{g_j}^t \leq s_{g_j}^{t_0} E_{g_j}, \quad \text{for } j = 1, \dots, n_g, \\
& \tilde{s}_{b_j}^{t_0} E_{b_j} \leq \sum_{t=t_0}^{T+t_0-1} P_{b_j}^t \leq s_{b_j}^{t_0} E_{b_j}, \quad \text{for } j = 1, \dots, n_b, \\
& 0 \leq P_{g_j}^t \leq W_{g_j}, \quad \text{for } j = 1, \dots, n_g, t \in \mathbb{T}_{t_0}, \\
& -W_{b_j} \leq P_{b_j}^t \leq W_{b_j}, \quad \text{for } j = 1, \dots, n_b, t \in \mathbb{T}_{t_0}, \\
& 0 \leq P_s^t \leq W_s, \quad \text{for } t \in \mathbb{T}_{t_0}, \\
& \sum_{j=1}^{n_d} x_j^t \hat{P}_{L_j}^t + \sum_{j=1}^{n_c} \hat{P}_{c_j}^t = \sum_{j=1}^{n_g} P_{g_j}^t + \sum_{j=1}^{n_b} P_{b_j}^t \\
& \quad + \sum_{j=1}^{n_p} \hat{P}_{r_j}^t + P_s^t, \quad \text{for } t \in \mathbb{T}_{t_0}, \\
& x_j^t \in \{0, 1\}, \quad \text{for } j = 1, \dots, n_d, \quad \text{for } t \in \mathbb{T}_{t_0},
\end{aligned}$$

where the objective function is:

$$\begin{aligned}
\mathcal{C}(x, p_g, p_b, s_b, s_g) = & \sum_{t=t_0}^{T+t_0-1} \left[\sum_{j=1}^{n_b} c_{b_j}(s_{b_j}^{t_0}, P_{b_j}^t) \right. \\
& \left. - \sum_{j=1}^{n_g} c_{g_j}(\eta_{g_j}(P_{g_j}^t)) + \sum_{j=1}^{n_d} c_{\ell_j}(1 - x_j^t) \hat{P}_{L_j}^t \right].
\end{aligned}$$

The dispatch decisions are based on renewable generation forecasts, predicted profiles of the loads in the CI, and information about the latest SOC and SOE (the amount of energy equivalent fuel remaining relative to its reserve capacity), of the GFL battery inverter systems and diesel generator sets, respectively. This process is repeated at the subsequent time instants with updated sets of measurements. More details about the constraints and the objective function design of the HoV engine optimization problem can be found in [13].

5.6 Designed Test Cases for the EPSS of the CI

This section describes three test cases that are designed based on the challenges and the proposed modifications of the EPSS for the CI.

5.6.1 Test Case 1:

Test Case 1 is described in Table 5.1. Test Case 1 is designed to:

- Test the capability of having seamless transitions from on-grid to off-grid mode of the CI after a grid interruption with the help of the proposed modification in the GFM inverters, as described in Section 5.5.1, while serving critical loads.
- Test the operation of the CI with successive increases in generation through the available diesel generation sets and by serving more critical loads while maintaining stringent power qualities and regulations with the NLM engine, as described in Section 5.5.2.
- Test the capability of the CI to prioritize critical loads with the NLM engine, as described in Section 5.5.2, when solar irradiance is reduced while maintaining standard power qualities and regulations.

Table 5.1: Test case designed to demonstrate the capability of the proposed epss for a ci to have seamless transitions from on-grid to off-grid mode after grid failure and viable operation by serving critical loads via local energy resources.

Time [s]	Primary Source	Load Condition	Event Description
$t = 0$	Grid + PV GFL Inverter = 33kVA	Critical Load = 233kVA +Non-critical Load = 266kVA	<u>Initial Condition</u> • Grid and PV inverter supplies the entire load of the CI. • GFM Inverters are always connected to not supplying power. <u>Occurrence of Event</u> • Anomaly in grid is detected, grid disconnected.
$t = t_1$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA	Critical Load = 166kVA	<u>Initial Condition</u> • Diesel Gen – Sets are brought back online <u>Occurrence of Event</u> • Seamlessly transferring the load from ongrid to offgrid • achieve generation demand balance by NLM Engine.
$t = t_2$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA + Diesel Gen – Set = 67kVA	Critical Load = 233kVA	<u>Initial Condition</u> • Battery, PV Inverters are supplying loads initially. • Diesel Gen Sets achieve synchronous speed. <u>Occurrence of Event</u> • Diesel Gen-Sets connected, power balance by NLM.
$t = t_3$	Battery GFM Inverter = 133kVA + PV GFL _{inverter} = 6.7kVA + Diesel Gen – Set = 67kVA	Critical Load = 206.7kVA	<u>Initial Condition</u> • Battery, PV Inverters and Diesel Gensupply loads initially. <u>Occurrence of Event</u> • Solar irradiance decreases, curtailing loads by NLM.

Table 5.2: Test case designed to demonstrate the capability of the proposed eps for a ci to have black-start operation and viable operation by serving critical loads via local energy resources.

Time [s]	Primary Source	Load Condition	Event Description
$t = 0$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA	Critical Load = 166kVA	<u>Initial Condition</u> • CI is in off – grid mode and completely de – energized. <u>Occurrence of Event</u> • Black – start by GFM, PV and NLM engine
$t = t_1$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA + Diesel Gen – Set = 67kVA	Critical Load = 233kVA	<u>Initial Condition</u> • Diesel Gen Sets are slow to achieve synchronous speed <u>Occurrence of Event</u> • Diesel Gen Sets brought online, NLM increases loads.
$t = t_2$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA + Diesel Gen – Set = 67kVA + Mobile Inverter = 67kVA	Critical Load = 300kVA	<u>Initial Condition</u> • Diesel Gen Sets, GFM and PV supply loads initially. <u>Occurrence of Event</u> • Mobile GFL brought online, NLM increases loads.
$t = t_3$	Battery GFM Inverter = 133kVA + PV GFL Inverter = 33kVA + Mobile Inverter = 67kVA	Critical Load = 233kVA	<u>Initial Condition</u> • Diesel Gen Sets, GFM, PV and Mobile GFL supply load <u>Occurrence of Event</u> • Diesel Gen Sets are tripped, curtailing load by NLM.
$t = t_4$	Grid + PV GFL Inverter = 33kVA	Critical Load = 233kVA + Non-critical Load = 266kVA	<u>Initial Condition</u> • Diesel Gen Sets, GFM, PV and Mobile GFL supply load <u>Occurrence of Event</u> • Grid returns, GFMs connected, not supplying power.

5.6.2 Test Case 2

Test Case 2 is described in Table 5.2. Test Case 2 is designed to:

- Test the capability of the EPSS to black-start the CI using local generation sources and using the fast-acting NLM engine, as described in Section 5.5.2, while maintaining stringent power qualities and regulations.
- Test the capability of the CI to demonstrate the performance of the NLM engine in tackling contingencies, such as tripping diesel generation sets, while maintaining power qualities and regulations according to standards.
- Test the capability of having seamless transitions from off-grid to on-grid mode of the CI after the grid is restored with the help of the proposed modification in the GFM inverters.

5.6.3 Test Case 3

Test Case 3 is designed to:

- Test the capability of the EPSS to operate the CI for a longer time horizon using the proposed HoV engine, as described in Section 5.5.3, compared with the Base Case engine, as described in Section V.A of [11].
- Showcase the advantage of the proposed HoV engine in the EPSS over the existing Base Case engine.

To evaluate the performance of the developed HoV engine, the following metrics are defined:

Total remaining energy (TRE) The TRE (in kWh) of the DERs in the CI (i.e., battery inverters and diesel generators) at the end of the time horizon $[0, \mathcal{T}]$ is defined as:

$$\text{TRE}(\mathcal{T}) := \sum_{j=1}^{n_g} s_{g_j}^{\mathcal{T}} E_{g_j} + \sum_{j=1}^{n_b} s_{b_j}^{\mathcal{T}} E_{b_j}. \quad (5.15)$$

Table 5.3: Metrics on Voltage, Frequency Regulation Limits and Power Quality Measures.

Metric Type		Metric Value
Regulation Limit	Frequency range	$\leq \pm 0.5$ Hz
	Voltage range	$\leq \pm 10\%$ pu
Power Quality Limit	THD in voltage	$\leq 8\%$ pu
	Individual distortion	$\leq 5\%$ pu

Cost-weighted energy (CWE) The total CWE delivered to all critical and deferrable loads in the microgrid over a time horizon $[0, \mathcal{T}]$ is defined as:

$$\text{CWE}(\mathcal{T}) := \int_0^{\mathcal{T}} \text{CWP} \, dt = \int_0^{\mathcal{T}} \sum_{i=1}^{n_d+n_c} c_{\ell_i} P_{L_i}^t \, dt, \quad (5.16)$$

where CWP is the cost-weighted power delivered; and $P_{L_i}^t$ is the per-unit power of load, i . The metrics in (5.15) and (5.16) capture the trade-off between the guaranteed viability of power for the CI units (higher remaining energy implies the availability of more energy in the generation resources of the CI units) and the level of service of the required loads of the CI (higher CWE implies that more higher-priority loads are running). Note that the performance of the proposed EPSS system in all three test cases will be examined based on the accredited standards on voltage and frequency regulation limits and power quality measures, i.e., total harmonic distortion (THD) and individual harmonic distortion, as recommended by [71, 115] as tabulated in Table 5.3.

5.7 CHIL-based Validation and Results

5.7.1 CHIL Configuration

A CHIL-based real-time (RT) simulation studies are conducted on a test-system under study that is developed based on the electrical network of a commercial-scale medical center [121]. The CI shown in Fig. 5.1 is an R-L network of 3- ϕ , 60Hz, 480V (line-to-line), and it has critical loads of 150kW, 72kVAR; sub-critical loads of 60kW, 30kVAR; and non-critical loads of 240kW, 115kVAR. The main feeder and the DERs of the CI are tabulated in Table 5.4. The entire CI network with the grid and DERs is emulated inside the OP5700 real-time simulator (RTS) manufactured by OPAL-RT

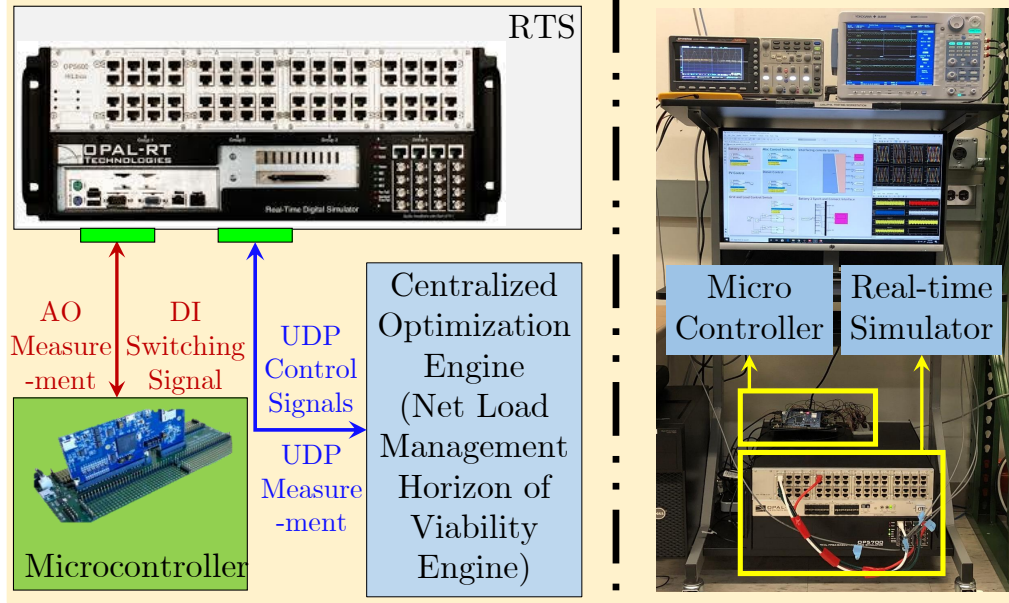


Figure 5.5: Controller-hardware-in-the-loop simulation set-up with real-time simulator interfaced with physical micro-controllers and server workstations.

using the eMEGASIM platform for electromagnetic transient simulation. The control loops of one GFL solar inverter system and one GFM battery inverter system are realized on two Texas Instruments TMS320F28379D, 16/12-bit floating-point 200MHz Delfino microcontroller boards. The NLM and HoV optimization engines are implemented on a server workstation, developed using Python 3.7.1, with commercial optimization solver, MOSEK. A device control gateway is realized using standard User Datagram Protocol (UDP), which emulates the communication layer for continuously listening to the measurements sent from the RTS and dispatching commands from the NLM/HoV engine to the RTS, as shown in Fig. 5.5. The CHIL experiment is conducted to evaluate the performance in Test Case 1, Test Case 2, and Test Case 3, and results are discussed next.

5.7.2 CHIL Simulation-based Validation and Results

The CHIL-based validation of Test Case 1 for 360s and corresponding results are shown in Fig. 5.6. The active and reactive power output of GFM battery-supported inverters are shown in Fig. 5.6(a), Fig. 5.6(b), respectively. Fig. 5.6(c), and Fig. 5.6(d) show the active and reactive power outputs of the grid and solar GFL inverter, respectively. It is

Table 5.4: 3-Phase Parameters of Grid Feeder and DERs of the CI

Inverter	Value of the Parameters
Ratings (3- ϕ)	480 V (L-L), 60 Hz, 125 kVA, 0.85 pf
DC Side	$V_{dc} = 850$ V, $f_{sw} = 10$ kHz
LCL Filter	$L_f = 150$ μ H, $L_g = 15$ μ H, $C_f = 110$ μ F
Droop Control	$n_i=4 \frac{\text{mHz}}{\text{kW}}$, $m_i=0.2 \frac{\text{volt}}{\text{kVAr}}$, $k_i=0.67 \frac{\text{volt/s}}{\text{kVAr}}$
GFM Control	$k_{pc,i}=0.45$, $k_{ic,i}=150$, $k_{pv,i}=0.1$, $k_{iv,i}=37.1$,
GFL Control	PLL's PI Regulator: $k_p^{\text{pll}}=0.05$, $k_i^{\text{pll}}=0.5$
Diesel Set	Value of the Parameters
Ratings (3- ϕ)	480 V (L-L), 60 Hz, 135 kVA, 0.85 pf
Droop Control	$n_g(\text{P-f}) = 4$ mHz/kW, $m_g(\text{Q-V}) = 0.2$ V/kVAr
Grid	Value of the Parameters
Ratings (3- ϕ)	13.8 kV fed Y-Y 0.75 MVA, 13.8/0.48 kV Xr.
Feeder Line	$R_{l,g} = 5$ m Ω , $L_{l,g} = 30$ μ H

observed that all the critical, sub-critical and non-critical loads of the CI are supplied by the grid and the solar GFL inverter alone in on-grid mode. With the help of the modified droop law employed to GFM inverters, the EPSS achieves seamless transition from on-grid to off-grid mode around $t \approx 180$ s. The diesel generation sets are brought online and synchronized after time $t_2 \approx 250$ s; Fig. 5.6(e) and Fig. 5.6(f) show the active and reactive power output of diesel gen-set, respectively. At $t_3 \approx 300$ s, the output power of the solar inverter curtails due to low irradiance. The NLM engine performs a fast-acting control to increase/decrease loads to maintain balance between generation and demand. Similarly, the CHIL-based validation of Test Case 2 for 540s and the corresponding results are shown in Fig. 5.7. The active and reactive power output of GFM battery-supported inverters are shown in Fig. 5.7(a), and Fig. 5.7(b). Fig. 5.7(c), and Fig. 5.7(d) show the active and reactive power outputs of the grid and solar GFL inverter, respectively. The NLM engine is making sure to supply the major critical load and some sub-critical loads during the initial black start condition and increases the load when the diesel generation sets are brought online at $t_1 \approx 300$ s and the GFL mobile battery inverter system is synchronized at $t_2 \approx 360$ s, as shown in Fig. 5.7(e) and Fig. 5.7(f), respectively and decrease load at $t_3 \approx 420$ s when the diesel generation sets are disconnected due to exhausting the fuel limits. At $t_4 \approx 480$ s, the grid returns, and the entire load is supplied alone by the grid and solar GFL inverter system. Fig. 5.6(g), Fig. 5.6(h), Fig. 5.6(g), and Fig. 5.6(h) show the voltage RMS,

the system frequency, and the instantaneous voltage measured at the emergency bus during Test Case 1 and Test Case 2, respectively and it is observed that all the metrics, as tabulated in Table 5.3, are satisfied during both test cases. Note that the GFM inverters, which are still connected to the emergency bus, have regulated the output active and reactive power by the enhancement in the droop curve, as mentioned in Section 5.5.1, once the grid comes online.

Figs. 5.8 demonstrates the performance of the HoV engine compared to the Base Case with respect to the TRE and CWE metrics, in (5.15) and (5.16), respectively. Fig. 5.8(a) demonstrates that the HoV engine provides more total CWE than the Base Case engine during the 1-hour horizon; therefore, the HoV engine provides power to more high-priority critical and sub-critical loads in a resource-constrained scenario during the time duration of the experiment. Fig. 5.8(b) provides a comparison based on the TRE. The HoV optimization engine optimally schedules the available generation sources that result in a higher TRE available at the end of 1 hour that can be used to serve the critical and sub-critical loads in the CI at a later time instant. This shows the advantage of the proposed HoV engine over the existing Base Case engine in the sense of extending the horizon of the viable operation of the CI. It is observed that, initially the HoV engine is inclined toward running more load in the CI, which is similar to Base Case engine; therefore, initially, the HoV engine is aggressive in terms of maintaining a good quality of service to the load, but it adjusts to conserve generation for future times of operation. During the end of the time horizon, when the SOC level of the batteries is low, the HoV engine jumps to a more energy-saving mode compared to the Base Case engine.

5.8 PHIL Based Validation and Result

5.8.1 PHIL Configuration

The PHIL-based experimental configuration and the large-scale laboratory-based hardware systems are shown in Fig. 5.9(a) and Fig. 5.9(b). The hardware under test of the PHIL setup has the following components:

- 1) two 3- ϕ , 125-kVA, 4-quadrant GFM inverters, manufactured by Dynapower; interfaced with Aerovironment AV-900 and NH Research 9300 bidirectional battery storage systems,
- 2) two 3- ϕ , 125-kVA, 4-quadrant GFL inverters, manufactured by Dynapower, two interfaced with Magna Power MT250-1000 unidirectional battery and PV emulator systems,

- 3) four 3- ϕ , 250-kW LoadTec OSW4C-based load banks,
- 4) one 3- ϕ , 135-kW Cummins C150D2RE diesel genset.

The control parameters of the GFM inverter, GFL inverter, and diesel generation units are tabulated in Table 5.4. The PHIL interface consists of the following:

- 1) one 3- ϕ , 270-kVA Ametek RS-series power amplifier,
- 2) three LEM DVL750 voltage transducers,
- 3) three LEM LF-350-S current transducers.

An Opal RT OP5700-series real-time simulator with a time step of $T_s = 50\mu s$ is used to execute the in-the-loop system, which consists of P - Q loads, the emulated grid, and a novel PHIL interface control block, as discussed in [164], which consists of a virtual GFM inverter model and three discrete-time PHIL interface compensators. The parameters of the controller and the power circuit of the virtual inverter are similar to the physical GFM inverter system under study with the same control parameters as tabulated in Table 5.4. PHIL experiment is conducted to evaluate the performance of Test Case 1 and Test Case 2, and, results are discussed next.

5.8.2 PHIL Simulation-based Validation and Results

Table 5.5: Resulting Values of the Metrics on Voltage, Frequency Regulation Limits, and Power Quality Measures in PHIL.

Metric		TC 1	TC 2	TC 3
Freq.	Max [$\uparrow$]	+0.03 Hz	+0.05 Hz	+0.04 Hz
Range	Max [$\downarrow$]	−0.31 Hz	−0.32 Hz	−0.29 Hz
Volt.	Max [$\uparrow$]	+0.9%	+0.8%	+0.9%
Range	Max [$\downarrow$]	−8.9%	−8.1%	−9.2%
THD	Max	6.3%	6.8%	6.7%
Volt.	Min	1.1%	1.5%	1.9%
Ind.	Max	4.3%	4.9%	4.7%
Dist.	Min	0.9%	1.2%	1.4%

The PHIL-based validation of Test Case 1 for 300s and the corresponding results are shown in Fig. 5.10. The active and reactive power flow of both physical and virtual GFM battery-based inverter systems during the events of Test Case 1 are shown in Fig. 5.10(a) and Fig. 5.10(b), respectively. As emphasized in the figures, the enhanced droop control for the GFM inverters, as described in Section 5.5.1, enables the EPSS to achieve seamless transition capability from on-grid to off-grid mode ($t_1 \approx 100s$)

with substantially less transients. Moreover, during on-grid mode, the output active and reactive power of both GFM inverters are regulated to the desired value (in this case, zero); therefore, during on-grid mode, all the critical, sub-critical, and non-critical loads of the CI are supplied by the grid and the solar inverter system alone, as shown in Fig. 5.10(c) and Fig. 5.10(d) respectively. With this seamless transition to off-grid mode, however, the GFM inverters along with the solar inverter systems supply the critical loads and some sub-critical loads based on the decision made by fast-acting NLM engine, as mentioned in Section 5.5.2. The NLM engine is able to make the decision of optimal loading based on available generation within 96ms. Diesel generation sets are brought online and synchronized after time $t_2 \approx 180s$, as shown in Fig. 5.10(e) and Fig. 5.10(f), respectively. Eventually, the NLM engine decides to add and supply more critical loads due to the increase in the available generation capability of the EPSS and is able to make the decision within 97ms. Moreover, at $t_3 \approx 240s$, the output power of the solar inverter curtails due to low irradiance, and the fast-acting NLM engine takes care of the contingency situation by curtailing some sub-critical loads to maintain balance between generation and demand. It is observed that during every contingencies, the fast acting NLM engine is able to find the optimal solution of loading by 100ms based on generation capability.

Similarly, the PHIL-based validation of Test Case 2 for 300s and the corresponding results are shown in Fig. 5.11. The active and reactive power flow of both the physical and virtual GFM battery-based inverter systems during the events of Test Case 2 are shown in Fig. 5.11(a) and Fig. 5.11(b), respectively. With the advanced and fast-acting NLM engine decision within 97.14ms and GFM capability, the major critical load and some sub-critical loads are being supplied during the initial black-start condition. The diesel generation sets are brought online and synchronized at time $t_1 \approx 65s$, as shown in Fig. 5.11(e) and Fig. 5.11(f), respectively. Eventually, the NLM engine decides to add and supply more critical loads due to the increase in the available generation capability of the EPSS and the decision is made by 95ms. Moreover, at $t_2 \approx 120s$, the GFL mobile battery inverter system is synchronized and connected, and as a result, the fast-acting NLM engine adds and supplies more critical and sub-critical loads, as shown in Fig. 5.11(c) and Fig. 5.11(d), respectively. At $t_3 \approx 180s$, the diesel generation sets are disconnected as they exhaust the fuel limits, and the NLM engine tackles the contingency situation by curtailing some loads to maintain balance between generation and demand. In this major contingency, the NLM engine completes the optimal solution by 95ms. At $t_4 \approx 240s$, the grid returns, and all the loads of the CI are supplied alone by the grid

and solar inverter system once the grid is synchronized and connected to the CI. Note that the GFM inverters, which are still connected to the emergency bus, have regulated output active and reactive power by the enhancement in the droop curve, as mentioned in Section 5.5.1, once the grid comes back online.

Figs. 5.12 demonstrate the performance of the HoV engine compared to the Base Case with respect to the ratio of TRE and CWE metrics as mentioned in (5.15) and (5.16) respectively. These two figures show the run time ratios of the CWE and TRE metrics (ρ_{CWE} and ρ_{TRE}). Fig. 5.12(a) demonstrates that the HoV engine provides more total CWE than the base case engine over the 1 hour horizon (the ratio is more than 1); therefore, the HoV engine provides power to more high-priority critical and sub-critical loads in a resource-constrained scenario during the time duration of the experiment. Fig. 5.12(b) provides a comparison based on the total remaining energy. The HoV optimization engine optimally schedules the available generation sources that result in a higher total remaining energy available at the end of 1 hours that can be utilized to serve the critical and sub-critical loads in the CI at a later time instant. This shows the advantage of the proposed HoV engine over the existing base case engine in the sense of extending the horizon of the viable operation of the CI. It is observed that initially the HoV engine is inclined toward running more load in the CI that is similar to base case engine. Therefore, initially, the HoV engine is aggressive in terms of maintaining a good quality of service to the load, but it adjusts itself to conserve generation for future times of operation. However, during the end of the time horizon, when the SOC level of the batteries are low, the HoV engine jumps to more energy saving mode compare to the base case engine case. It can be observed that the HoV engine is equally conservative in restoring the energy of the batteries in the CI but more aggressive in terms of supplying a higher number of prioritized loads compared to the base case engine (indicated by $\rho_{CWE} > 1$ and $\rho_{TRE} \approx 1$). Yet, toward the end of the horizon, the HoV engine is more aggressive in restoring the energy of the batteries in the CI but equally aggressive in terms of supplying prioritized loads compared to the base case engine (indicated by $\rho_{CWE} \approx 1$ and $\rho_{TRE} > 1$).

5.9 Discussions on Results

Based on the initial CHIL-based, and the large scale PHIL-based demonstrations and the corresponding results, the following observations can be made:

1. During on-grid to off-grid and off-grid to on-grid transition of the CI, the integrator in the Q-V droop law ensures that the reactive power output of the GFM inverters are zero and the P-f droop law ensures the active power output is zero. The experimental data establishes the effectiveness of the always grid-forming operation and control of inverters in meeting power delivery objectives when on-grid and off-grid under various kinds of loads and scenarios while minimizing transients, i.e. frequency is kept under ± 0.5 Hz and the voltage is kept under $\pm 10\%$ of the nominal voltage rating, before, after and during transitions. These strict regulations on the frequency and voltage of the system are well under the limits recommended by IEEE standards [71, 110]. The novel mode-dependent droop controller for the GFM inverters enhances the seamless transition capability of the EPSS that is a significant improvement over the existing solutions for EPSS of CI.
2. During the test cases, various contingencies are tested during the operation of the EPSS of the CI. During major contingencies such as on-grid to off-grid jump, off-grid to on-grid jump, tripping of diesel gen-sets, drop in solar irradiance, etc., the fast-acting NLM engine was always able to maintain the balance in total available generation and the power demand by determining the optimal amount of loads and power set-points of GFL inverters. These optimal decisions were successfully determined within 100 ms during various cases, that is significantly fast in action in comparison with conventional EPSS for CI. The performance of the fast-acting NLM engine during these contingencies corroborate the fact that the operation of the EPSS employed with the NLM engine regulates the voltage and frequency of the system under the strict limit recommended by the grid codes [71, 110]. Table 5.5 shows that the maximum frequency deviation is -0.32 Hz and the maximum voltage deviation is -9.2% of the nominal voltage during all three test cases. These numbers are well within the recommended limits suggested by IEEE 1547 Std. Moreover, the maximum total harmonic distortion (THD) and the individual harmonic distortion limits of the voltage waveform at emergency bus are 6.7% and 4.9% during all three test cases, that follow the recommendations of the IEEE Std. 519 [73].

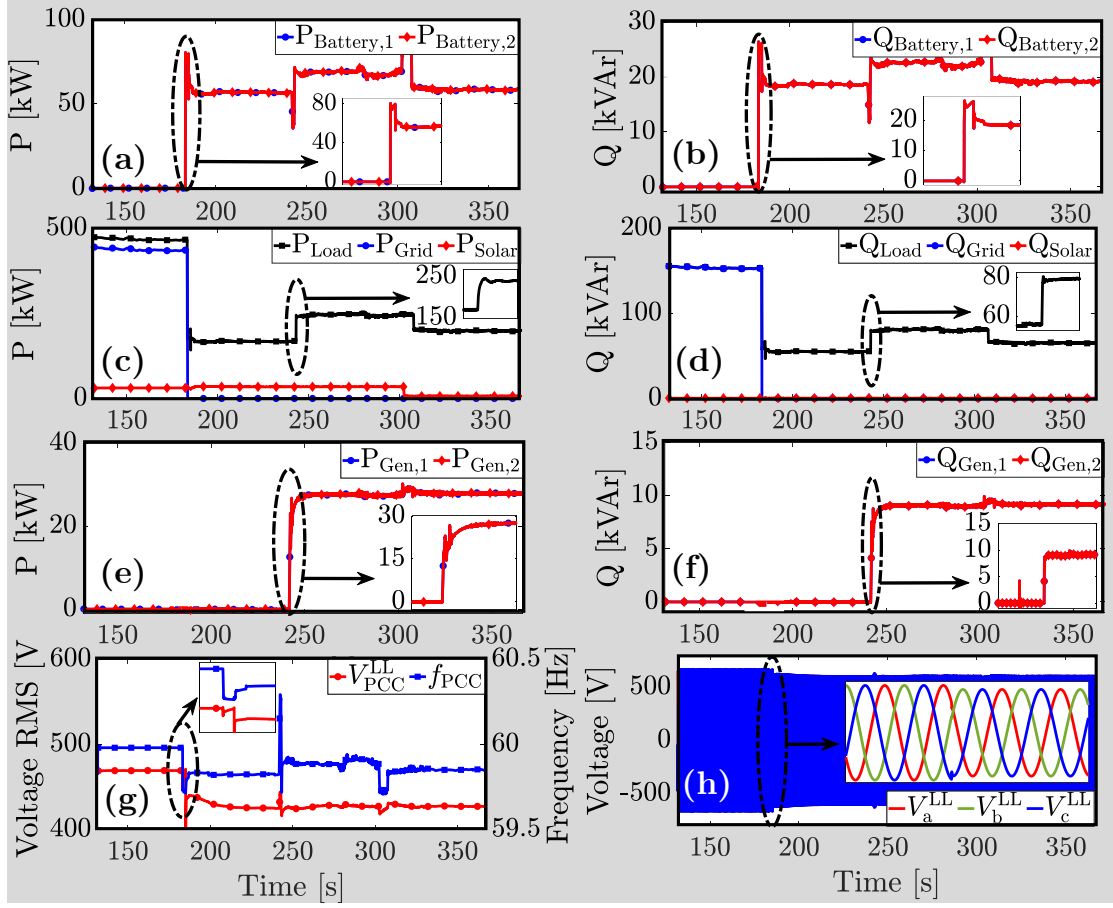


Figure 5.6: CHIL results of (a) active power output, (b) reactive power output of GFM battery inverter systems, (c) active power output, (d) reactive power output of the grid, loads, and solar inverter system, (e) active power output, (f) reactive power output of the diesel generation sets, (g) line-to-line voltage RMS and frequency, (h) instantaneous voltage waveform measured as emergency bus of the CI system of Fig. 5.1 during Test Case 1 of Table 5.1.

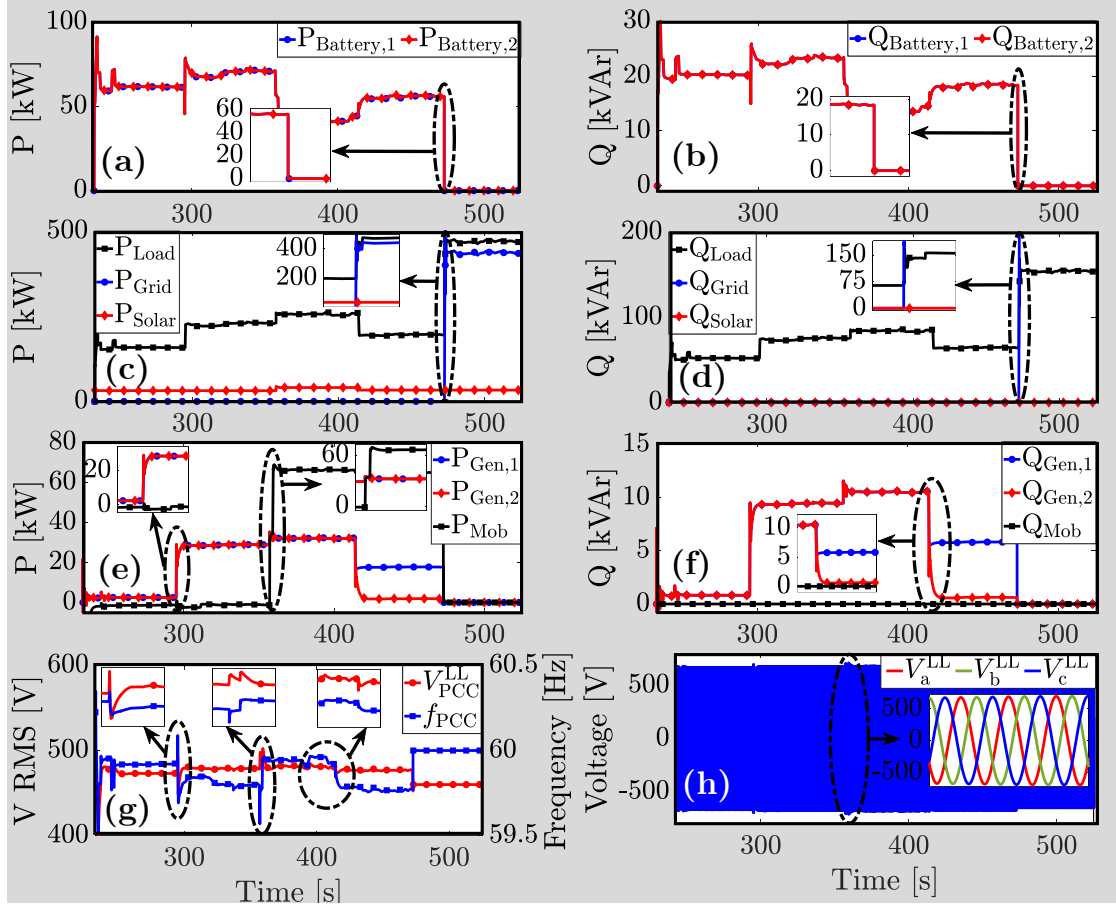


Figure 5.7: CHIL results of (a) active power output, (b) reactive power output of GFM battery inverter systems, (c) active power output, (d) reactive power output of the grid, loads, and solar inverter system, (e) active power output, (f) reactive power output of the diesel generation sets and mobile grid-following inverter system, (g) line-to-line voltage RMS and frequency, (h) instantaneous voltage waveform measured as emergency bus of the CI system of Fig. 5.1 during Test Case 2 of Table 5.2.

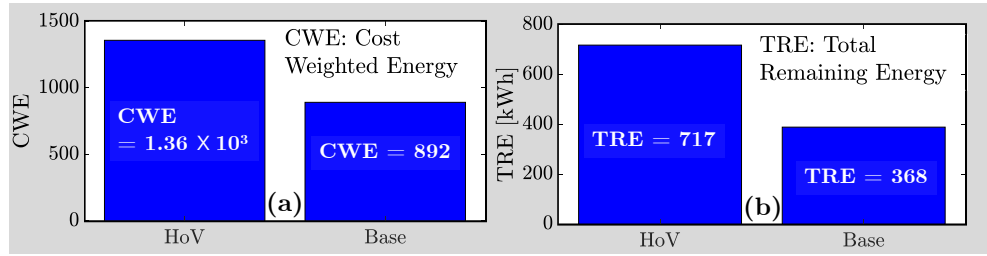
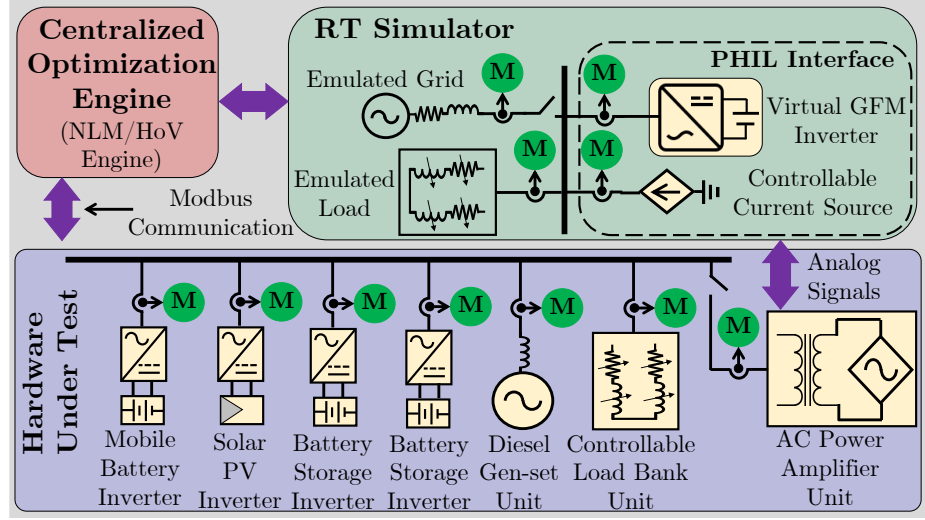
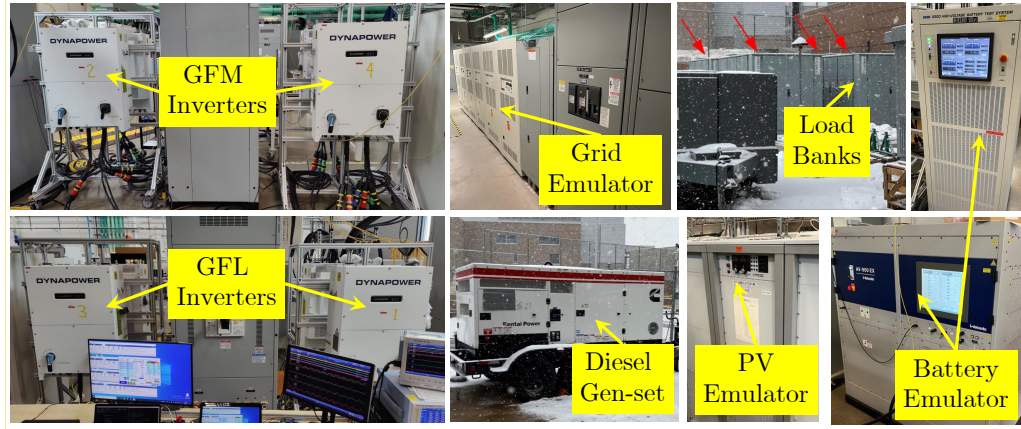


Figure 5.8: Comparative results of (a) total cost-weighted energy and (b) total remaining energy (in kWh), resulting from HoV engine and base case engine during 1 hours of the CHIL experiment on the test system of Fig. 5.1.



(a)



(b)

Figure 5.9: The diagram of (a) the entire power-hardware-in-the-loop configuration of the test system under study, (b) laboratory-based at-scale hardware setups with various types of sources and loads.

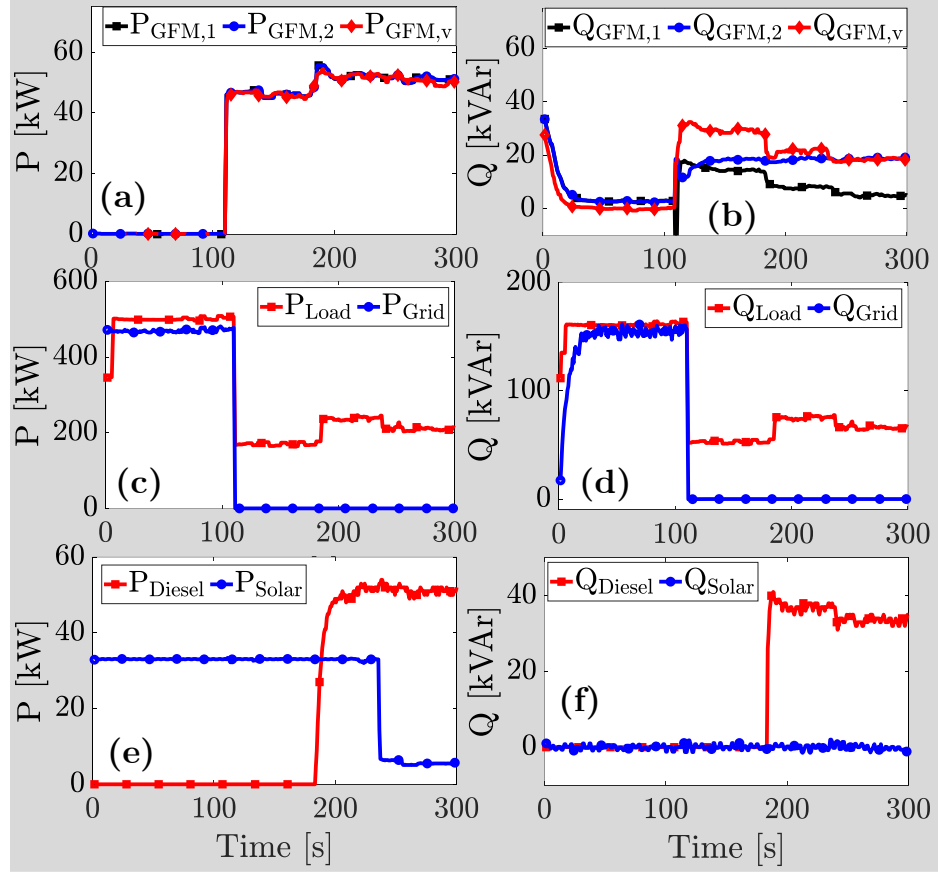


Figure 5.10: PHIL results of (a) active power output, (b) reactive power output of GFM battery inverter systems, (c) active power output, (d) reactive power output of the grid, loads, and solar inverter system, (e) active power output, (f) reactive power output of the diesel generation sets of the CI system of Fig. 5.1 during Test Case 1 of Table 5.1.

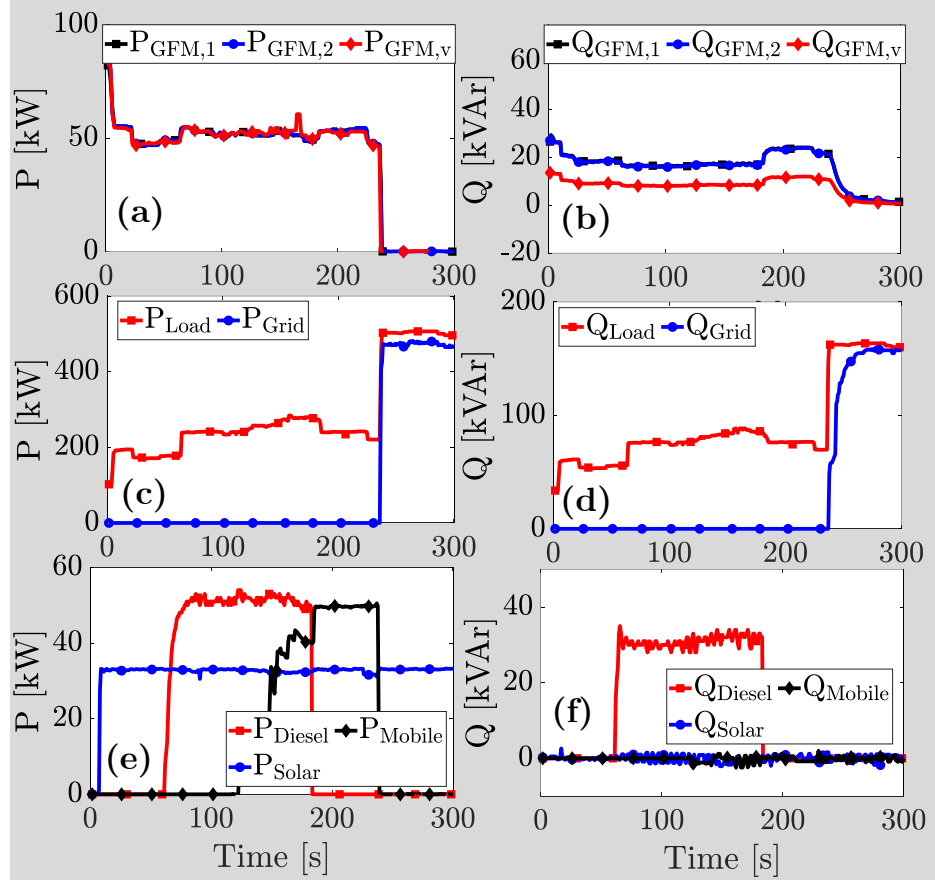


Figure 5.11: PHIL results of (a) active power output, (b) reactive power output of GFM battery inverter systems, (c) active power output, (d) reactive power output of the grid, loads, and solar inverter system, (e) active power output, (f) reactive power output of the diesel generation sets and mobile grid-following inverter system of the CI system of Fig. 5.1 during Test Case 2 of Table 5.2.

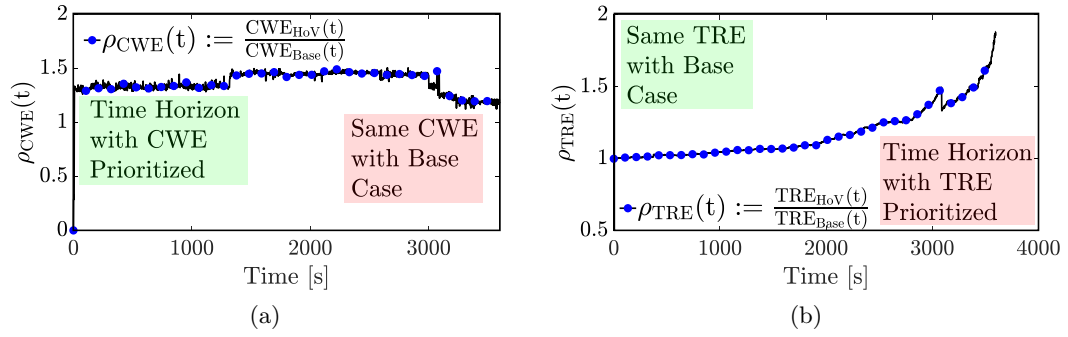


Figure 5.12: Comparative results of (a) the ratio of the CWE metric value resulting from HoV engine and base case engine, (b) ratio of the TRE metric value resulting from HoV engine and base case engine during 1 hours of PHIL experiment on the test system of Fig. 5.1.

Chapter 6

Design of a Robust Power Hardware-in-the-Loop Digital Real-time Simulation Platform

6.1 Gist of the Chapter

The power hardware-in-the-loop (PHIL) framework provides an attractive platform for evaluating new frameworks and power hardware including transformers, converters and machines. It allows for conducting diverse tests on a laboratory scale, providing results that closely resemble real-world scenarios, all while avoiding the high costs and risks associated with experimentation on a full-scale power system. However, despite its advantages of the PHIL framework, considerable challenges remain where often the performance is not robust and the fidelity to the real-world needs to be quantified and improved. This chapter addresses these challenges by synthesizing an interface between the hardware and software components of PHIL while employing a modern control perspective for managing inherent uncertainties. The chapter describes a methodology that quantifies model uncertainty caused by the point-of-connection in the emulated power network of the software system and the model uncertainty stemming from the varying nature of the hardware-under-test. To overcome these uncertainties, the chapter proposes a robust PHIL interface controller based on μ -synthesis. This controller ensures multiple objectives are met that includes robust stability, performance, accuracy, and tracking capabilities. To assess the effectiveness and viability of the proposed PHIL

interface, a PHIL experiment is conducted. The experiment involves interfacing an emulated software system based on a 225-bus, 110V, 60Hz, 1MW residential sub-network of the University of Minnesota and suburb Minneapolis. The interface is established using a real-time simulator and a physical hardware system consisting of two $1-\phi$ 1.67kVA inverter systems and one linear and nonlinear load, each with a capacity of 1.8 kVA.

6.2 State-of-the-art on PHIL Interface Scheme

Enabled by advances in computational technology, real-time simulation (RTS) provides an effective framework for emulating the behavior of power systems in real-time. RTS provide high fidelity emulation by solving dynamical equations that model the power system employing a real-world clock [165]. The impact of RTS is substantially enhanced in the power hardware-in-the-loop (PHIL) framework, where, real physical hardware components (termed hardware-under-test (HUT)) such as electrical machines, converters and various kinds of loads, can be tested where the HUT interacts with a power system emulated in RTS software. It serves as an efficient in-laboratory testing methodology for the power hardware under close-to-real conditions [166].

Developing a PHIL simulation requires special attention to be paid to the interface between the RTS and the HUT to ensure proper operation. Stability and fidelity issues that arise due to mismatch between the RTS and HUT caused, for example, from time-delays between the analog signals in HUT and their representations provided to the digital RTS, require detailed analysis followed by mitigation strategies [167]. A variety of interface algorithms (IAs) are reported; a summary is provided in [168, 169]. Common algorithms include; *i*) ideal transformer model (ITM) method [170, 171, 172], *ii*) partial circuit duplication (PCD) method [173, 174], *iii*) feedback current filtering (FCF) method [175], *iv*) damping impedance (DIM) method [176, 177], *v*) time-variant first-order approximation (TFA) method [178], and *vi*) transmission-line model (TLM) method [179]. The ITM method is most commonly used due to its superior accuracy, ease of implementation, and suitability for a large range of ratings of HUT [180, 181]. However, the original ITM method often suffers from stability issues due to varying nature of HUT, as discussed in [182]. As a result, various improvements are proposed in the ITM method that can be broadly classified as: *a*) uncompensated open-loop ITM methods [183, 184, 185, 186, 187], and *b*) compensated closed-loop ITM methods [188, 189, 190, 191, 192]. Table 6.1 provides a comparative study of these enhancements of ITM-based interface algorithms proposed since last decade. Existing ITM-based

Table 6.1: a summary of major improvements in the itm-based ia.

Ref.	Pros and Cons
[183, 184, 185, 186, 187]	• Classical uncompensated open-loop ITM interfaces are modified using various enhancements to improved stability. • E.g., Smith predictor compensator eliminating the impact of delays, open-loop voltage-source inverter to be used as power amplifier for better bandwidth and accuracy, multirate partitioning using small time step interfaces, etc. • Suffers stability with active HUT (GFL/GFM inverters). • Most are designed for passive load based HUT and emulated systems without considering uncertainties.
[188, 189]	• Feedback compensation is used in the ITM interface to stabilize PHIL test designed with GFL inverter-based HUT. • Proportional-resonant controller-based compensator is used for better voltage tracking compared to [183, 184, 185, 186, 187]. • Not agnostic to nature of HUT, PoC of the emulated network. Lacks robustness in accuracy under uncertainties.
[190, 191]	• $\mathcal{H}_\infty$-based robust feedback compensator is designed for GFL inverter-based HUT modeled as a fixed current source. • Nominally enhanced stable interface is achieved with improved accuracy, performance compared to prior works. • Lacks robustness in accuracy under uncertainties of HUT. • Robustness not guaranteed under varying emulated PoC.
[192]	• Classical proportional-resonant controller-based feedback compensation with virtual GFM inverter inside emulated system is used to stabilize PHIL test designed for a GFM inverter-based HUT with better performance and accuracy. • Suffers robustness in performance and accuracy under uncertain nature of HUT, and the varying emulated PoC.

IAs, both uncompensated open-loop and compensated closed-loop methods do not *a priori* address the uncertainty in the emulated system and in the types of HUT. To the authors' best knowledge, the state-of-the-art lacks a study of PHIL IA under the impact of uncertainties imposed by the RTS and the HUT; however the need for robust IA is evident. Moreover, even-though the concerns lend themselves to a multi-objective description, current interface design addresses these interacting objectives in a ad-hoc manner.

6.3 Problem Formulation and Motivation

A description of a closed-loop compensated ITM-based PHIL interface, is shown in Fig. 6.1 which consists of the following systems:

- 1) **Software System:** It comprises the emulated power system model terminated by the equivalent current source model of the HUT and the closed-loop interface controller.
- 2) **Hardware System:** It comprises hardware interfacing components such as analog-to-digital (A/D) converters, digital-to-analog (D/A) converters, voltage sensor, current

sensor and power amplifier, across which the HUT is terminated.

The instantaneous voltage waveform, v_o , and current waveform, i_o^* , across the power amplifier are relayed from the *Hardware System* back to the *Software System*, as v'_o , i'_o respectively, that involves the voltage and current sensors along with the A/D converters. The interface controller has v'_o , i'_o and the voltage waveform at the PoC in the emulated system (across which the HUT is considered to be connected), v_o^* , as input signals. The controller generates a voltage signal, u_o , relayed via a D/A converter to the power amplifier for generating a voltage waveform at the terminal and across the HUT. The controller also provides the current signal, i_o , as an input to the equivalent current source model of the HUT in the *Software System*. It is important to note that the additional path from v_o to v'_o realized using a voltage sensor and an A/D converter followed by the *Interface Controller* with an output u_o makes the PHIL interface to be closed-loop which is different from conventional uncompensated open-loop interfaces of [170, 171, 172, 183, 184, 185, 186, 187]. The block diagram of the uncompensated ITM-based PHIL interface is shown in Fig. 6.2. $\mathcal{G}_{AO}^v(s)$, $\mathcal{G}_{AI}^i(s)$ are the transfer functions representing the D/A converter in the voltage path and A/D converter in the current path respectively. $\mathcal{G}_{CS}(s)$ is the transfer function that models the current sensor. $\mathcal{G}_{PA}(s)$ is the power amplifier's transfer function, and $\mathcal{Z}_{PA}(s)$ is the output impedance of the power amplifier. The mathematical models and parameters used in this work are summarized in Table 6.2. $\mathcal{K}_{CS}^{\text{up}} = 1/A_{cs}^{\text{gn}}$ and $\mathcal{K}_{PA}^{\text{down}} = 1/A_{pa}^{\text{gn}}$ are the current and voltage scaling factors. $v_{Th}(s)$ and $\mathcal{Z}_{Th}(s)$ are the Thévenin equivalent voltage source, and Thévenin equivalent series impedance, respectively, that represent the equivalent model of the electrical network of the *Software System* of Fig. 6.1. $\mathcal{Y}_{HUT}(s)$ is the equivalent admittance model of the HUT in the *Hardware System* of Fig. 6.1. The loop-gain of the SISO closed-loop system of Fig. 6.2 is given by:

$$\mathcal{P}_{\text{un}}^{\text{op}}(s) = \mathcal{L}_v(s) \frac{\mathcal{Z}_{Th}(s)\mathcal{Y}_{HUT}(s)}{1 + \mathcal{Z}_{PA}(s)\mathcal{Y}_{HUT}(s)} \mathcal{L}_i(s), \quad (6.1)$$

where, $\mathcal{L}_v(s) = \mathcal{K}_{PA}^{\text{down}}\mathcal{G}_{AO}^v(s)\mathcal{G}_{PA}(s)$ and $\mathcal{L}_i(s) = \mathcal{G}_{CS}(s)\mathcal{G}_{AI}^i(s)\mathcal{K}_{CS}^{\text{up}}$. Transfer function from v_{Th} to v_o , $\mathcal{G}_{v_{Th}}^{v_o}(s)$, and v_{Th} to i_o , $\mathcal{G}_{v_{Th}}^{i_o}(s)$, are given by:

$$\mathcal{G}_{v_{Th}}^{v_o}(s) = \frac{\mathcal{L}_v(s)}{1 + \mathcal{P}_{\text{un}}^{\text{op}}(s)}, \quad \mathcal{G}_{v_{Th}}^{i_o}(s) = \frac{\mathcal{P}_{\text{un}}^{\text{op}}(s)/\mathcal{Z}_{Th}(s)}{1 + \mathcal{P}_{\text{un}}^{\text{op}}(s)} \quad (6.2)$$

respectively. The stability of the system of Fig. 6.2 can be assessed by applying Nyquist

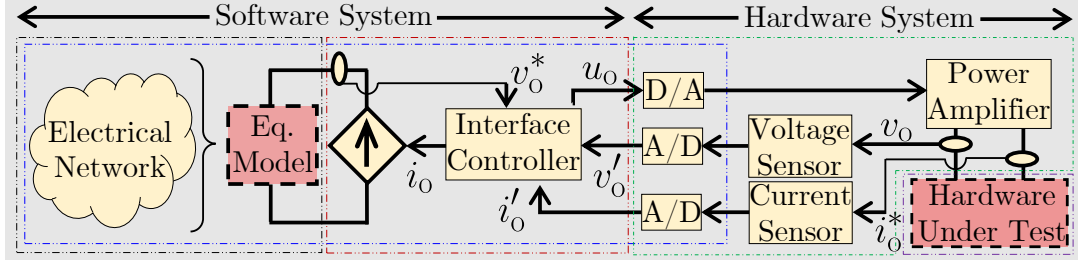


Figure 6.1: Closed-loop compensated ITM-based PHIL configuration.

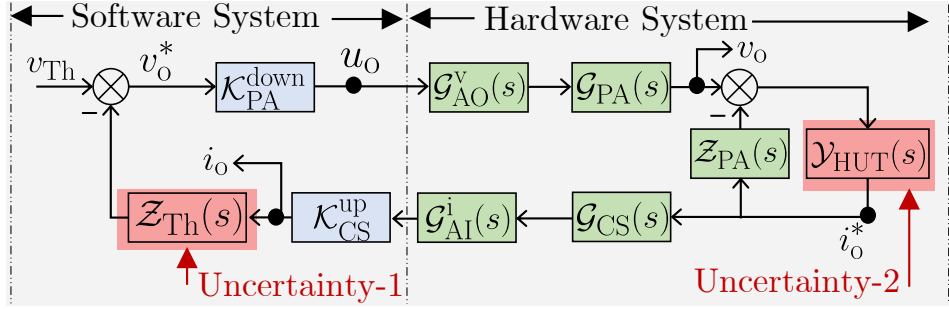


Figure 6.2: Block diagram of the uncompensated ITM-based PHIL interface.

stability criteria to the characteristic equation, given by $1 + \mathcal{P}_{un}^{op}(s) = 0$. Two sources of uncertainties are described below.

- uncertainty caused by the emulated electrical network inside the *Software System* (shown as Uncertainty-1 in Fig. 6.2). The emulated electrical network may have many possible PoC, at which the physical hardware system is connected. For each PoC, the parameters of the equivalent model, $\mathcal{Z}_{Th}(s)$, will be different depending on the electrical position and stiffness at the PoC in the emulated network.
- uncertainty caused by the HUT in the *Hardware System* (shown as Uncertainty-2 in Fig. 6.2). The HUT can either include active elements (e.g. GFL or GFM inverters) or passive elements (e.g. linear and non-linear loads) or any combination of these. Various combinations of varying nature of the HUT cause uncertainty in $\mathcal{Y}_{HUT}(s)$.

Thus the controller has to provide performance with respect to multiple objectives while being robust to uncertainty in models of both hardware and software subsystems. A case study of the performance of the system described in Fig. 6.2 with an extreme high

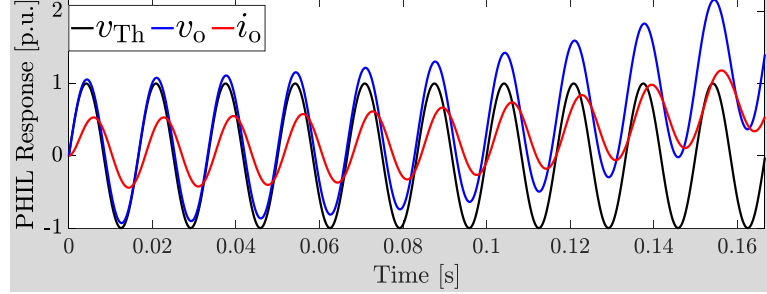


Figure 6.3: An unstable instantaneous voltage and current response of the conventional uncompensated ITM-based PHIL interface method with the extreme candidates of the uncertain $\mathcal{Z}_{Th}(s)$ and $\mathcal{Y}_{HUT}(s)$.

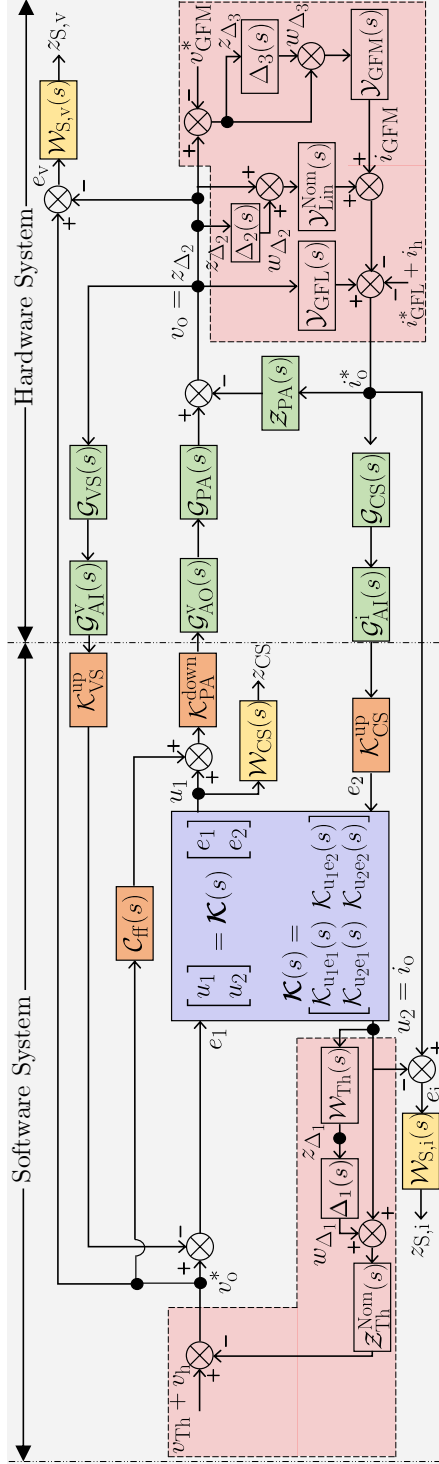
value of $\mathcal{Z}_{Th}(s)$ (signifies weak-grid at PoC) and an equivalent $\mathcal{Y}_{HUT}(s)$ (parallel combination of a GFM and a GFL inverter) is considered here. Fig. 6.3 shows that for a bounded $v_{Th}(t)$, the instantaneous voltage waveform, $v_o(t)$, and current waveform, $i_o(t)$, of the system of Fig. 6.2 collapse after few cycles. Indeed, it is observed that for the selected $\mathcal{Z}_{Th}(s)$ and $\mathcal{Y}_{HUT}(s)$, the system is unstable which can also be assessed based on Nyquist stability criteria using $1 + \mathcal{P}_{un}^{op}(s) = 0$. Thus, these uncertainties pose significant challenges and if unaddressed will result in unacceptable performance and can impact the stability of the system [193, 194, 195, 196]. This chapter provides a methodology for designing a *robust-enough* and *stable* PHIL interface controller based on μ -synthesis which meets the satisfactory performance criteria under *worst-case* admissible uncertainties while addressing multiple objectives.

6.4 μ -synthesis-based Control Framework

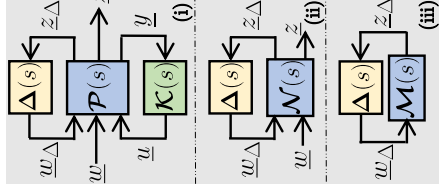
Motivated by Section 6.3, this section will describe the control framework and the required modeling for μ -synthesis-based robust PHIL interface controller synthesis.

Table 6.2: Models and parameters of the phil interface.

Model	Parameters	Values
$\mathcal{G}_{\text{PA}} = \frac{A_{\text{pa}}^{\text{gn}}}{1 + sT_{\text{pa}}^{\text{bw}}} e^{-sT_{\text{pa}}^{\text{dl}}}$	$A_{\text{pa}}^{\text{gn}}$: Amplifier gain $T_{\text{pa}}^{\text{bw}}$: Time constant $T_{\text{pa}}^{\text{dl}}$: Amplifier I/O delay	$30\sqrt{2}$ 10^{-3}s $50 \times 10^{-6}\text{s}$
$\mathcal{Z}_{\text{PA}} = R_{\text{pa}} + sL_{\text{pa}}$	R_{pa} : Amplifier resistance L_{pa} : Amplifier inductance	0.5Ω 1 mH
$\mathcal{G}_{\text{AI}}^{\text{v}} = \frac{A_{\text{ai}}^{\text{gn}}(1 - e^{-sT_{\text{ai}}^{\text{sm}}})}{sT_{\text{ai}}^{\text{sm}}(1 + sT_{\text{pa}}^{\text{bw}})}$ $= \mathcal{G}_{\text{AI}}^{\text{i}}$	$A_{\text{ai}}^{\text{gn}}$: Filter gain $T_{\text{ai}}^{\text{sm}}$: Sample time $T_{\text{ai}}^{\text{bw}}$: Filter cutoff	1 $2.5 \times 10^{-6}\text{s}$ $16 \times 10^{-6}\text{s}$
$\mathcal{G}_{\text{AO}}^{\text{v}} = e^{-sT_{\text{ao}}^{\text{cn}}}$	$T_{\text{ao}}^{\text{cn}}$: Conversion time	10^{-6}s
$\mathcal{G}_{\text{VS}} = \frac{A_{\text{vs}}^{\text{gn}}}{1 + sT_{\text{vs}}^{\text{bw}}} e^{-sT_{\text{vs}}^{\text{dl}}}$	$A_{\text{vs}}^{\text{gn}}$: Voltage sensor gain $T_{\text{vs}}^{\text{bw}}$: Sensor time constant $T_{\text{vs}}^{\text{dl}}$: Sensor I/O delay	10^{-3} $40 \times 10^{-9}\text{s}$ $20 \times 10^{-9}\text{s}$
$\mathcal{G}_{\text{CS}} = \frac{A_{\text{cs}}^{\text{gn}}}{1 + sT_{\text{cs}}^{\text{bw}}} e^{-sT_{\text{cs}}^{\text{dl}}}$	$A_{\text{cs}}^{\text{gn}}$: Current sensor gain $T_{\text{cs}}^{\text{bw}}$: Sensor time constant $T_{\text{cs}}^{\text{dl}}$: Sensor I/O delay	10^{-3} $100 \times 10^{-6}\text{s}$ $175 \times 10^{-6}\text{s}$



(a)



(b)

Figure 6.4: Figure of (a) the block diagram of the proposed robust PHIL interface, (b) Δ - $\mathcal{P}$ - $\mathcal{K}$, Δ - $\mathcal{N}$ and Δ - $\mathcal{M}$ representation of the structure of Fig. 6.4(a)

6.4.1 Developing the Control Framework

A block diagram representation of the entire ITM interface-based PHIL configuration of Fig. 6.1 is shown in Fig. 6.4(a). The proposed robust PHIL interface controller, $\mathcal{K}(s)$, is shown inside the blue box. The uncertain equivalent model of the electrical network in the *Software System*, and the uncertain model of the generic HUT in the *Hardware System* are shown inside the red boxes. The shaping transfer functions for designing the robust controller are shown in yellow boxes. These components are all discussed in later sections. Rest of the components that includes D/A and A/D converters, voltage and current sensors and the power amplifier of Fig. 6.1 are modelled judiciously in this section. $\mathcal{C}_{\text{ff}}(s) = 1$ is the feedforward gain for the conventional uncompensated ITM interface, $\mathcal{G}_{\text{AI}}^{\text{v}}(s)$, is the transfer functions representing the D/A converter in voltage path. $\mathcal{G}_{\text{VS}}(s)$ is the transfer function that models the voltage sensor. Rest of the modelings was described earlier in Section. 6.3. The mathematical models and parameters used in this work are summarized in Table 6.2. $\mathcal{K}_{\text{VS}}^{\text{up}} = 1/A_{\text{VS}}^{\text{gn}}$ is the voltage scaling factors. A second-order Padé approximation is used to model the exponential time delay (e^{-sT}) [197].

Modeling of Uncertainty

A systematic approach is followed here for characterizing and modeling the uncertainty for both *Software* and *Hardware System* as discussed below:

a) Uncertainty of Software System: The equivalent model of the emulated electrical network at a particular PoC is designed based on Thévenin equivalent circuit representation with voltage harmonic distortion as follows:

$$v_o^*(s) = v_{\text{Th}}(s) - \mathcal{Z}_{\text{Th}}(s)i_o(s) + v_h(s), \quad (6.3)$$

where, $v_{\text{Th}}(s)$, $\mathcal{Z}_{\text{Th}}(s)$ are the Thévenin equivalent voltage source, and Thévenin equivalent series impedance respectively. In accordance with IEEE Std. 519 [73], the harmonics component, $v_h(s)$, is defined as $v_h(s) := \sum_d v_d(s)$. where $d \in \{3 \ 5 \ 7 \ 9\}$. Equivalent Thévenin impedance, $\mathcal{Z}_{\text{Th}}(s)$ varies due to ever-changing generation-demand situation [198], change in topology due to network contingencies [199], and variation in system frequency [200]. These variations are captured by modeling the $\mathcal{Z}_{\text{Th}}(s)$ as a multiplicative

uncertainty of the following form:

$$\mathcal{Z}_{\text{Th}}(s) := [1 + \mathcal{W}_{\text{Th}}(s)\Delta_1(s)]\mathcal{Z}_{\text{Th}}^{\text{Nom}}(s), \quad (6.4)$$

where, $\mathcal{Z}_{\text{Th}}^{\text{Nom}}(s)$ is the nominal Thévenin equivalent impedance, and $\Delta_1(s)$ is a normalized dynamic Linear Time-invariant (LTI) system such that $\|\Delta_1(s)\|_\infty \leq 1$. $\mathcal{W}_{\text{Th}}(s)$ is a stable, proper weighting function selected such that a large spectrum of variation of $\mathcal{Z}_{\text{Th}}(s)$ can be captured. The short-circuit ratio (SCR) is used to characterize the grid stiffness and is employed as the nominal Thévenin equivalent impedance at the PoC, $\mathcal{Z}_{\text{Th}}^{\text{Nom}}(s)$. SCR is defined as $(V_{\text{PoC}}^{\text{Nom}})^2 / [S_{\text{B}}|\mathcal{Z}_{\text{Th}}^{\text{Nom}}(j\omega_{\text{Nom}})|]$, where $V_{\text{PoC}}^{\text{Nom}}$ and ω_{Nom} are the nominal voltage and frequency at PoC, and S_{B} is the rated apparent power of HUT at PoC. Usually the grid at PoC is considered as weak when the SCR is less than 3 [71]. In this work, with a pre-specified SCR (< 3) and $\omega_{\text{Nom}}\mathcal{L}_{\text{Th}}^{\text{Nom}}/\mathcal{R}_{\text{Th}}^{\text{Nom}}$ ratio (< 10), $\mathcal{Z}_{\text{Th}}^{\text{Nom}}(s) := \mathcal{R}_{\text{Th}}^{\text{Nom}} + s\mathcal{L}_{\text{Th}}^{\text{Nom}}$ is determined.

Remark 4. *Note that a wide range of conditions including a stiff ($\mathcal{W}_{\text{Th}}(s) = 1$ and $\Delta_1(s) = -1$) and extremely weak ($\mathcal{W}_{\text{Th}}(s) > 0$ and $\Delta_1(s) = 1$) PoC are accommodated with this uncertainty modeling. $v_{\text{h}}(s)$ entering the system as an exogenous disturbance and provides a means to accommodate PoC voltages that are near sinusoidal to signals that are heavily distorted by higher harmonics..*

b) Uncertainty of Hardware System: Characterizing the nature of the HUT is essential in the process of designing the interface controller. In this work, the HUT is modeled by a parallel combination of four components. First component is the linear passive load which is modeled as an admittance, $\mathcal{Y}_{\text{Lin}}(s)$. To capture the variation of the loading, the following multiplicative uncertainty is defined:

$$\mathcal{Y}_{\text{Lin}}(s) := [1 + \Delta_2(s)]\mathcal{Y}_{\text{Lin}}^{\text{Nom}}(s), \quad (6.5)$$

where, $\mathcal{Y}_{\text{Lin}}^{\text{Nom}}(s) := 1/[\mathcal{R}_{\text{Lin}}^{\text{Nom}} + s\mathcal{L}_{\text{Lin}}^{\text{Nom}}]$. Considering 50% rated VA of the linear load as $S_{\text{rat.}}^{\text{Lin}}$, with active power, $P_{\text{rat.}}^{\text{Lin}}$, and reactive power, $Q_{\text{rat.}}^{\text{Lin}}$, the following holds:

$$R_{\text{Lin}}^{\text{Nom}} = (V_{\text{PoC}}^{\text{Nom}})^2 \frac{P_{\text{rat.}}^{\text{Lin}}}{(S_{\text{rat.}}^{\text{Lin}})^2}, L_{\text{Lin}}^{\text{Nom}} = (V_{\text{PoC}}^{\text{Nom}})^2 \frac{Q_{\text{rat.}}^{\text{Lin}}}{(S_{\text{rat.}}^{\text{Lin}})^2} \frac{1}{\omega_{\text{Nom}}}.$$

$\Delta_2(s)$ is a normalized dynamic LTI uncertainty such that $\|\Delta_2(s)\|_\infty \leq 1$, where, $\|\cdot\|_\infty$ refers to the $\mathcal{H}_\infty$ norm. In accordance with IEEE Std. 519 [73], the second component is the non-linear harmonic load, $i_{\text{h}}(s)$, is defined as $i_{\text{h}}(s) := \sum_{\text{d}} i_{\text{d}}(s)$. where $\text{d} \in$

{3 5 7 9}. The third and forth component includes active source-based HUT i.e. GFL and GFM inverter system respectively. The GFL inverter system is represented by the Norton equivalent model as $i_{\text{GFL}} = i_{\text{GFL}}^* - \mathcal{Y}_{\text{GFL}}(s)v_o$. Usually $\mathcal{Y}_{\text{GFL}}(s)$ includes the admittance of the filter capacitor of the GFL inverter [201]. The GFM inverter system is represented by the Thévenin equivalent model as $i_{\text{GFM}} = Y_{\text{GFM}}(s)[v_o - v_{\text{GFM}}^*]$, where $Y_{\text{GFM}}(s)$ is modeled by a multiplicative uncertainty as $Y_{\text{GFM}}(s) := [1 + \Delta_3(s)]\mathcal{Y}_{\text{GFM}}(s)$ to capture the varying power transfer by the GFM inverter. Usually $\mathcal{Y}_{\text{GFM}}(s)$ includes the admittance of the grid-side filter inductance with parasitic resistance of the GFM inverter [202]. $\Delta_3(s)$ is a normalized dynamic LTI uncertainty such that $\|\Delta_3(s)\|_\infty \leq 1$. As a result, the generic uncertainty model of the HUT is given by:

$$i_o^*(s) = \mathcal{Y}_{\text{Lin}}(s)v_o + i_h - i_{\text{GFL}} - i_{\text{GFM}}, \quad (6.6)$$

Remark 5. *It is observed that no-load ($\Delta_2(s) = -1$) to full rated ($\Delta_2(s) = 1$) linear load, and no GFM connection ($\Delta_3(s) = -1$) to GFM connection with non-zero power transfer ($\Delta_3(s) \neq -1$) are accommodated with this uncertainty modeling. $i_h(s)$ and $i_{\text{GFL}}(s)$ work as exogenous disturbances and $= 0 / \neq 0$ capture the absence/presence of non-linear load and GFL inverter system in the HUT.*

Shaping of Transfer Functions

The closed-loop objectives in designing the PHIL interface controller ($\mathcal{K}_{u_1 e_1}(s)$, $\mathcal{K}_{u_1 e_2}(s)$, $\mathcal{K}_{u_2 e_1}(s)$, $\mathcal{K}_{u_2 e_2}(s)$ of Fig. 6.4(a)) are as follows:

- i) Reference Tracking:** v_o and i_o track v_o^* and i_o^* with minimum tracking error respectively.
- ii) Disturbance Rejection:** Effect of i_o^* on v_o and effect of v_o^* on i_o are attenuated.
- iii) Control Effort Reduction:** u_1 satisfies the bandwidth limitations of the power amplifier.

Based on the objectives, user-defined weighting transfer functions with design guidelines are provided below.

a) Selection of $\mathcal{W}_{S,v}(s)$ and $\mathcal{W}_{S,i}(s)$: To shape the sensitivity transfer functions of voltage and current loop, the weighting function, $\mathcal{W}_{S,v}(s)$ and $\mathcal{W}_{S,i}(s)$ are introduced so that the tracking errors, $e_v := v_o^* - v_o$ and $e_i := i_o^* - i_o$ at fundamental and selected harmonic frequency are small. Both are modeled to have peaks around fundamental

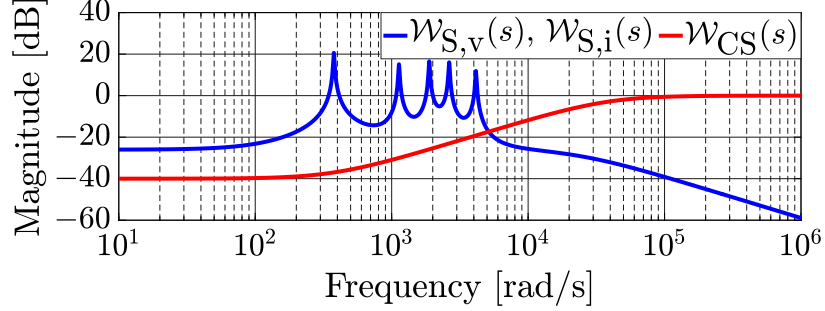


Figure 6.5: Bode plots of magnitudes of weighting transfer functions.

and harmonic frequencies with 2nd order roll-off, $k_S(s)$ and formed as:

$$\mathcal{W}_{S,v}(s) = \mathcal{W}_{S,i}(s) = k_S(s) \prod_{h=1,3,5,7,9} \frac{s^2 + 2k_{s,h}\zeta h\omega_N s + h^2\omega_N^2}{s^2 + 2\zeta h\omega_N s + h^2\omega_N^2},$$

where, $k_{s,h}$ is selected to exhibit peaks at h^{th} harmonic and ζ addresses the off-nominal frequency around the nominal.

b) Selection of $\mathcal{W}_{CS}(s)$: It is designed to reduce high-frequency control effort to shape the performance of u_1 for the power amplifier. Hence, it is designed as a high-pass filter with cut-off frequency, governed by the power-amplifier's bandwidth., with the form:

$$\mathcal{W}_{CS}(s) = k_{CS} \frac{s + k_{CS,1}\omega_N}{s + k_{CS,2}\omega_N}, \text{ where } k_{CS,1} \ll k_{CS,2}.$$

Representative of weighting functions are shown in Fig. 6.5.

Preparing the Generalized Plant

Toward designing a robust PHIL controller, the multi-loop closed-loop block diagram in Fig. 6.4(a) is consolidated into the general control configuration in Fig. 6.4(b)(i) [72]. Here, $\mathcal{P}(s)$, $\mathcal{K}(s)$ and $\Delta(s) := \text{diag}(\Delta_1, \Delta_2, \Delta_3)$ are the generalized MIMO plant model, controller model and the structured uncertainty respectively. Moreover, $\underline{\mathbf{w}} := \begin{bmatrix} v_{Th} + v_h & v_{GFM}^* & i_{GFL}^* + i_h \end{bmatrix}^\top$, $\underline{\mathbf{w}}_\Delta := \begin{bmatrix} w_{\Delta_1} & w_{\Delta_2} & w_{\Delta_3} \end{bmatrix}^\top$, $\underline{\mathbf{z}} := \begin{bmatrix} z_{S,v} & z_{S,i} & z_{CS} \end{bmatrix}^\top$, $\underline{\mathbf{z}}_\Delta := \begin{bmatrix} z_{\Delta_1} & z_{\Delta_2} & z_{\Delta_3} \end{bmatrix}^\top$, $\underline{\mathbf{u}} := \begin{bmatrix} u_1 & u_2 \end{bmatrix}^\top$ and $\underline{\mathbf{y}} := \begin{bmatrix} y_1 & y_2 \end{bmatrix}^\top$. The generalized MIMO

plant maps as follows:

$$\begin{bmatrix} \underline{z}_\Delta \\ \underline{z} \\ \underline{y} \end{bmatrix} = \begin{bmatrix} \mathcal{P}_{\Delta\Delta}(s) & \mathcal{P}_{\Delta w}(s) & \mathcal{P}_{\Delta u}(s) \\ \mathcal{P}_{z\Delta}(s) & \mathcal{P}_{zw}(s) & \mathcal{P}_{zu}(s) \\ \mathcal{P}_{y\Delta}(s) & \mathcal{P}_{yw}(s) & \mathcal{P}_{yu}(s) \end{bmatrix} \begin{bmatrix} \underline{w}_\Delta \\ \underline{w} \\ \underline{u} \end{bmatrix}. \quad (6.7)$$

The detailed MIMO transfer function models of (6.7) are not provided due to limited space. $\mathcal{P}(s)$ and $\mathcal{K}(s)$ can be combined together by a lower LFT [72] to form $\mathcal{N}(s)$ in Fig. 6.4(b)(ii) as follows:

$$\begin{aligned} \mathcal{N} &= \begin{bmatrix} \mathcal{P}_{\Delta\Delta} & \mathcal{P}_{\Delta w} \\ \mathcal{P}_{z\Delta} & \mathcal{P}_{zw} \end{bmatrix} + \begin{bmatrix} \mathcal{P}_{\Delta u} \\ \mathcal{P}_{zu} \end{bmatrix} \mathcal{K} [\mathbf{I} - \mathcal{P}_{yu} \mathcal{K}]^{-1} \begin{bmatrix} \mathcal{P}_{y\Delta} \\ \mathcal{P}_{yw} \end{bmatrix}^\top \\ &= \begin{bmatrix} \mathcal{N}_{\Delta\Delta} & \mathcal{N}_{\Delta w} \\ \mathcal{N}_{z\Delta} & \mathcal{N}_{zw} \end{bmatrix}, \text{ where } \begin{bmatrix} \underline{z}_\Delta \\ \underline{z} \end{bmatrix} = \mathcal{N} \begin{bmatrix} \underline{w}_\Delta \\ \underline{w} \end{bmatrix}. \end{aligned} \quad (6.8)$$

All variables are functions of the Laplace variable, s ; not explicitly shown for notational convenience. The uncertainty closed-loop transfer function from $\underline{w}$ to $\underline{z}$, $\underline{z} = \mathcal{F}\underline{w}$, is related to $\mathcal{N}$ and Δ by an upper LFT where $\mathcal{F} = \mathcal{N}_{zw} + \mathcal{N}_{z\Delta}\Delta(\mathbf{I} - \mathcal{N}_{\Delta\Delta}\Delta)^{-1}\mathcal{N}_{\Delta w}$. To analyze robust stability of $\mathcal{F}$, the system can be re-arrange into $\mathcal{M}$ - Δ structure of Fig. 6.4(b)(iii) where $\mathcal{M} = \mathcal{N}_{\Delta\Delta}$.

6.5 Controller Synthesis and Stability Analysis

With reference to the configuration of Fig. 6.4(b), the μ -synthesis-based robust control problem is to determine a stabilizing controller $\mathcal{K}(s)$ that solves

$$\min_{\mathcal{K}(s) \text{ stabilizing}} \|\mathcal{N}\|_\infty. \quad (6.9)$$

This problem is solved using the MATLAB Robust Control Toolbox [203]. An algorithm for solving (6.9) along with the theoretical underpinnings of this optimization problem can be found in [72]. Upon finding a stabilizing controller, the requirement of stability and performance of the closed-loop system are checked and can be summarized

as follows:

$$\text{Nominal Stable(NS)} : \mathcal{N} \text{ is internally stable,} \quad (6.10)$$

$$\text{Nominal Performance(NP)} : \|\mathcal{N}_{\mathbf{zw}}\|_{\infty} < 1 \text{ \& NS,} \quad (6.11)$$

$$\text{Robust Stable(RS)} : \mu_{\Delta}(\mathcal{M}) < 1, \forall \omega, \text{ \& NS,} \quad (6.12)$$

$$\text{Robust Performance(RP)} : \mu_{\bar{\Delta}}(\mathcal{N}) < 1, \forall \omega, \text{ \& NS,} \quad (6.13)$$

where, $\mu_{\Delta}(\mathcal{M})$ and $\mu_{\bar{\Delta}}(\mathcal{N})$ are the structured singular values of $\mathcal{M}$ and $\mathcal{N}$ for the allowed structure of Δ and $\bar{\Delta} := \text{diag}(\Delta, \Delta_P)$ respectively with Δ_P being an unstructured uncertainty [72]. In this work, the following iterative approach is followed for the μ -synthesis problem:

Step-1. Based on the design guidelines, determine the performance weights, i.e. $\mathcal{W}_{S,v}(s)$, $\mathcal{W}_{S,i}(s)$, $\mathcal{W}_{CS}(s)$,

Step-2. Construct the generalized plant, $\mathcal{P}(s)$,

Step-3. Solve equation (6.9) with the following iterations:

i) **Synthesis-stage:** Synthesize an $\mathcal{H}_{\infty}$ controller for $\mathcal{P}(s)$, with initial choice of parameters as guided by Step 1-2.

ii) **Checking-stage:** Check if the conditions of (6.10)-(6.13) are satisfied or not. If conditions are satisfied, μ -synthesized optimal controller is achieved as the problem in (6.9) is solved. If the conditions are not satisfied, return to Step-1 to lower the specifications on performance imposed via the filters $\mathcal{W}_{S,v}(s)$, $\mathcal{W}_{S,i}(s)$, $\mathcal{W}_{CS}(s)$ and continue until conditions of (6.10)-(6.13) are all satisfied. The optimal controller $\mathcal{K}(s)$ will have an order similar to the order of $\mathcal{P}(s)$. Thus, before implementation in actual RT simulator, model order reduction is used to obtain a lower order controller using MATLAB's *balred* command. Moreover, bilinear transformation is used in the discretization stage of the resulting controller. As a result, an 18th order MIMO $\mathcal{K}(s)$ is found to perform well. The closed-loop stability and desired robust performances under uncertainties are met by obtaining NS = True, NP = 0.48, RS = 0.93 and RP = 0.96.

6.6 Experimental Results

6.6.1 Experimental Configuration

The PHIL experiment setup, as shown in Fig. 6.6 and Fig. 6.7, consists of the following:

i] Software System: A 225 bus, 110V, 60Hz, 1MW residential sub-network of University of Minnesota and suburb Minneapolis is emulated using multi-timescale real-time simulation method, with time-step of $T_s = 50\mu\text{s}$ and $T_{ss} = 5\text{ms}$ via eMEGASIM and ePHASORSIM platform respectively using the OP5700 RT-simulator (RTS) manufactured by OPAL-RT. The test system is modified to make it compatible with the laboratory capability while respecting the recommended limits of harmonic distortions as mentioned in IEEE Std. 519[73].

ii] Software Interface: It consists of the discrete-time proposed robust PHIL interfacing compensators.

iii] Hardware Interface: It consists of two types of hardware components; 1) one OP5330-based and two OP5340-based analog-out and analog-in port respectively, both realized using OP5700 RTS, 2) one power amplifier realized by 4kVA Chroma 61605 programmable ac source, and one voltage and current sensor realized by BK Precision PR-60 active differential probe and AEMC MN261 current probe respectively.

iv] Hardware System: It consists of the generic HUT under study realized by two 1- ϕ , 110V, 60Hz, 1.67kVA voltage source inverter systems (one operating in GFL and other in GFM mode). These inverter systems are controlled by Texas-Instruments TMS320F28379D, 16/12-bit floating-point 200 MHz Delfino micro-controller boards and interfaced with MAGNA-POWER programmable DC power. Linear and non-linear load are realized using 1.8kVA Sorensen SLH-500-6-1800 and 1.8kVA Chroma 63800 electronic load bank respectively.

6.6.2 Metrics for Performance Evaluation

In order to evaluate the performance of the proposed PHIL interface method, the following accuracy measures are defined: (1) instantaneous voltage tracking error, $e_v^{\text{inst}}(t) := v_o^*(t) - v_o(t)$; (2) fundamental voltage RMS error, $e_{\text{Fund}}^{\text{vRMS}}[\%] := 100 \times [V_{o,1}^* - V_{o,1}]/V_{o,1}^*$; (3) fundamental voltage phase error, $e_{\text{Fund}}^{\text{vPhase}} := \phi_1^{v_o^*} - \phi_1^{v_o}$; (4) instantaneous current tracking error, $e_i^{\text{inst}}(t) := i_o^*(t) - i_o(t)$; (5) fundamental current RMS error, $e_{\text{Fund}}^{\text{iRMS}}[\%] := 100 \times [I_{o,1}^* - I_{o,1}]/I_{o,1}^*$; (6) fundamental current phase error, $e_{\text{Fund}}^{\text{iPhase}} := \phi_1^{i_o^*} - \phi_1^{i_o}$; (7) total harmonic distortion error, $e_{\text{THD}}^{\text{V}}[\%] := 100 \times [\text{THD}_{v_o^*} - \text{THD}_{v_o}]/\text{THD}_{v_o^*}$; (8)

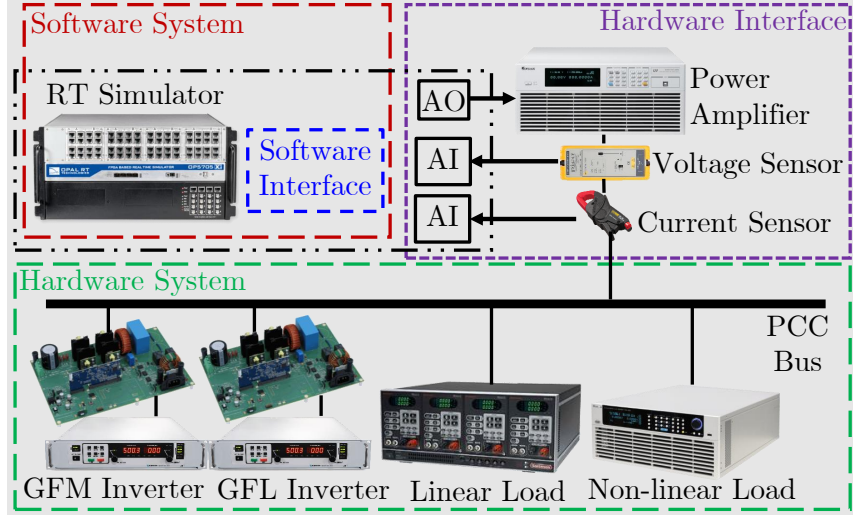


Figure 6.6: Figure of the experimental hardware configuration.

Table 6.3: Performance of the proposed phil interface

Error	Error Metrics of V			Error Metrics of I		
	$e_{\text{Fund}}^{\text{VRMS}}$	$e_{\text{Fund}}^{\text{VPhase}}$	$e_{\text{THD}}^{\text{V}}$	$e_{\text{Fund}}^{\text{iMag}}$	$e_{\text{Fund}}^{\text{iPhase}}$	$e_{\text{TDD}}^{\text{I}}$
Max	1.85%	0.08°	2.77%	1.29%	0.09°	3.59%
Min	0.19%	0.02°	1.09%	0.48%	0.04°	1.24%
Mean	0.55%	0.04°	1.23%	0.88%	0.06°	2.25%

total demand distortion error, $e_{\text{TDD}}^{\text{I}}[\%] := 100 \times [\text{TDD}_{\text{i}_o^*} - \text{TDD}_{\text{i}_o}]/\text{TDD}_{\text{i}_o^*}$; (9) active power tracking error, $e_{\text{P}}[\%] := 100 \times [P_{\text{HUT}} - P_{\text{EMU}}]/P_{\text{HUT}}$; and (10) reactive power tracking error, $e_{\text{Q}}[\%] := 100 \times [Q_{\text{HUT}} - Q_{\text{EMU}}]/Q_{\text{HUT}}$; where $V_{o,1}^*$, $I_{o,1}^*$, $V_{o,1}$, $I_{o,1}$, ϕ_1^{v*} , ϕ_1^{i*} , $\phi_1^{v_o}$, $\phi_1^{i_o}$, $\text{THD}_{v_o^*}$, THD_{v_o} , $\text{TDD}_{i_o^*}$, TDD_{i_o} are the RMS values and phase angles of the fundamental component and total harmonic distortion (THD) and total demand distortion (TDD) of the voltage and current reference and output signals of Fig. 6.4(a). Moreover, active and reactive power of the HUT (i.e. P_{HUT} , Q_{HUT}) are determined based on power calculations using $v_o(t)$ and $i_o^*(t)$, whereas active and reactive power of the emulated system (i.e. P_{EMU} , Q_{EMU}) are determined based on power calculations using $v_o^*(t)$ and $i_o(t)$.

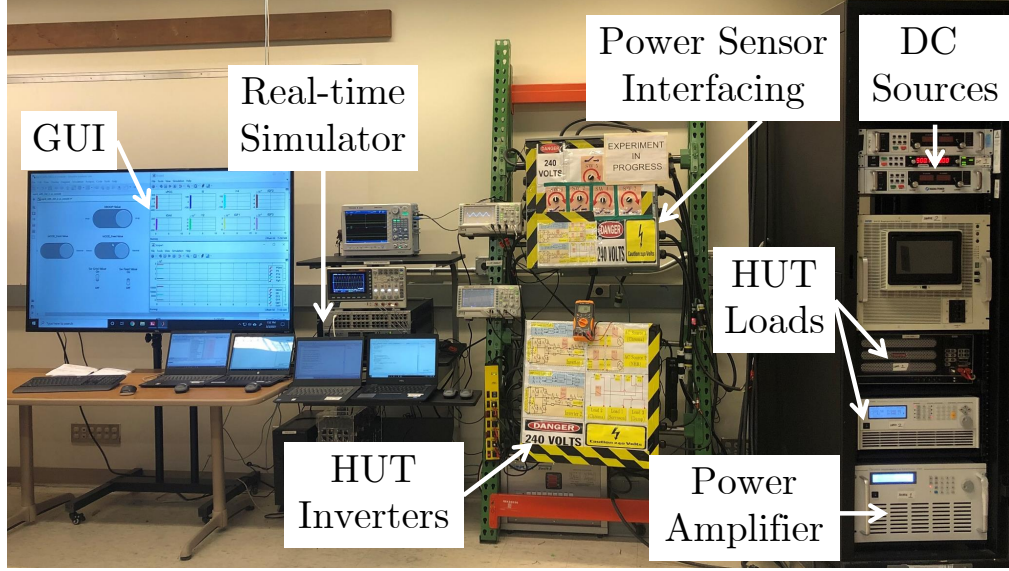


Figure 6.7: Figure of the laboratory-based experimental hardware setup.

6.6.3 Experimental Results and Discussions

Fig. 6.8 and Fig. 6.9 show the performance of the instantaneous voltage and current tracking capability of the proposed robust PHIL interface controller and provide a comparison with the conventional uncompensated ITM-based PHIL interface method [172]. Here, a combined HUT of $P = 1.96\text{kW}$ and $Q = 0.35\text{kVAr}$ containing GFM and GFL inverters and linear and non-linear loads are used. It is observed that after $t = 1\text{ s}$, while operating with the proposed robust PHIL interface, e_v^{inst} is within $\pm 2.4\text{V}$ and e_i^{inst} is within $\pm 0.25\text{A}$ which are significantly improved from the case before $t = 1\text{ s}$, where e_v^{inst} is within $\pm 35.91\text{V}$ and e_i^{inst} is within $\pm 7.94\text{A}$, where the conventional uncompensated ITM-based PHIL interface is employed. Fig. 6.10 shows the improvement of the active and reactive power tracking capability of the proposed PHIL interface method ($e_P = -0.47\%$ and $e_Q = -0.53\%$) over the uncompensated ITM-based PHIL interface ($e_P = 16.41\%$ and $e_Q = -13.2\%$) during the same case study. Over 100 PHIL simulation runs are conducted to showcase the robust performance of the proposed PHIL interface under varying HUT with various points of connection in the emulated system. Various loading of the power amplifier (0%-100%) caused by various combinations of the HUT are realized by varying connections of GFL and GFM inverters along with linear and non-linear loads with varying power set-points. Moreover, multiple point of connection

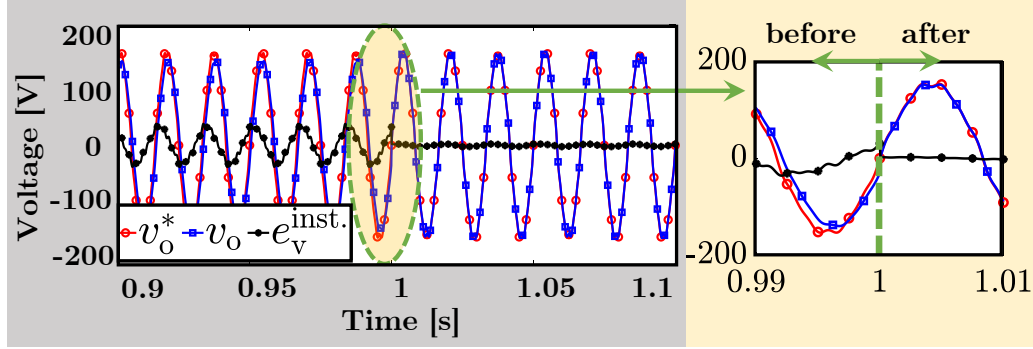


Figure 6.8: Instantaneous voltage tracking capability after switched to the proposed robust PHIL interface method from conventional uncompensated ITM-based PHIL interface method.

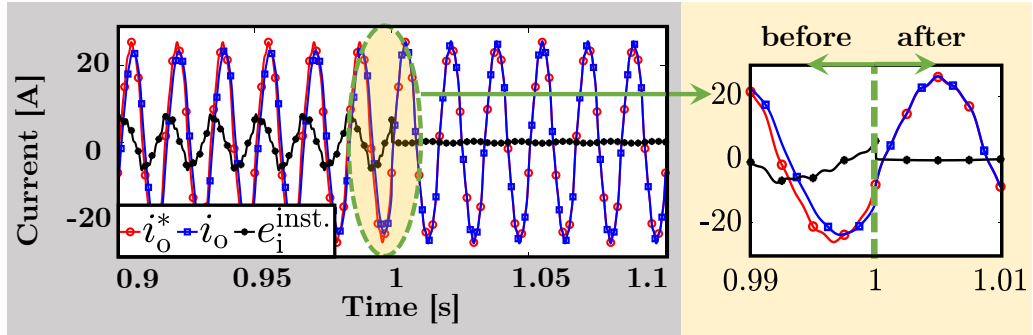


Figure 6.9: Instantaneous current tracking capability after switched to the proposed robust PHIL interface method from conventional uncompensated ITM-based PHIL interface method.

in the emulated system are realized with varying equivalent Thevenin Impedance at PoC, $Z_{Th} \in [0 \ 5 + 5j]$ PU. Table 6.3 shows the values of the metrics, as defined in Section 6.6.2, achieved by the proposed robust PHIL interface method. The following metrics are satisfied, $e_{Fund}^{vRMS} \in [0.19 \ 1.85]\%$ with a mean of 0.55%, $e_{Fund}^{vPhase} \in [0.02^\circ \ 0.04^\circ]$ with a mean of 0.04°, $e_{THD}^v \in [1.09 \ 2.77]\%$ with a mean of 1.23%, $e_{Fund}^{iRMS} \in [0.48 \ 1.29]\%$ with a mean of 0.48%, $e_{Fund}^{iPhase} \in [0.04^\circ \ 0.09^\circ]$ with a mean of 0.06°, $e_{TDD}^I \in [1.24 \ 3.59]\%$ with a mean of 2.25%. Moreover, the Whisker plots of Fig. 6.11 show the distributions of e_P and e_Q under varying equivalent Thevenin impedance of the PoC in the emulated system under four different types of HUT; Type 1: linear and non-linear load without inverter systems, Type 2: linear and non-linear load with GFL inverter, Type 3: linear

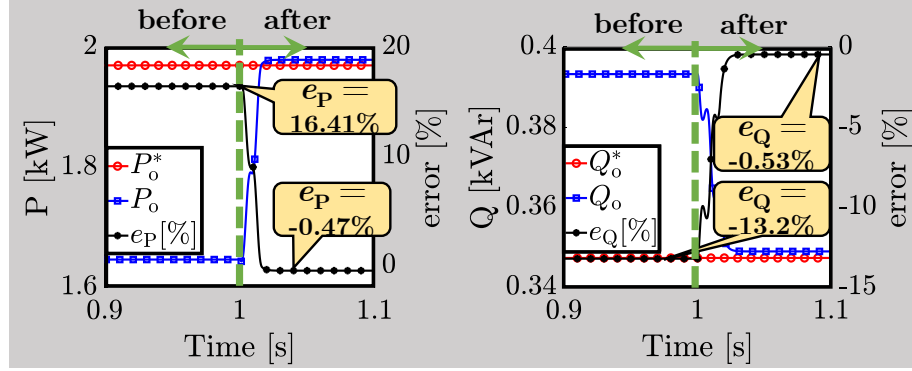


Figure 6.10: Active and reactive power tracking capability after switched to the proposed robust PHIL interface method from conventional uncompensated ITM-based PHIL interface method.

and non-linear load with GFM inverter, and Type 4: linear and non-linear load with both GFM and GFL inverter. Fig. 6.11(a), Fig. 6.11(b) for Type 1 HUT, Fig. 6.11(c), Fig. 6.11(d) for Type 2 HUT, Fig. 6.11(e), Fig. 6.11(f) for Type 3 HUT, Fig. 6.11(g), Fig. 6.11(h) for Type 4 HUT show the e_P and e_Q respectively and it is observed that the corresponding tracking errors are consistently within $\pm 2\%$ in all cases. These results corroborate that the proposed PHIL interface method exhibits significantly improved and robust performance under varying condition of HUT and PoC in the emulated system under study. A comparison study with two existing compensated ITM-based PHIL interface methods of [189, 191] is also conducted to showcase the superiority of the proposed robust PHIL interface method. However, the results are not added in this version due to limited space.

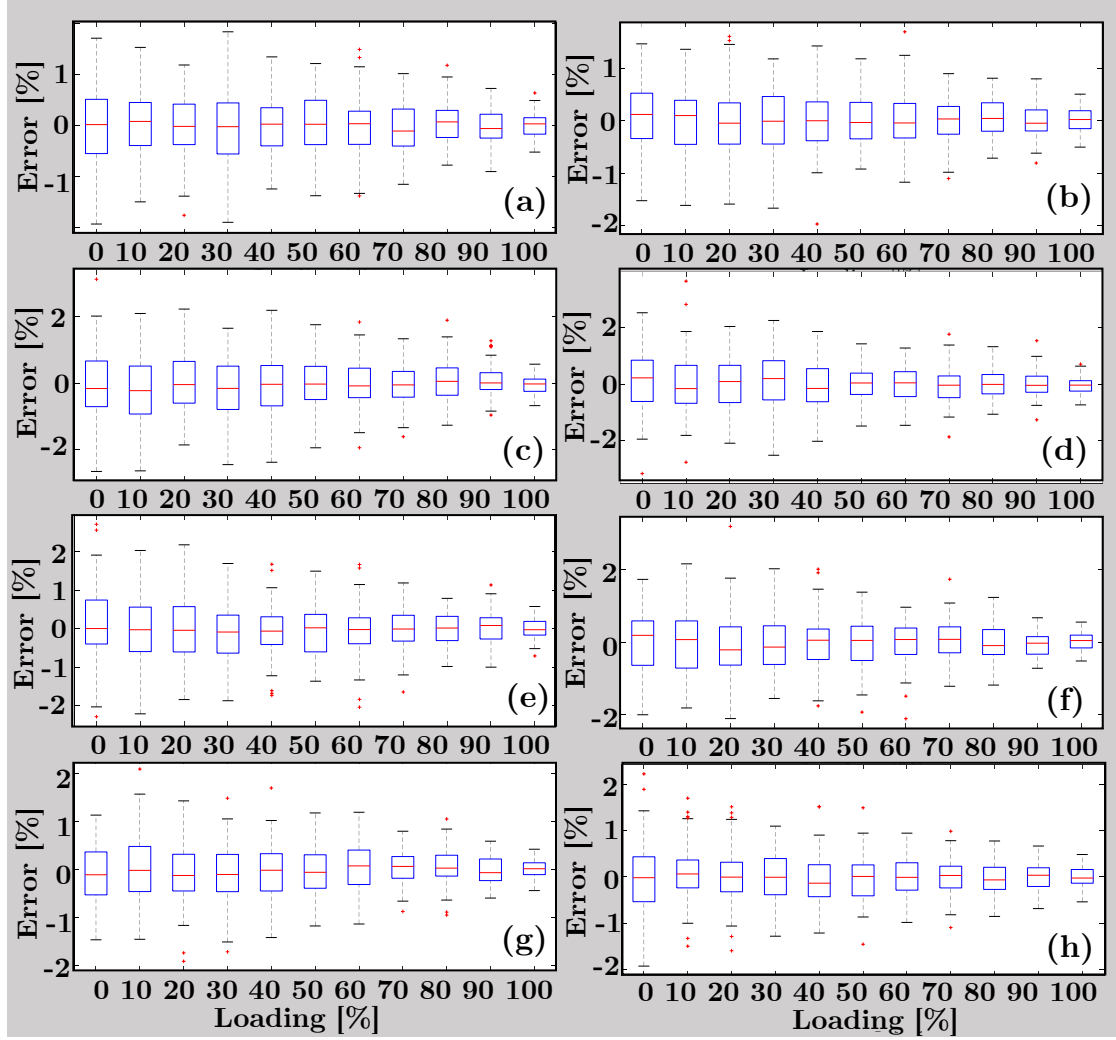


Figure 6.11: Tracking performance of (a) active power, (b) reactive power under Type 1 HUT, (c) active power, (d) reactive power under Type 2 HUT, (e) active power, (f) reactive power under Type 3 HUT, and (g) active power, (h) reactive power under Type 4 HUT, of the proposed robust PHIL interface method.

Chapter 7

Conclusions and Future Research

7.1 Summery of Completed Works

This thesis makes several contributions in developing a *robust, dynamic, resilient* microgrid that promises dramatic transformations in the way people make and consume reliable, resilient, clean and economically viable energy.

In Chapter 2, a generalized μ -synthesis-based robust control framework is designed utilizing the fact that there is a voltage-current duality in the plant dynamic model of GFL and GFM inverter. The uncertainties in grid impedance parameters and uncertainties in equivalent loading parameters for GFL and GFM inverters are modeled respectively. The generalized control framework results in the controllers that are single-loop, hence simple and cost-effective enough to be implemented, and optimal, in the sense of robustness in performance under uncertainties. The resulting current-controller for GFL inverter provides inherent active damping under grid parameter variation whereas the resulting voltage-controller for GFM inverter enhances the dynamic performance during load transients as desired by a *robust* microgrid.

Chapter 3 develops a seamless transition strategy for microgrid using GFM inverters. It is shown both analytically and experimentally that none of the inverters supply power to the network while operating in on-grid mode where the entire load is supplied by the grid as specified. During off-grid mode, both inverters share the required demand between them with a seamless transition from on-grid mode following a grid failure. The viability of the proposed control is corroborated by CHIL-based simulation studies for initial prototyping of the control strategy. A large-scale hardware experiment on real commercial inverters and the results validate the functionality of the proposed control

in the seamless transition and improved power flow control as desired by a *dynamic* microgrid.

In Chapter 4, a novel synchronization controller is designed that compensates the phase angle and voltage magnitude difference between the islanded microgrid and the incoming grid side to zero by adjusting the droop laws of all GFM inverters of the microgrid dynamically via secondary control layer. It is analytically shown that compensation on only phase angle and voltage magnitude is sufficient for synchronization as frequency is compensated by this proposed method indirectly by the phase angle compensation. A systematic approach for tuning the proposed synchronization controllers is also provided. It is concluded from the validation that the proposed synchronization method is relatively fast and makes the re-connection with the incoming grid relatively smooth with less transients as desired by a *dynamic* microgrid.

Chapter 5 is devoted to proposing a comprehensive design of the emergency power supply system for the uninterrupted operation of a microgrid by employing various novel techniques, including 1) novel mode-dependent droop-controlled GFM inverters for seamless transition capability; 2) a fast-acting optimal net-load management engine for the efficient and optimal operation of microgrid; and 3) a novel optimization-based HoV engine for longer, sustained, and viable operation in the aftermath of a grid failure as recommended in the standards. Both CHIL- and PHIL-based studies using defined test cases on the test system based on the realistic electrical network of a commercial-scale medical center are performed in the validation stage. The results corroborate the efficacy and empirical robustness of the proposed strategy for a *resilient* microgrid by achieving good regulation limits and power quality measures while considering the enhancements as recommended by the standards.

In Chapter 6, a μ -synthesis-based robust PHIL interface controller is developed. In the synthesizing process of the interface controller, two fundamental model uncertainties are quantified and characterized systematically. It includes model uncertainty caused by the point-of-connection in the emulated power network of the software system and the model uncertainty caused by the varying nature of the hardware-under-test of the hardware system. It is shown analytically that the proposed interface controller guarantees robust stability and robust performance under model uncertainties. Extensive PHIL demonstrations corroborate the fact that the proposed robust PHIL interface has improved accuracy, and tracking performance under varying conditions of software and hardware system. This design methodology is the first to design a multiple-input multiple output optimal controller.

7.2 Future Works

After delineating the contributions in developing a *robust, dynamic, resilient* microgrid, this thesis also highlights a few eminent open research questions that must be addressed to enable the widespread adoption of microgrids in future electric grid. A summary of open research topics that warrant further investigation is provided below:

- A relevant direction of future study is to investigate the nuances of the traditional voltage controllers of grid-forming IBRs. A grid-forming IBR becomes a pivotal entity for future microgrids. State-of-the-art covers various types of structures of voltage controllers for a GFM IBR that includes open-loop, single-loop and cascaded double-loop voltage controllers. The cascaded double-loop voltage controller, also known as the outer-voltage-inner-current, is the most common and well-accepted both in industry and academia. However, the current state-of-the-art lacks a good study that investigates the root cause of this dominance in acceptance of the double-loop structure over other structures of voltage controllers. However, it is essential to study and analyze the superiority of the double-loop voltage controller and document it. A re-look at the design objectives of a voltage control for a GFM IBR and establishing the problem statement in the robust control framework to investigate the optimality of the double-loop structure for voltage controller is a promising field of research that is currently under investigation.
- The works related to the seamless transition capability of microgrids offer room for several proximate extensions. The current method for the seamless transition from on-grid to off-grid condition, described in Chapter 3, is dominantly focused on the scenario of disengaging a microgrid from a stiff grid. The re-synchronization method for seamless and smooth off-grid to on-grid transition, described in Chapter 4, is also focused on the scenario of engaging a microgrid to a stiff grid. However, in multi-microgrid systems, it is important for every microgrid to have a similar seamless and smooth transition capability to engage/disengage to/from another microgrid. In the design stage of such seamless transition methods, it is essential to consider the stiffness of the microgrid systems. The role of controllers of the IBRs, especially the droop-based control law and its modification for developing such capabilities in microgrids is required to be investigated further.
- One essential extension in the PHIL-based DRTS system is to understand the phenomenon of parallel operation of multiple PHIL interfaces. In microgrids,

multiple DERs are dispersed across the network and connected at different point-of-connections. Therefore, operating multiple HUTs interfaced to the emulated system via multiple PHIL interface controllers running in parallel becomes necessary to emulate such realistic scenarios. However, most of the existing works on the design of PHIL interface controllers usually deal with a single HUT interfaced to the emulated system via a single PHIL interface controller. Current state-of-the-art lacks a study and analysis of the issues and challenges of operating multiple PHIL interfaces in parallel. It is necessary to understand the impact of one PHIL interface controller on another PHIL interface controller when both are running in parallel. A sufficient study of the impact will potentially form the base for modifying the existing type of interface controllers or designing a new type of interface controller to ensure stable operation in parallel. Hence, the study and design of PHIL interface controllers suitable for parallel operation is a further promising field of research. This extension is currently under investigation.

In conclusion, there are many challenging open questions regarding the design and implementation of a *robust, dynamic, resilient* microgrids. The author hopes that the work conducted in this thesis may help to answer some of these questions and serve as a base for a large variety of challenging future research directions.

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Appendix A

A.1 Reference Generation for Grid-following Inverter

The outer ‘Reference Generation Block’ of Fig. 2.1(a) eventually generates the i_{ref} signal using its pre-specified reference active power, P_{ref} , and reactive power, Q_{ref} , (defined locally/centrally) and output signals from phase-locked loop (PLL). The expression of i_{ref} is given by

$$i_{\text{ref}} = \sqrt{2} \frac{\sqrt{P_{\text{ref}}^2 + Q_{\text{ref}}^2}}{\tilde{V}} \sin \left(\tilde{\theta} - \arctan \frac{Q_{\text{ref}}}{P_{\text{ref}}} \right), \quad (\text{A.1})$$

where, a $1\text{-}\phi$ second order generalized integrator-based synchronous reference frame PLL (SOGI-SRF-PLL) operates with its grid-synchronization technique and generates the RMS value, $\tilde{V}$, and synchronized phase information, $\tilde{\theta}$, of v_{O} [204].

A.2 Reference Generation for Grid-forming Inverter

A $P\text{-}f/Q\text{-}V$ droop control strategy is adopted for outer ‘Reference Generation Block’ of Fig. 2.1(b). The droop characteristic equations are as follows [205]:

$$\omega = \omega_{\text{N}} - n_{\text{P}}P, \quad V = V_{\text{N}} - m_{\text{Q}}Q, \quad (\text{A.2})$$

$$v_{\text{ref}} = \sqrt{2}V \sin \left(\int \omega dt \right), \quad (\text{A.3})$$

where, ω_{N} and V_{N} are nominal frequency (in rad/sec) and voltage (RMS) respectively. P and Q are averaged active and reactive power output of GFM inverter. n_{P} and m_{Q} are the droop coefficients and the values are typically chosen such that ω and V are within

the allowed specification, defined by IEEE 1547 Standard [71], for all $P \in [0, P_{\text{rated}}]$ and $Q \in [-Q_{\text{rated}}, Q_{\text{rated}}]$ respectively. Here, P_{rated} and Q_{rated} are the rated active and reactive power of the GFM inverter.

A.3 Shaping Transfer Functions

$$\begin{aligned}
\mathcal{W}_s &= \frac{5.97 \times 10^9}{(s + 5278)^2} \times \frac{s^2 + 314s + (2\pi 60)^2}{s^2 + 31.4s + (2\pi 60)^2} \\
&\quad \times \frac{s^2 + 1.2 \times 10^4 s + 1.5 \times 10^8}{s^2 + 245s + 1.5 \times 10^8} \\
\mathcal{W}_{\text{CS}} &= 100 \times \frac{s + 2\pi 60}{s + 1500 \times 2\pi 60} \\
\mathcal{W}_d &= \frac{7539.8}{s + 7540} \times \frac{s^2 + 628.1s + (2\pi 60)^2}{s^2 + 31.4s + (2\pi 60)^2} \\
&\quad \times \frac{s^2 + 314s + 9 \times (2\pi 60)^2}{s^2 + 31.4s + 9 \times (2\pi 60)^2} \\
&\quad \times \frac{s^2 + 314s + 25 \times (2\pi 60)^2}{s^2 + 31.4s + 25 \times (2\pi 60)^2} \\
&\quad \times \frac{s^2 + 314s + 49 \times (2\pi 60)^2}{s^2 + 31.4s + 49 \times (2\pi 60)^2}
\end{aligned}$$

Appendix B

B.1 System Model in On-grid Condition

System Model in on-grid condition with $\mathcal{I}_{\text{gs}} = 1$:

$$\dot{\theta}_2 = \omega_{\text{r},2} - \omega_{\text{r},1}, \text{ where } \theta_2 := \theta_{\text{r},2} - \theta_{\text{r},1}, \quad (\text{B.1})$$

$$\dot{\theta}_{\text{g}} = \omega_{\text{grid}} - \omega_{\text{r},1}, \text{ where } \theta_{\text{g}} := \theta_{\text{grid}} - \theta_{\text{r},1}, \omega_{\text{grid}} = \omega_{\text{nom}}, V_{\text{g}} = V_{\text{nom}} \quad (\text{B.2})$$

$$\tau_{\text{S},1} \dot{\omega}_{\text{r},1} = \omega_{\text{nom}} - \omega_{\text{r},1} - n_{\text{i}} p_1, \quad (\text{B.3})$$

$$\tau_{\text{S},2} \dot{\omega}_{\text{r},2} = \omega_{\text{nom}} - \omega_{\text{r},2} - n_2 p_2, \quad (\text{B.4})$$

$$\tau_{\text{S},1} \dot{V}_{\text{r},1} = k_{\text{m},1} (V_{\text{nom}} - V_{\text{r},1} - m_{\text{int},1} \psi_1^{\text{Q}}) - m_1 q_1, \quad (\text{B.5})$$

$$\tau_{\text{S},2} \dot{V}_{\text{r},2} = k_{\text{m},2} (V_{\text{nom}} - V_{\text{r},2} - m_{\text{int},2} \psi_2^{\text{Q}}) - m_2 q_2, \quad (\text{B.6})$$

$$m_1 \dot{\psi}_1^{\text{Q}} = V_{\text{nom}} - V_{\text{r},1} - m_{\text{int},1} \psi_1^{\text{Q}}. \quad (\text{B.7})$$

$$m_2 \dot{\psi}_2^{\text{Q}} = V_{\text{nom}} - V_{\text{r},2} - m_{\text{int},2} \psi_2^{\text{Q}}. \quad (\text{B.8})$$

$$L_{\text{l},1} \dot{i}_{\text{o},1}^{\text{d}} = V_{\text{r},1} - R_{\text{l},1} i_{\text{o},1}^{\text{d}} - v_{\text{PCC}}^{\text{d}} + \omega_{\text{r},1} L_{\text{l},1} i_{\text{o},1}^{\text{q}}, \quad (\text{B.9})$$

$$L_{\text{l},1} \dot{i}_{\text{o},1}^{\text{q}} = -R_{\text{l},1} i_{\text{o},1}^{\text{q}} - v_{\text{PCC}}^{\text{q}} - \omega_{\text{r},1} L_{\text{l},1} i_{\text{o},1}^{\text{d}}, \quad (\text{B.10})$$

$$L_{\text{l},2} \dot{i}_{\text{o},2}^{\text{d}} = V_{\text{r},2} \cos \theta_2 - R_{\text{l},2} i_{\text{o},2}^{\text{d}} - v_{\text{PCC}}^{\text{d}} + \omega_{\text{r},1} L_{\text{l},2} i_{\text{o},2}^{\text{q}}, \quad (\text{B.11})$$

$$L_{\text{l},2} \dot{i}_{\text{o},2}^{\text{q}} = V_{\text{r},2} \sin \theta_2 - R_{\text{l},2} i_{\text{o},2}^{\text{q}} - v_{\text{PCC}}^{\text{q}} - \omega_{\text{r},1} L_{\text{l},2} i_{\text{o},2}^{\text{d}}, \quad (\text{B.12})$$

$$L_{\text{l},\text{g}} \dot{i}_{\text{g}}^{\text{d}} = V_{\text{g}} \cos \theta_{\text{g}} - R_{\text{l},\text{g}} i_{\text{g}}^{\text{d}} - v_{\text{PCC}}^{\text{d}} + \omega_{\text{r},1} L_{\text{l},\text{g}} i_{\text{g}}^{\text{q}}, \quad (\text{B.13})$$

$$L_{\text{l},\text{g}} \dot{i}_{\text{g}}^{\text{q}} = V_{\text{g}} \sin \theta_{\text{g}} - R_{\text{l},\text{g}} i_{\text{g}}^{\text{q}} - v_{\text{PCC}}^{\text{q}} - \omega_{\text{r},1} L_{\text{l},\text{g}} i_{\text{g}}^{\text{d}}. \quad (\text{B.14})$$

$$L_{\text{L}} \dot{i}_{\text{L}}^{\text{d}} = v_{\text{PCC}}^{\text{d}} - R_{\text{L}} i_{\text{L}}^{\text{d}} + \omega_{\text{r},1} L_{\text{L}} i_{\text{L}}^{\text{q}}, \text{ where } i_{\text{L}}^{\text{d}} = i_{\text{g}}^{\text{d}} + i_{\text{o},1}^{\text{d}} + i_{\text{o},2}^{\text{d}} \quad (\text{B.15})$$

$$L_{\text{L}} \dot{i}_{\text{L}}^{\text{q}} = v_{\text{PCC}}^{\text{q}} - R_{\text{L}} i_{\text{L}}^{\text{q}} - \omega_{\text{r},1} L_{\text{L}} i_{\text{L}}^{\text{d}}, \text{ where } i_{\text{L}}^{\text{q}} = i_{\text{g}}^{\text{q}} + i_{\text{o},1}^{\text{q}} + i_{\text{o},2}^{\text{q}} \quad (\text{B.16})$$

B.2 System Model in Off-grid Condition

System Model in off-grid condition with $\mathcal{I}_{\text{gs}} = 0$:

$$\dot{\theta}_2 = \omega_{\text{r},2} - \omega_{\text{r},1}, \text{ where } \theta_2 := \theta_{\text{r},2} - \theta_{\text{r},1}, \quad (\text{B.17})$$

$$\tau_{\text{S},1}\dot{\omega}_{\text{r},1} = \omega_{\text{nom}} - \omega_{\text{r},1} - n_1[p_1 - P_{\text{ref},1}], \quad (\text{B.18})$$

$$\tau_{\text{S},2}\dot{\omega}_{\text{r},2} = \omega_{\text{nom}} - \omega_{\text{r},2} - n_2[p_2 - P_{\text{ref},2}], \quad (\text{B.19})$$

$$\tau_{\text{S},1}\dot{V}_{\text{r},1} = V_{\text{nom}} - V_{\text{r},1} - m_1[q_1 - Q_{\text{ref},1}], \quad (\text{B.20})$$

$$\tau_{\text{S},2}\dot{V}_{\text{r},2} = V_{\text{nom}} - V_{\text{r},2} - m_2[q_2 - Q_{\text{ref},2}], \quad (\text{B.21})$$

$$L_{\text{l},1}\dot{i}_{\text{o},1}^{\text{d}} = V_{\text{r},1} - R_{\text{l},1}i_{\text{o},1}^{\text{d}} - v_{\text{PCC}}^{\text{d}} + \omega_{\text{r},1}L_{\text{l},1}i_{\text{o},1}^{\text{q}}, \quad (\text{B.22})$$

$$L_{\text{l},1}\dot{i}_{\text{o},1}^{\text{q}} = -R_{\text{l},1}i_{\text{o},1}^{\text{q}} - v_{\text{PCC}}^{\text{q}} - \omega_{\text{r},1}L_{\text{l},1}i_{\text{o},1}^{\text{d}}, \quad (\text{B.23})$$

$$L_{\text{l},2}\dot{i}_{\text{o},2}^{\text{d}} = V_{\text{r},2} \cos \theta_2 - R_{\text{l},2}i_{\text{o},2}^{\text{d}} - v_{\text{PCC}}^{\text{d}} + \omega_{\text{r},1}L_{\text{l},2}i_{\text{o},2}^{\text{q}}, \quad (\text{B.24})$$

$$L_{\text{l},2}\dot{i}_{\text{o},2}^{\text{q}} = V_{\text{r},2} \sin \theta_2 - R_{\text{l},2}i_{\text{o},2}^{\text{q}} - v_{\text{PCC}}^{\text{q}} - \omega_{\text{r},1}L_{\text{l},2}i_{\text{o},2}^{\text{d}}, \quad (\text{B.25})$$

$$L_{\text{L}}\dot{i}_{\text{L}}^{\text{d}} = v_{\text{PCC}}^{\text{d}} - R_{\text{L}}i_{\text{L}}^{\text{d}} + \omega_{\text{r},1}L_{\text{L}}i_{\text{L}}^{\text{q}}, \text{ where } i_{\text{L}}^{\text{d}} = i_{\text{o},1}^{\text{d}} + i_{\text{o},2}^{\text{d}} \quad (\text{B.26})$$

$$L_{\text{L}}\dot{i}_{\text{L}}^{\text{q}} = v_{\text{PCC}}^{\text{q}} - R_{\text{L}}i_{\text{L}}^{\text{q}} - \omega_{\text{r},1}L_{\text{L}}i_{\text{L}}^{\text{d}}, \text{ where } i_{\text{L}}^{\text{q}} = i_{\text{o},1}^{\text{q}} + i_{\text{o},2}^{\text{q}} \quad (\text{B.27})$$