

Smart three-phase power converter for more electric powertrains

by

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Abstract

In this dissertation, a smart three-phase power converter for more electric powertrains is presented. In the context of a more electric powertrain, the *smartness* of a power converter is tied to its ability to adapt to rapid input-frequency/voltage variations, load changes, and extreme ambient conditions fluctuations while regulating the desired variables with seamless dynamics and satisfactory power quality. Founded on these principles, this dissertation advances the state-of-the-art in power converter control schemes for more electric powertrains. More specifically, for ac-dc three-phase power converters fed by a variable-speed electric generator. At first, a model predictive step-ahead controller is developed and experimentally validated. This control scheme's uniqueness lies in its integration with an instantaneous phase-detection technique, which enables ultrafast dynamics for the dc-bus and power factor regulation when the power converter is connected to a variable-frequency/constant-amplitude input voltage supply. Despite the step-ahead controller's ultrafast dynamic performance, it is shown and experimentally verified that it suffers a performance degradation when there is a parameter mismatch between hardware and controller due to the variable-frequency and ambient operating conditions in a more electric powertrain. Accordingly, a Lyapunov-based adaptive parameter estimation algorithm that provides accurate fast-tracking of the system's parameters is developed and experimentally validated in this dissertation. The estimated parameters are fed to the step-ahead controller, minimizing the detrimental effects of parameter mismatch. However, despite its sound performance and enhanced capabilities, the step-ahead control scheme shows a reactive power steady-state tracking error proportional to the input frequency. Consequently, a direct model reference adaptive control (MRAC) scheme is developed and experimentally verified in this dissertation. The novelty

of this control scheme resides in its ability to seamlessly regulate the output dc-bus voltage and input reactive power in the presence of rapid input-frequency/voltage variations, load changes, and system parameter variations. These smooth dynamics are achieved through a multi-variable direct MRAC formulation, which allows the controller's gains to automatically adjust to external perturbations without any knowledge of the system's parameters. The validity and performance effectiveness of all the developed control schemes is verified experimentally through a proof-of-concept laboratory-scaled three-phase 1.5kW 270V SiC-based two-level voltage-source-converter (2L-VSC) using a variable-frequency programmable grid emulator. Furthermore, a traditional proportional-integral (PI) controller is implemented experimentally with the same 2L-VSC as a benchmark to highlight the merits of the developed controllers. Lastly, to further validate the concepts introduced in this dissertation, all the developed controllers are experimentally tested under variable-frequency/variable-amplitude input voltage, which emulates a more electric powertrain variable-speed permanent magnet generator that is directly connected to its primer mover, e.g., internal combustion engine or jet engine. Herein, the superiority of the developed MRAC scheme is once again experimentally verified in comparison to the step-ahead and PI-based controllers, thus solidifying the proposed smart converter concept for more electric powertrains.

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Dedication

To my family.

Chapter 1

Introduction

This introductory chapter contains four sections. The first section describes the background of the problem. Subsequently, the second section presents the literature review of state-of-the-art technologies. The third section articulates the objectives and contributions of this dissertation in three-phase ac-dc converters for more electric powertrain. Lastly, section four introduces the general structure and organization of this dissertation.

1.1 Background of the Problem

Modern transportation systems are shifting towards more electrification, i.e., electrified-mobility (E-Mobility), technologies to address fuel efficiency and emission requirements, as well as market demands for lower operational costs. Power electronic converters are the enabling technology leading the technological transition towards E-Mobility [1]. As a first step toward all-electric transportation, many airborne and hybrid vehicles are adopting the *power-by-wire* concept to achieve higher system efficiencies and reduce fuel consumption while still implementing conventional carbon-based energy sources. In these vehicles, subsystems that were conventionally powered by hydraulic, pneumatic, or mechanical energy sources are electrified [2]. Two clear examples of such vehicles are hybrid electric vehicles (HEV) [3], and more-electric/hybrid aircraft [4, 5].

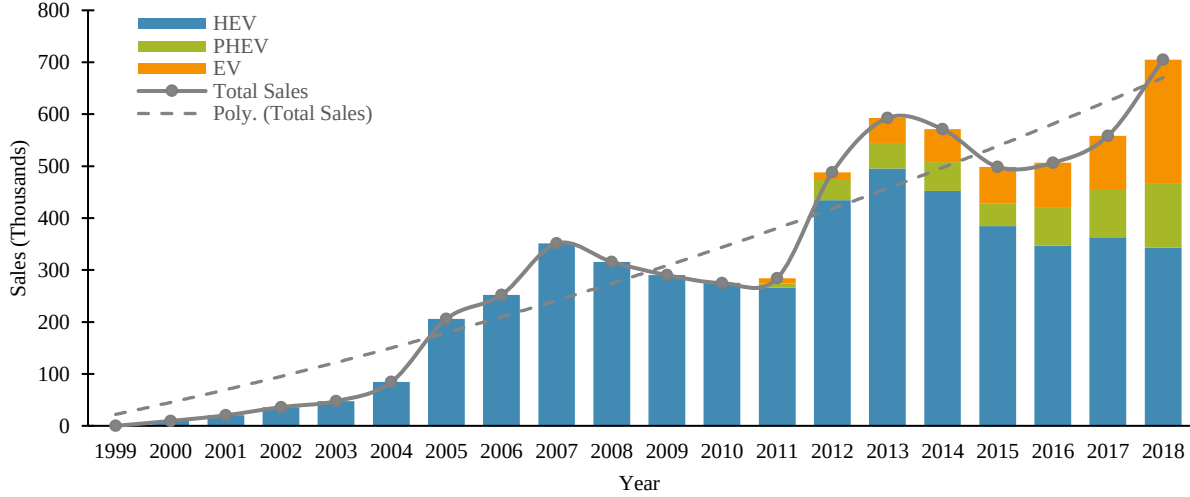


Figure 1.1: Hybrid (HEV), plug-in (PHEV), and all-electric (EV) vehicles sold between 1999-2018 in the United States of America [3].

Automotive electrification has dramatically increased over the last decade with hybrid and all-electric vehicles leading the trend [6]. Fig. 1.1 depicts the Hybrid (HEV), the plug-in (PHEV), and all-electric (EV) vehicles sold between 1999-2018 in the United States of America [3]. Here, since 1999, the majority of the EV sold were hybrid. This type of EVs combines classical internal combustion engines (ICE) with electric motors and energy storage devices to achieve higher miles-per-gallon ratios throughout the whole speed range of the vehicle [7]. The combined operation of the ICE and electric motor leads to a wide variety of topologies for the configuration of the powertrain. In general, the main powertrain topologies can be categorized as series, parallel, and parallel/series (power-split), with the series and power-split architectures being the most popular in the automotive industry [8]. Herein, the onboard electric power is generated by a variable-speed electric generator directly driven by an ICE. The voltage generated by the machine is converted to dc using a three-phase ac-dc converter or active rectifier. Depending on the electric machine utilized, the input voltage of the converter can exhibit a variable-frequency/constant amplitude or variable-frequency/amplitude. In both cases, the active rectifier is subject to rapid input frequency/voltage variations that demand novel control schemes to address such external perturbations.

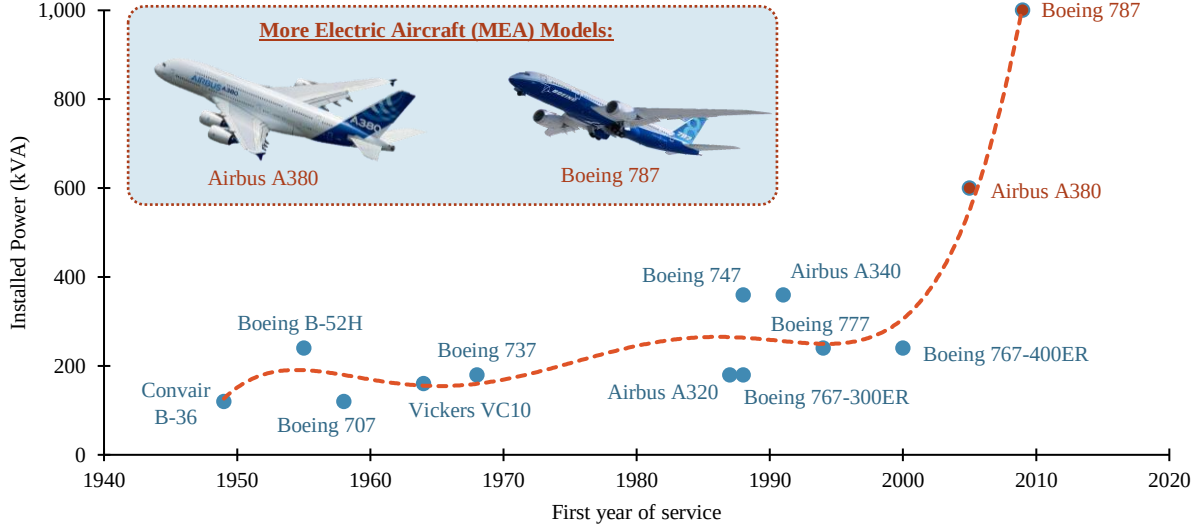


Figure 1.2: Evolution of onboard electrical power generation of some common aircraft [4].

On the other hand, from the beginning of commercial and military aircraft flight operation, the onboard electrical power generation has gradually increased throughout the years, reaching maximums of 600 kVA and 1,000 kVA for the Airbus A380 and Boeing 787, respectively. Fig. 1.2 shows the evolution of the onboard electrical power generation of some common aircraft models starting in 1949 up to the start of the MEA era in 2005 [9,10]. Since then, MEA technologies have enabled aerospace manufacturers to reduce fuel consumption, weight, environmental impact, and operating costs by electrifying the majority of the onboard subsystems [2]. More specifically, subsystems that were conventionally powered by hydraulic, pneumatic, or mechanical energy sources are now electrified, thus reducing the fuel consumption of the aircraft by approximately 9% [10].

The electrification of the aircraft translates into a substantial increase in the onboard electric power requirements, hence shifting the attention of researchers to the power generation system and its associated power converters [11]. In most legacy aircraft, a variable-ratio transmission gearbox is used to mechanically couple the jet engines with the main and auxiliary electric generators. As a result, the generators produce an ac voltage with a constant magnitude and frequency. Conversely, in a MEA, the gearbox is removed, and the jet engines are directly coupled with the variable-speed generators [4]. This variable-speed operation

condition imposes novel design challenges in terms of the electric generator's ac-dc power converters, which must now operate with a variable-amplitude/frequency input voltage.

Despite the remarkable fuel consumption reduction figure achieved in MEA, modern aircraft must further reduce their emissions to keep up with national and global subsonic airliner performance goals, such as the 2020-2030 NASA 40-70% fuel burn reduction, and the European Commission in the Flightpath 2050 with 75% CO₂ emission decrease [12]. In a MEA, 90% of the fuel is used for propulsion [13]. Consequently, a definitive step toward fuel and cost reduction can be taken by shifting toward a more electric powertrain architecture, in which the propulsion of the aircraft is carried out by electric motors and/or turbines. In light of this, the National Academies of Sciences, Engineering, and Medicine's Committee on Propulsion and Energy Systems to Reduce Commercial Aviation Carbon Emissions has outlined several hybrid and turboelectric powertrain architectures that can potentially close the gap between the more and all-electric aircraft [5].

Finally, taking the aerospace and automotive industries as examples of how modern transportation systems are transitioning toward E-mobility by implementing more electric powertrains in hybrid propulsion architectures, power electronic converters are now more than ever the center of this technological transition acting as the main enablers. Moreover, the simplification or elimination of mechanical conversion systems, such as gearboxes and complex transmissions, constitutes a significant advance in E-Mobility transportation systems. Herein, the onboard power generation system will be driven at variable-speed. This variable-speed operation can generate a wide-range variable-frequency/constant-amplitude or a variable-frequency/amplitude set of voltages, which constitutes an entirely novel operation scenario for power converters that are typically implemented at constant industrial frequency levels in grid-tied [14–21], stand-alone applications [22–24], or narrower variable-frequency ranges in motor drives or renewable energy interfacing [25]. Furthermore, onboard any moving vehicle, power electronic converters are exposed to changeable ambient temperature and pressure conditions that can produce important fluctuations in the parameters of the systems. In this vein, it is clear how the new generation of power converters for more electric powertrains requires an extra layer of awareness or *smartness* that allows the

converter to adapt to external perturbations while controlling the desired variables. It is then based on this philosophy, that the concept of *smart* power converters for more electric powertrains becomes not only a reality but an active research area for the automobile and aerospace industry.

1.2 Literature Review

In this section, a literature review of the main components of a more electric powertrain is presented. Starting with electric machines, i.e., motors and generators, and followed by power converters, this section briefly describes the state-of-the-art technologies of the main apparatus present in a more electric powertrain. Finally, since the focus of this dissertation is on the ac-dc power converter and its controller, the last subsection presents a comprehensive literature search on control schemes applicable to three-phase ac-dc power converters.

1.2.1 Electric Machines for More Electric Powertrains

Hybrid and all-electric propulsion systems or powertrains have proven their effectiveness in the automotive, locomotive, and marine vehicles [7]. In these applications, the electric motors are only a few hundred kilo-Watts; whereas, aircraft powertrains are expected to range from hundreds of kW to several MW, even up to tens of MW for wide-body airplanes. As a result, hybrid/electric vehicle (HEV/EV) powertrains can tolerate machines with 1-3 kW/kg power density. Conversely, an aircraft propulsion motor is required to have above 10 kW/kg to minimize the weight of the aircraft [5, 13, 26]. Fig. 1.3 shows the state-of-the-art electric machine power density as a function of power. Here, Induction Machine (IM), Radial Flux Permanent Magnet Machine (RFPM), Synchronous Reluctance Machine (SynRM), Switched Reluctance Machine (SRM), Wound-Field Synchronous Machine (WFSM), and Axial-Flux Permanent Magnet Machine (AFPM) are compared showing the advantages of certain configurations over others. Although it is difficult to establish a clear trend per-machine type, Fig. 1.3 demonstrates how far the current state-of-the-art is from the requirements of a mod-

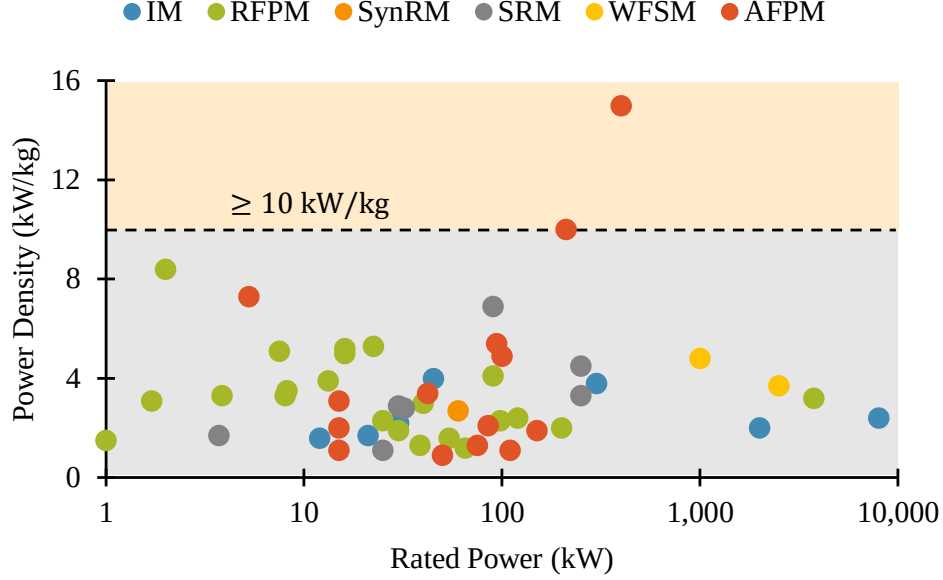


Figure 1.3: State-of-the-art electric machine power density as a function of power [13].

ern more electric powertrain. In fact, the authors of [5] project that only after 20 years, a power density of 9 kW/kg will be achieved for machines with 1-3 MW. Based on these projections, innovative and transforming designs must be explored to make an electric powertrain feasible.

1.2.1.1 Variable-Speed Electric Generators

Stand-alone power systems, such as those present in vehicles and aircraft, typically use dc-distribution systems to supply part of the onboard load and charge the energy storage devices. Often, this dc-voltage is generated with an electric generator–rectifier combination [4]. The classical topology for such systems comprises a WFSM and a passive multi-pulse diode-based rectifier [27]. In this vein, the concept of variable-speed electric generators, where the prime mover engine is directly coupled to a WFSM, translates to a variable-frequency/constant-amplitude voltage generation.

An example of a variable-frequency/constant-amplitude voltage power system can be found in a MEA, where the jet engines are directly coupled with a three-stage WFSM [2]. Consequently, the output voltage of the generators exhibit a variable-frequency characteristic that ranges between 360 – 800 Hz [11]. For MEA applications, the predominant state-of-the-

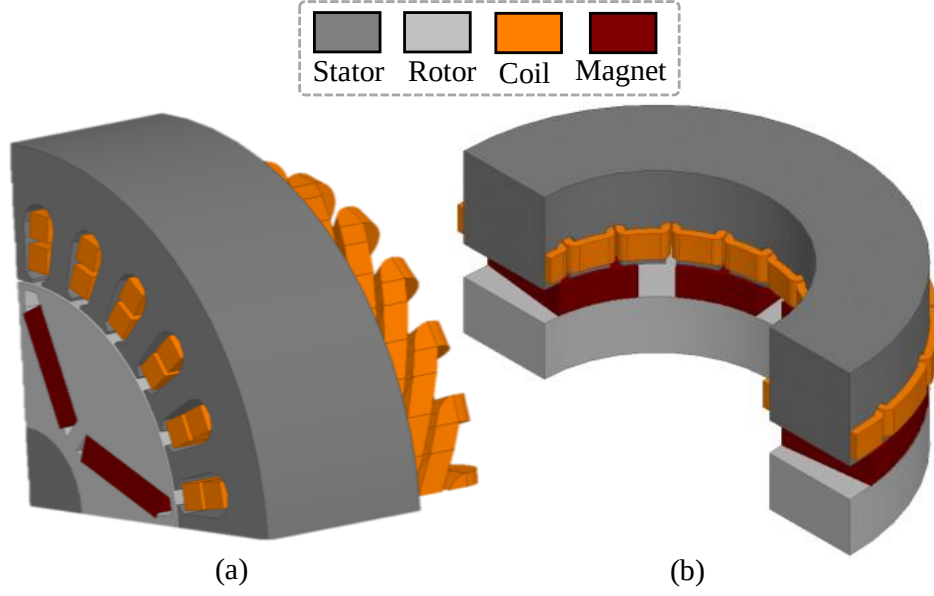


Figure 1.4: Permanent magnet synchronous machines; (a) radial and (b) axial flux.

art machine-related research is centered around two types of electromechanical converters: (i) permanent magnet synchronous generator (PMSG), and (ii) hybrid excitation synchronous generator (HESG) [28]. Several papers have studied the control of the PMSG with its active rectifier in MEA applications [27,29–32]. These papers only focus on the control scheme of the active rectifier, in which the reactive power circulation is minimized to reduce the losses in the generator. Conversely, in [28,33,34], the emphasis of the research is on the electromechanical design of the HESG in MEA applications. Similarly, in [35], an HESG is implemented as a starter-generator for HEV applications. However, in the automobile industry, the PMSG is the proffered choice given its high power density and efficiency [7,8,36].

1.2.1.2 Propulsion Electric Motors

Similar to electric generators, permanent magnet (PM) machines have increasingly replaced dc and induction motor for traction applications [7]. These machines offer high power density and efficiency at the expense of higher costs due to the permanent magnet materials implemented in their construction. Recently, AFPM, see Fig. 1.4(b), have gained popularity in the transportation electrification sector due to their high-power density and specific power compared to classic radial PM machines, Fig. 1.4(a) [37]. Furthermore, in [37], the

authors prove that for a given magnet material and air-gap flux density, the rotor moment of inertia in an axial flux machine tends to be smaller compared to its radial counterpart. Thus resulting in faster transient responses of the motor drive system. On the other hand, in terms of their cost, the axial flux motor is less costly than radial flux machines [38]. Based on [37–39], axial flux motors are generally regarded as ideally suited for direct-drive medium speed high torque applications and postulate themselves as the ideal electric motor to carry the technological transition toward the more electric powertrain.

AFPM motors have been studied for the past 30 years, reaching technological maturity in a wide variety of applications, from renewable energy to transportation. Recent advances in AFPM machines are summarized in [38]. Herein, the maximum power density reported is 2 kW/kg for 15 kW, 4500 r/min machine for use as a generator directly coupled to the internal combustion engine in a series HEV drive-train [40]. In [13], the authors outline the state-of-the-art AFPM motors for aircraft propulsion available both as commercial models and research-generated prototypes. Here, the reported power densities for AFPM motors range from 0.9-4.9 kW/kg, with the EMRAX early prototypes achieving the maximum reported power density. A 94 kW AFPM model with 5.4 kW/kg power density is presented in [41]. Conversely, higher power densities are now possible for the EMRAX 348 rated at 210 kW with a power density of 10 kW/kg at a maximum speed of 4000 r/min [42]. The highest commercial AFPM motor prototype is the Magnax AXF275 rated at 400 kW with a power density of 15 kW/kg [43]. Despite their attractive power density levels, state-of-the-art AFPM exhibit power rating in the lower-hundreds kW, see Fig. 1.3, limiting their propulsion applications to EV in-wheel traction [38]. With this in mind, future research in AFPM motors must tackle this limitation to achieve high power density machines at the MW range.

A further glance in the future of electric motors may postulate superconducting machines as attractive candidates for aerospace applications due to their high efficiency and reduced weight [44]. Recent studies conducted by NASA estimate that a cryogenic motor design can achieve 125 times the power density of a conventional electric motor, and 3.3 times the power density of a large turbine engine, particularly for larger capacity airplanes [45]. With this in mind, researchers at the NASA Glenn Research Center have oriented their efforts

toward developing MW high-efficiency superconducting electric propulsion motors such as the High Efficiency Megawatt Motor (NASA-HEMM) with a superconducting rotor. Despite their benefits, superconductors introduce several challenges such as cryogenic temperature requirements, loss of superconductivity, fragility compared to copper and aluminum, and cost [46].

1.2.2 Power Converters for More Electric Powertrains

For years the automobile and aerospace industry have relied on power electronic converters to interconnect the onboard electric systems. The continuous advancements in digital computing and semiconductors have amplified the usage of power converters in aircraft and any electrified vehicle [1]. The MEA power system is a clear example of power converters acting as the technological enablers in a system with dc and ac loads at different voltage levels. However, in a MEA, the power converters are designed to be robust and reliable at the expense of low-functionality and control scheme complexity.

As industries advance toward more electric powertrains, the onboard power converters must now act as the pillars of each powertrain architecture to guarantee performance, reliability, and fault tolerance under any operating scenario. These characteristics are the key features of a smart converter. A smart converter takes a step beyond the conventional power converter as it is able to smartly adapt to input variations, load changes, and system parameter variations while providing ancillary services with fault-tolerant capabilities [47–50]. As such, the smart converter is the key enabler to the more electric powertrain.

1.2.2.1 DC-AC Power Converters

In an electric powertrain, the motor drive converter is in charge of controlling the propulsion motor. This converter must be bidirectional, as it must offer regenerative braking capabilities. Typically from an electric perspective, the dc-ac converter and its associated motor are considered as one in terms of efficiency and power density. As a result, for a more electric powertrain, the two main desired features for a dc-ac converter are minimum losses

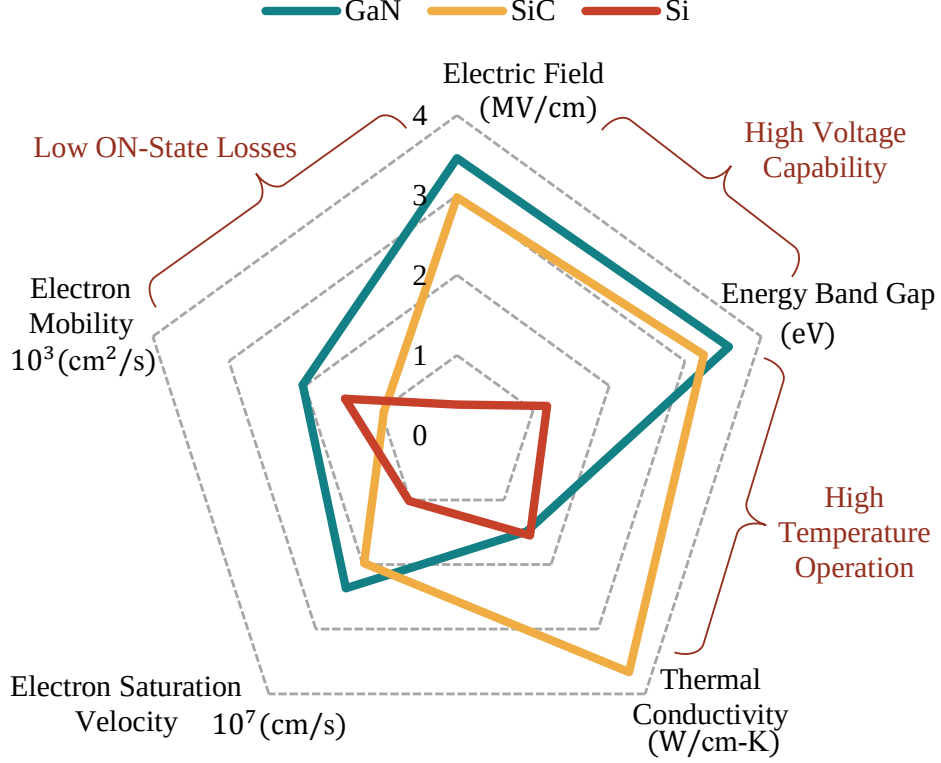


Figure 1.5: Graphical summary of Si, SiC, and GaN relevant material properties [51,52].

and weight. The recently emerging wide-bandgap semiconductors, i.e., Silicon Carbide (SiC) and Gallium Nitride (GaN), are the key enablers to make electric powertrains more efficient. These semiconductors offer lower conduction and switching losses at higher temperatures and lower footprints than their Silicon counterparts [53,54], as shown in Fig. 1.5. Although the current GaN switches available in the market offer limited voltage and current ratings, it is expected that a full GaN inverter could potentially outperform both Si- and SiC-based inverters in the power range of tens or hundreds of kW scale [55]. For MW-scale power converters, high-voltage SiC modules, such as 1.7 kV SiC modules, can be used to boost the efficiency and power density. It should be noted that the switching dv/dt and di/dt with SiC and GaN devices are much higher than that with the conventional Silicon counterparts, so extra attention should be paid to the electromagnetic interference (EMI) filter design, in order to meet industry EMI standards. Finally, in addition to efficiency and power density, a propulsion inverter must be fault-tolerant as it is responsible for the propulsion of the vehicle. Fault-tolerant motor drives have been extensively discussed in the literature [47],

and their application to more electric powertrains could be the next step.

1.2.2.2 AC-AC Power Converters

Among ac-ac power converters, direct and indirect matrix converters have been proposed for MEA applications due to their advantages of lower weight and smaller physical volume, as well as higher reliability due to the elimination of dc-link capacitors which are vulnerable components typically existing in conventional voltage source converters. In [56], a direct matrix converter is developed for A320 electro-hydrostatic aircraft actuator (EHA) system, in which the 3-phase matrix converter provides direct ac-ac power converter to feed a high-speed permanent magnet motor. In a direct matrix converter, 9 arrays of bi-directional semiconductor switches are implemented. Each semiconductor switch is configured by two IGBTs connected in anti-series. An input LC filter is typically required to attenuate harmonics related to the converter switching frequency. No bulky intermediate dc-link capacitors or dc inductors are needed, thus enabling high-power density and reliability.

In [57], an indirect matrix converter (IMC) has been proposed for aircraft starter/generator systems. The circuit topology of the IMC includes two power stages, the front-end current-source-rectifier, and the back-end voltage-source-inverter. The rectifier stage is composed of six bi-directional switches, with each switch containing two IGBTs connected in common source, while the inverter stage consists of six unidirectional switches, similar to the conventional voltage-source-inverter. Although such power architecture looks similar to the conventional ac-dc-ac two-stage power converter, no dc-link energy storage components are needed for an IMC. One additional benefit offered by the IMC topology is the dc-link access to interconnect a battery energy storage system in MEA, which is not possible for a direct matrix converter [58].

1.2.2.3 DC-DC Power Converters

Buck-boost dc-dc converters interface between the energy storage system and the power converters. In MEA, simple buck dc-dc topologies are implemented to generate 28Vdc for

the instrumentation equipment [2]. Nonetheless, for an all-electric or hybrid powertrain, the dc-dc converter must be bidirectional, efficient, and light. Bidirectional dc-dc converters have been studied for years [59], and the leap between a classic and smart dc-dc converter resides in its controller.

1.2.2.4 AC-DC Power Converters

Active rectifiers or ac-dc power converters are implemented in hybrid and turboelectric powertrain architectures, see Fig. 2.6, to convert the electric generator ac output voltage to a constant dc-bus (≥ 270 V) from where the electric propulsion drive is connected. In some cases, the dc-bus can also be equipped by an energy storage system, e.g., hybrid architectures. The main functionalities of an electric powertrain ac-dc converter are to regulate the dc-bus voltage, control the power factor on the generator-side, and maintain a low total harmonic distortion (THD) in the generator currents [60].

Active rectifier topologies and control schemes have been extensively studied throughout the years [60–62]. Nonetheless, in a MEA and potentially a hybrid or turboelectric powertrain, the generator is directly coupled with the jet engine, thus generating a variable-frequency/constant-amplitude three-phase ac-voltage. In a MEA, the ac-dc conversion is performed by a passive multi-pulse autotransformer rectifier unit (ATRU), which generates low-order harmonics and offers no power factor regulation [63]. Consequently, the new generation of smart ac-dc converters must operate at variable-frequency and potentially at variable-amplitude/frequency when the electric generator is a PM synchronous machine.

1.2.3 Control Schemes for Three-Phase AC-DC Power Converters

The control of grid-tied ac-dc three-phase power converters has been well documented in the literature throughout the years with several review articles [60–62]. However, variable-frequency applications have started to recently attract the attention of researchers. In [27,29], the control scheme of an ac-dc three-phase power converter and its electric generator is designed based on the machine’s dynamics, adding complexity to the controller. On the other

hand, the authors in [64], demonstrate that a six-switch three-level Vienna-Type rectifier [65] is a good fit for MEA variable-frequency applications, with the trade-off of having higher conduction losses compared to a conventional two-level-voltage-source-converter (2L-VSC) topology. To address this issue, the authors in [66] developed a two-level Δ -switch rectifier (2L- Δ SR) with reduced conduction losses that utilizes fewer semiconductors than the three-level Vienna rectifier, but still requires three additional thyristors than the 2L-VSC topology. In light of this, the authors in [67] presented a discontinuous conduction mode (DCM) three-phase buck-boost power factor corrector converter with reduced semiconductor count. Nonetheless, the DCM operation increases the current stress in the switches. Furthermore, the converter developed in [67] requires three inductors to boost the input voltage to $270 V_{dc}$. That being said, regardless of their benefits and disadvantages, all the previously mentioned power converters topologies and controllers offer significant advantages compared to a passive diode converter, such as the ATRU used in MEA.

In addition to semiconductor count, which impacts the reliability of the system, a ac-dc power converter or active rectifier for more electric powertrain applications must be equipped with a controller that provides rapid and effective dynamics when connected to a variable-speed electric generator that is driven by an ICE or jet engine. These prime mover can rapidly vary their speeds, thus demanding fast responses from the active rectifier control scheme. Among the numerous controllers implemented for three-phase ac-dc converters, model predictive control (MPC) has become one of the preferred choices for applications where the dynamic performance is the main priority [68]. Multiple authors have presented articles comparing the performance of MPC techniques with proportional-integral (PI)-based linear control [69]. Many of these papers explore the use of finite control set MPC (FCS-MPC) in applications such as (i) 2L-VSC [70,71], (ii) matrix converters [72], and (iii) electric drives [73,74].

Despite its merits, MPC schemes are known to suffer from performance degradation under model parameter mismatch [75,76], which can be a relevant factor in a more electric powertrain since the parameters of the system may variate with the frequency and ambient conditions [77–80]. To address performance degradation due to parameter mismatch in MPC

schemes, the authors in [81] implement an integrator in the FS-MPC algorithm cost function to minimize the steady-state current error in an H-bridge converter. This solution requires the tuning of an additional weight factor that may deteriorate the dynamic performance of the controller [82]. Conversely, the authors in [83] propose an adaptive deadbeat MPC that minimizes the current steady-state current error by computing the resistance of the inductive input filter of the power converter. Herein, it is assumed that the inductance of the filter remains relatively unchanged, which is not applicable to a more electric powertrain variable-frequency system. Alternatively, in [76], both the inductance and parasitic resistance of the inductive filter (L filter) are estimated online using a least-squares matrix approach for a three-phase FS-MPC active rectifier. Herein, the previously mentioned parameters are accurately estimated in steady-state operation. In [84], the authors propose a disturbance observer-based MPC technique that exhibits constant switching frequency. This controller requires the tuning of the observer gains, which can affect the performance of the system since their adjustment tends to compromise the bandwidth of the controller. Instead, the authors in [85] propose a technique based on the combination of the control schemes presented in [83] and [81] in conjunction with a trajectory-based algorithm. Herein, the developed controller demonstrated improved performance in comparison with the FS-MPC with integrator solution presented in [81] for single-phase converters.

Newer studies related to parameter estimation in MPC-based power converters can be found in [86–89]. In [86], a Lyapunov-based estimation algorithm is applied for a sensorless-FS-MPC three-phase active rectifier. In this work, the parasitic resistance of the inductive filter is not estimated; thus the performance degradation associated with the resistance variations is not addressed. Alternatively, in [87], only the inductance of the filter is estimated as part of a deadbeat MPC scheme in a modular multi-level converter. The estimation is performed employing the discrete-time model of the system, which requires a computationally expensive moving average filter to ensure stability. Finally, in [88], a frequency-adaptive virtual flux (VF) estimator-based MPC with dc-bus ripple suppression is presented for an active rectifier supplied by an unbalanced grid. Herein, the double frequency ripple in the dc-bus and active power is minimized by adaptively estimating the frequency and compensating for its

variations. The advantages of the controller are valid even under filter inductance parameter mismatch. However, for an inductance parameter mismatch, the reactive power exhibits a steady-state error.

In all the aforementioned MPC techniques, the voltage and frequency at the ac-input of the power converter are assumed constant or quasi-constant, e.g., $50\text{-}60\text{ Hz} \pm 0.1\text{ Hz}$. This assumption limits the implementation of the controllers to grid-tied applications, hence excluding variable-frequency/amplitude voltage applications such as a more electric powertrain. Furthermore, FS-MPC schemes have a variable switching frequency that can produce undesirable Electromagnetic interference (EMI) propagation, audible noise, and obscure the heatsink design of the converter [90].

As systems become more complex and automated, adaptive control schemes become an attractive and feasible solution. In particular, direct model reference adaptive control (MRAC) schemes are now implemented in applications such as (i) active filters [91,92], (ii) electric drives [93,94], (iii) and power converters [95–97]. The main advantage of MRACs is that they combine system parameter identification and control into one scheme [98]. Herein, the gains of the control scheme are continually adjusted online based on the operating point of the system to force the control variables of the plant to follow the response of a pre-designed reference model. In [91], the authors implement a MRAC to minimize the harmonics and reactive components of a single-phase nonlinear load utilizing a linearized second-order model of the plant. Alternatively, in [92], a third-order nonlinear model is used to adaptively estimate the parameters and control a three-phase shunt active filter. Lastly, the authors in [95] present simulation results for an adaptive control combined with a H_∞ algorithm for a three-phase ac-dc power converter.

1.3 Objectives and Contributions of the Dissertation

All-electric transportation systems are envisioned as the ultimate target for zero-emission vehicles. Nonetheless, the technological transition from fuel-based to all-electric is mainly limited by the advances in energy storage systems. As a result, for automobiles and aircraft,

the technology intersecting fuel-based and all-electric vehicles is a hybrid system. These architecture allow the combination of mature carbon-based powertrains with modern electric apparatus, resulting in the so-called more electric powertrain. Enabled by power electronic converters, a more electric powertrain uses a combination of energy sources to drive vehicles efficiently. To achieve higher reliability, efficiency, and reduce weight, modern more electric powertrains directly couple the prime mover of the vehicle, e.g., internal combustion or jet engine, to an electric machine that generates the onboard power. Depending on the type of machine, the generated voltages exhibit a variable-frequency/constant amplitude or variable-frequency/amplitude pattern. As a result, more electric powertrain power converters are subject to rapid input-frequency/amplitude variations, load changes, and extreme ambient conditions fluctuations; thus requiring novel control schemes that not only regulate the desired variable but also adapt to external perturbations. More specifically, this dissertation focuses on the three-phase ac-dc converter that is directly connected to the main electric generator. The function of this active rectifier is to regulate its output dc-bus voltage while minimizing the stator copper losses by operating the generator at unity power factor. The general structure of the system investigated in this dissertation is presented in Fig. 1.6.

In this sense, the main objective of this dissertation is to advance the state-of-the-art in power converter control schemes for more electric powertrains. More specifically, for ac-dc three-phase power converters fed by a variable-speed electric generator. Aligned with the aforementioned objective, this dissertation brought forth the following contributions:

- An ultrafast model predictive step-ahead controller for more electric powertrains is developed and experimentally validated. This control scheme's uniqueness lies in its integration with an instantaneous phase-detection technique, which enables ultrafast dynamics for the dc-bus and power factor regulation when the power converter is connected to a variable-frequency/constant-amplitude input voltage supply [99, 100]. The merit of this contribution is to offer an active rectifier solution to substitute the passive rectifier in several applications, such as MEA.
- This research presented a comprehensive performance comparative experimental anal-

ysis between a classical PI-based model-based control scheme and the step-ahead controller developed in this dissertation. The performance comparison is executed with a variable-frequency/constant-amplitude input voltage supply applicable to more electric powertrains. This comparison investigates the dynamic performance of the controllers in terms of dc-bus voltage regulation and unity power factor operation under load steps increase/decrease, frequency step-changes, and frequency ramping [63].

- An indirect adaptive online parameter estimation technique is developed and experimentally validated to minimize performance degradation under parameter mismatch between hardware and controller due to the variable-frequency and ambient operating conditions in a more electric powertrain [101, 102]. The merit of this contribution is to enhance the performance of the step-ahead controller by providing accurate online parameter tracking.
- A direct model reference adaptive control (MRAC) scheme for more electric powertrains is developed and experimentally validated. The novelty of this control scheme resides in its ability to seamlessly regulate the output dc-bus voltage and input reactive power in the presence of rapid input-frequency/voltage variations, load changes, and system parameter variations. These smooth dynamics are achieved through a multi-variable direct MRAC formulation, which allows the controller's gains to automatically adjust to external perturbations without any knowledge of the system's parameters [103]. The merit of this contribution is to provide seamless dynamics without any information of the system.
- This research experimentally investigated all the developed controllers under variable-frequency/variable-amplitude input voltage, which emulates a more electric powertrain variable-speed permanent magnet generator directly connected to its primer mover, e.g., internal combustion engine or jet engine. Herein, the superiority of the developed MRAC scheme is once again verified in comparison to the step-ahead and PI-based controllers [104].

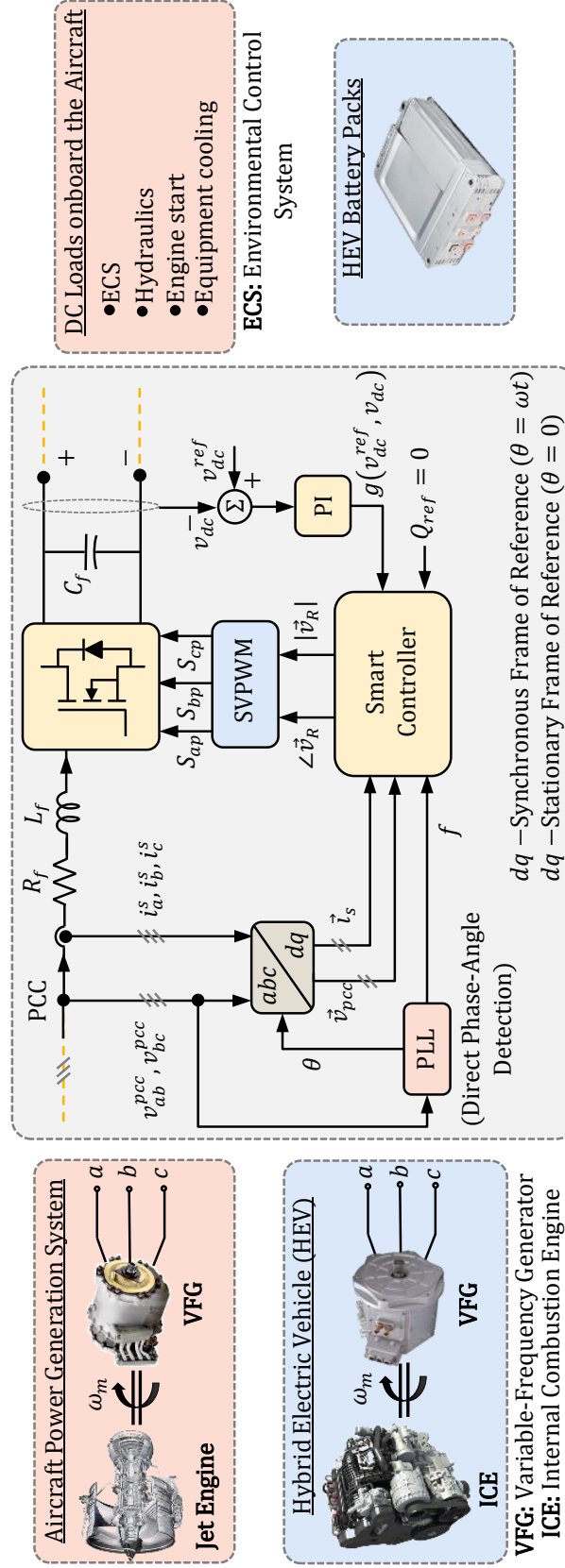


Figure 1.6: General block diagram of the specific system investigated in this dissertation. Left column, input of the system. Middle column, the developed smart power convert for more electric powertrains. Right column, output of the system.

1.4 Organization of the Dissertation

After an introductory chapter, Chapter 1, in which the problem statement, objectives, and contributions of the dissertation, as well as a literature search, were presented, Chapter 2 presents a detailed description of a more electric powertrain system. Herein, two main technologies have been outlined that highlight the usage of more electric powertrains (i) hybrid vehicles and (ii) more-electric/hybrid/all-electric aircraft. These applications are described focusing on the importance and functionalities of power electronic converters. With this mind, the focus is shifted to the ac-dc three-phase active rectifier utilized in this work as the main converter. A comprehensive primitive coordinate and dq —arbitrary reference-frame modeling of three-phase active rectifiers is presented in Section 2.2, followed by the description of the phase-angle and frequency detection scheme utilized in this dissertation. Lastly, the classic model-based dq —control scheme used as the benchmark controller in this work is presented in Section 2.4.

Chapter 3 presents the mathematical formulation and implementation of the three-phase ac-dc power converter control schemes developed in this dissertation. Starting with the step-ahead model predictive controller, followed by the indirect adaptive step-ahead controller, and finalizing with the direct model reference adaptive controller.

Chapter 4 presents the experimental results validating and showcasing the merits of the control schemes developed in Chapter 3. In each case, the steady-state and dynamic performance is assessed, taking the model-based dq —controller as a benchmark, given its high industry implemented PI-based structure. The experimental results presented in Chapter 4 are obtained with a proof-of-concept laboratory-scale SiC-based 2L-VSC fed by a programmable grid emulator. The efficacy of each control scheme is evaluated under input-frequency ramps, frequency step-changes, load step-changes, and parameter mismatch. Similarly, the developed control schemes are experimentally tested under variable-voltage/frequency ramp and step-changes, with the results presented in the final section of Chapter 4.

Chapter 5 essentially provides a summary of this research work, along with some conclusions regarding the significant contributions of this dissertation. Suggestions are given for

future work on how the developed controllers can serve as a base for future hardware and system-level research paths for more electric powertrains in automobiles and aircraft.

Lastly, this dissertation contains three Appendix chapter (A) Arbitrary Reference Frame Theory for Power Converters, (B) A Model Predictive Control Scheme Formulation for Active Rectifiers with *LCL* Filter, and (C) Next Generation GaN Power Converter Hardware Development. Appendix A describes the $dq0$ —arbitrary reference-frame theory utilized as well as its origins from in electric machine theory up to its implementation in modern power electronics controllers. Appendix B presents a complementary model predictive control scheme formulation for an active rectifier using *LCL* filters. Appendix C presents the development of the new generation GaN-based power converter.

1.5 Summary

In summary, this chapter covered the background of the problem addressed in this dissertation, a literature review in electric machines for more electric powertrains, power converters for more electric powertrains, and control schemes for three-phase ac-dc power converters; the objectives and contributions, and organization of this dissertation.

Chapter 2

More Electric Powertrain System Description

This chapter describes the physical system, modeling, and control aspects of a more electric powertrain and its associated ac-dc converter. First, a description of hybrid and all-electric powertrains for vehicles and aircraft applications is presented. Here, hybrid electric vehicles, more electric aircraft, and hybrid/all-electric aircraft powertrains are explained. Section II presents the modeling of a three-phase active rectifier implemented in this work. Once the modeling of the converter is covered, Section III details the phase-angle detection technique used in this work. Lastly, the model-based dq -control scheme utilized in this work as the benchmark controller is explained.

2.1 Hybrid and All-Electric Powertrains for Vehicle and Aircraft Applications

Hybrid and all-electric powertrains can be implemented in vehicles and aircraft. The automotive industry has been implementing electric powertrains for a few decades, which have lead to mature designs of all-electric vehicles. That being said, hybrid vehicles still occupy almost half of the sales of the electric vehicles in the United States of America by

2018, as depicted in Fig. 1.1. Consequently, extensive power converter research is conducted for hybrid vehicles even after their maturity.

On the other hand, the aerospace industry is only starting the road toward electrification. The first step of the industry is the more electric aircraft (MEA), which enabled substantial fuel and weight savings. The success of the MEA has opened the door for hybrid and all-electric powertrain architectures for aircraft that will lead the efforts toward emission-free aerospace transportation and cargo system in the future.

2.1.1 Hybrid Electric Vehicles

The state-of-the-art hybrid/all-electric powertrain topologies can be categorized as series, parallel, and parallel/series (power-split) [8]. From the three previously mentioned powertrain HEV topologies, the power-split configuration has attracted the most attention among researchers and auto industries, such as Toyota, Lexus, General Motors, Mercury, Hyundai, Kia, Saturn, Nissan, and Mazda [105]. In this topology, the planetary gear splits the mechanical power coming from the ICE and electric traction motor to effectively drive the vehicle, and charge the battery packs with the electric generator. At any given road load point, the vehicle can be driven individually by the electric motor, ICE, or a combination of both. On the other hand, the battery packs are charged by the electric generator, driven by the ICE, and/or through the electric motor regenerative braking. Due to the mechanical properties of the planetary gear, the rotating speed of the ICE can be decoupled from the vehicle load, thus allowing the engine to operate efficiently. In many cases, depending on the energy optimization control strategy [8], the ICE will drive the electric generator at a constant speed, which results in a constant-amplitude/frequency voltage generation, assuming the electric generator is a PMSG [106, 107]. However, to operate the ICE at its maximum efficiency, the engine's speed must be continually varied depending on the required torque/power [108]; hence producing a variable-amplitude/frequency voltage profile, which directly impacts the design of the onboard power converters, i.e., ac-dc and dc-dc, and battery charging.

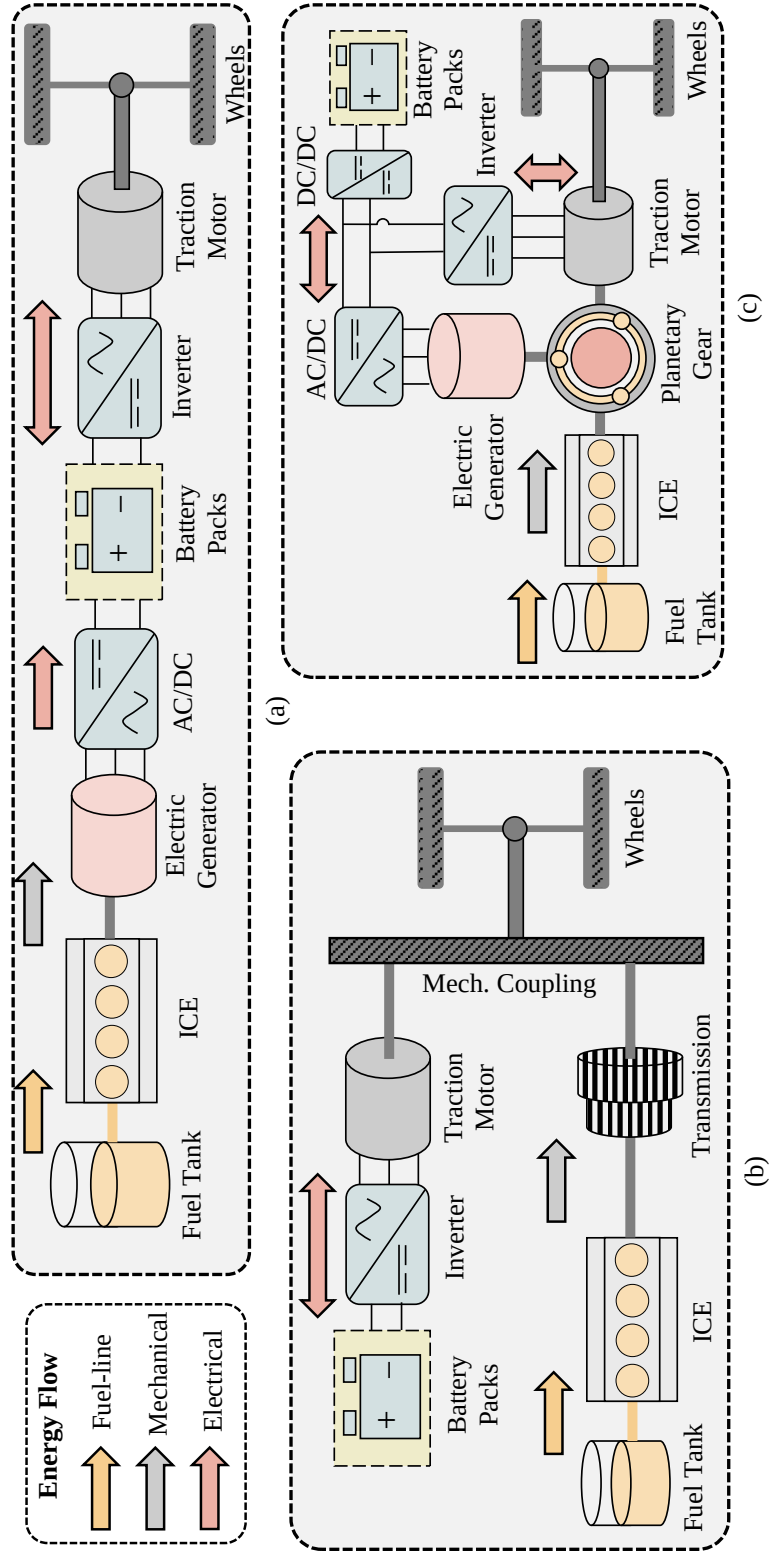


Figure 2.1: Hybrid vehicle powertrain configurations; (a) series, (b) parallel, and (c) parallel/series (power-split).

2.1.2 More Electric Aircraft

In most legacy aircraft, there are typically four sources of power derived from the jet engines (i) electrical, (ii) pneumatic, (iii) hydraulic, and (iv) mechanical. Approximately 96% of the engine's power is used for propulsion thrust, whereas from the remaining 4%, only 0.5% is used to produce electric power by means of a constant-speed gearbox-driven electric generator, as depicted in Fig. 2.2(a) [109,110]. The second alternative of a constant frequency generation system is presented in Fig. 2.2(b). Here, the constant-speed drive is removed, and a variable-speed electric generator is directly coupled with the jet engine. The variable-frequency/constant-amplitude voltage is then rectified such that an inverter can generate the constant-frequency voltage. Conversely, in a MEA, the gearbox is removed, and the jet engines are directly coupled with the variable-speed generators, as shown in Fig. 2.2(c) [4]. This variable-speed operation condition imposes novel design challenges in terms of the electrical, mechanical, and thermal design of the electric generators. Furthermore, these design challenges extend to the electric generator's ac-dc power converters, which must now operate with a variable-amplitude/frequency input voltage [63].

In a MEA power system, several variable-speed generators are connected in parallel to form the main ac-bus that feeds three separate subsystems [111]:

1. Variable-frequency ac-bus (115 V/200 V at 360 – 800 Hz): supplies approximately 40% of the on board loads.
2. Low-voltage dc (28 V): supplies close 10% of the onboard load, mainly consisting of flight instrumentation devices.
3. High-voltage dc-bus (± 270 V/540 V): supplies approximately 40% of the onboard load, e.g., hydraulics, engine start, equipment cooling, environmental control system (ECS) pressurization, ECS fans, ECS compressor, among others. Commercial MEA models such as the Airbus A380 and Boeing 787 use a ± 270 V dc-bus. Similarly, military aircraft, e.g., F-22, also implement a ± 270 V dc-bus. Higher voltage, e.g., 540 V, dc-bus configurations are currently under investigation for future MEA power systems [4,109].

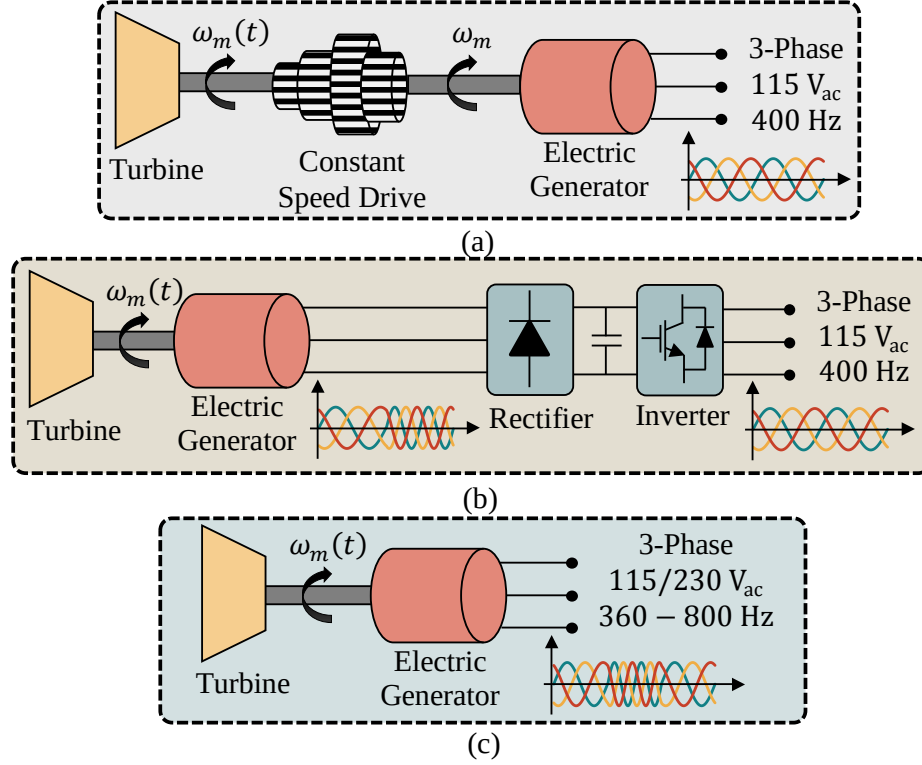


Figure 2.2: Aircraft electric power generation topologies; (a) constant-speed/constant-frequency, (b) variable-speed/constant-frequency, (c) and more electric aircraft variable-speed/variable-frequency.

Modern MEA utilize an ATRU to convert the variable-frequency voltage into ± 270 V_{dc}. The ATRU is known for its low circuit complexity and reliability [64]. Fig. 2.3 shows a typical 12-pulse ATRU conversion system formed by one phase-shifting isolation or autotransformer, two six-pulse diode bridge rectifiers, and a dc-bus electrolytic capacitor. Herein, the windings of the phase-shifting transformer are connected such that no circulation path is provided for the 5th and 7th input current harmonics, typically present 6-pulse diode bridge rectifiers. Consequently, the input line currents are predominately sinusoidal, as only harmonics with ordinal numbers, $n = 11, 13, 23, 25$ are present. Figs. 2.4(a) and 2.4(b) depict the simulated phase-to-neutral and line currents for the 12-pulse ATRU presented in Fig. 2.3 at 400 Hz and 800 Hz. Moreover, Figs. 2.4(c) and 2.4(d) show the input current harmonic spectrum for 400 Hz and 800 Hz, respectively. The simulations were performed using Matlab/Simulink at 1 kW resistive load, 115 V line-to-line input voltage, and 270 V_{dc} output.

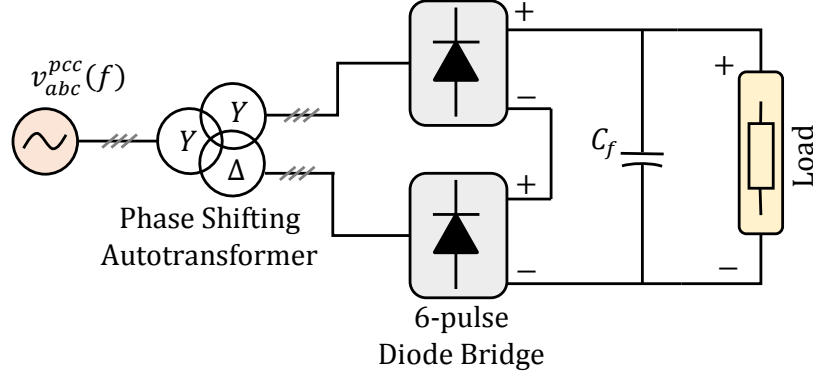


Figure 2.3: 12-Pulse ATRU circuit topology [63].

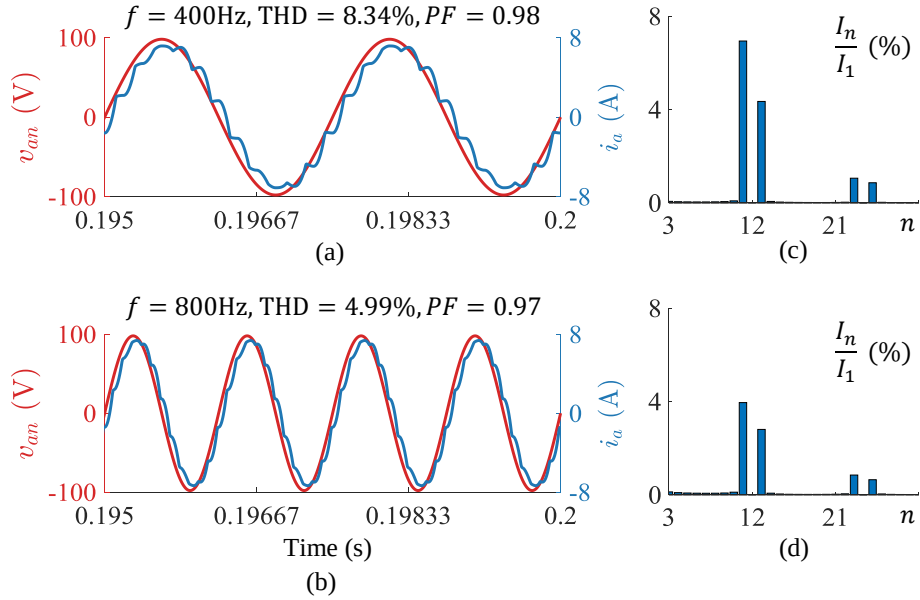


Figure 2.4: Simulation obtained steady-state waveforms of the phase-to-neutral voltages, line currents, and input current harmonics as a percentage of the fundamental component of the line current I_1 ; (a) and (c) at 400 Hz, and (b) and (d) at 800 Hz [63].

As demonstrated in Fig. 2.4, the ATRU has three main disadvantages that can be summarized as follows [64]:

1. Current THD is inversely proportional to the frequency, resulting in high THDs as the input frequency decreases.
2. Presence of low-order harmonics, e.g., $n = 11, 13, 23, 25$.
3. Load-dependent reactive power circulation.

Lastly, based on the simulation results shown in Fig. 2.4 and the investigation presented in [64], it is clear that the next step for a MEA power system is to replace the ATRU with an active rectifier to improve the power quality and efficiency of the system.

2.1.3 Hybrid and All-Electric Aircraft Powertrains

A typical aircraft propulsion system has three main blocks, see Fig. 2.5(a), (i) energy storage and power generation, (ii) powertrain, and (iii) propulsor. Most commercial aircraft implement a propulsion system consisting of a gas turbine engine powered by jet fuel, which drives a fan typically mounted to the turbine duct, i.e., turbofan engine, as shown in Fig. 2.5(b). The efficiency, defined as output shaft power divided by input fuel energy flow, for modern large commercial aircraft gas turbines is now about 55%. In contrast, a typical propulsor efficiency can achieve up to 80%, which totals a fuel-to-propulsor efficiency of about 45% [5]. Increasing this efficiency requires transitioning to a more electric propulsion system, as depicted in Fig. 2.5(c). Here, combined energy sources, together with high-efficiency electric machines and power converters, can potentially dramatically reduce the fuel burn.

An aircraft more electric propulsion system, as shown in Fig. 2.5(c), is formed by combined energy sources (jet fuel and electric energy storage device), electric machines (generators and motors), power converters, and a propulsion fan. These systems can be categorized into three main categories (i) all-electric, (ii) hybrid-electric, and (iii) turboelectric [5]:

2.1.3.1 All-Electric

The all-electric powertrain, as described in Fig. 2.6(a), uses batteries as the only energy source onboard. This architecture can achieve zero-emissions. Nonetheless, it is strongly tied to the development of high power density energy storage devices, e.g., batteries, fuel cells, capacitors, super-capacitors, among others.

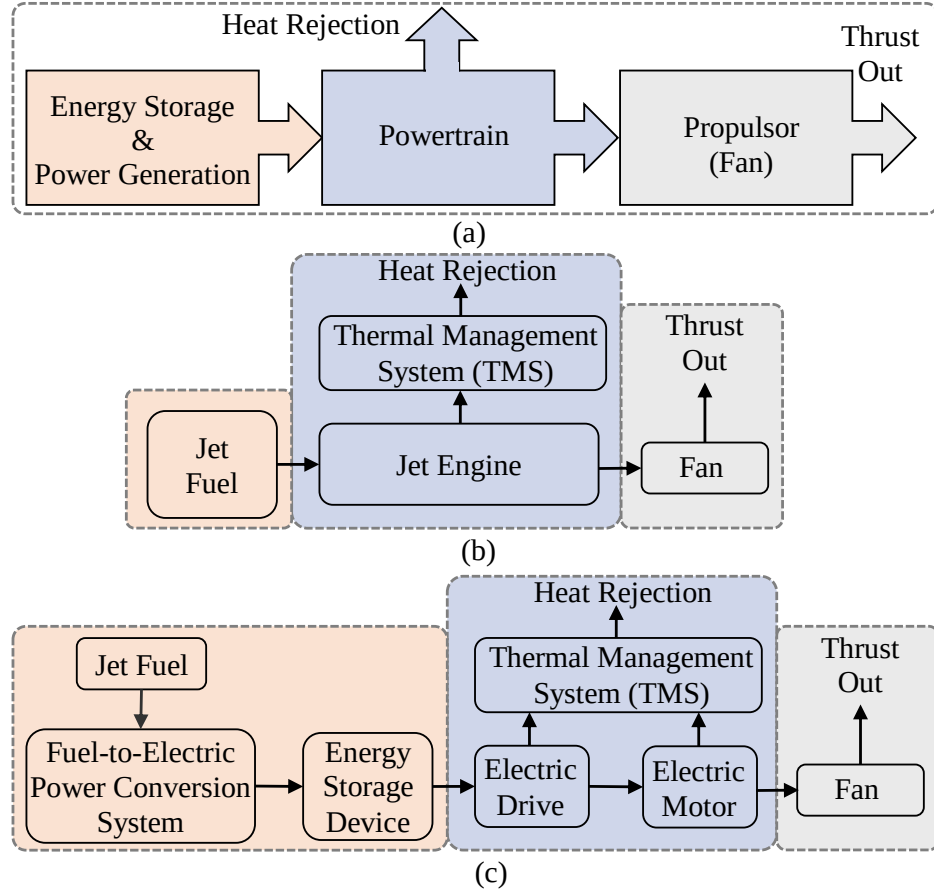


Figure 2.5: Schematic block diagram of an overall aircraft propulsion system; (a) general structure, (b) classic jet engine-based propulsion, (c) and more-electric propulsion.

2.1.3.2 Hybrid-Electric

The hybrid-electric powertrain uses a combination of jet fuel and electric energy storage devices as the onboard energy sources. Depending on the architecture, the propulsion is performed by an electric motor and/or a jet engine. The hybrid-electric powertrain architecture can be subdivided into three categories (i) parallel, (ii) series, and (iii) series/parallel. In the parallel hybrid architecture, as shown in Fig. 2.6(b), a battery-powered electric motor and a turbine are both mounted on a single shaft that drives the fan, such that either or both can provide propulsion any given time.

Conversely, in a series configuration, see Fig. 2.6(c), the propulsion is performed only by the electric motor. Herein, a gas turbine is implemented to drive an electric generator, which powers the propulsion motor drive and/or charges the battery packs. Lastly, in a

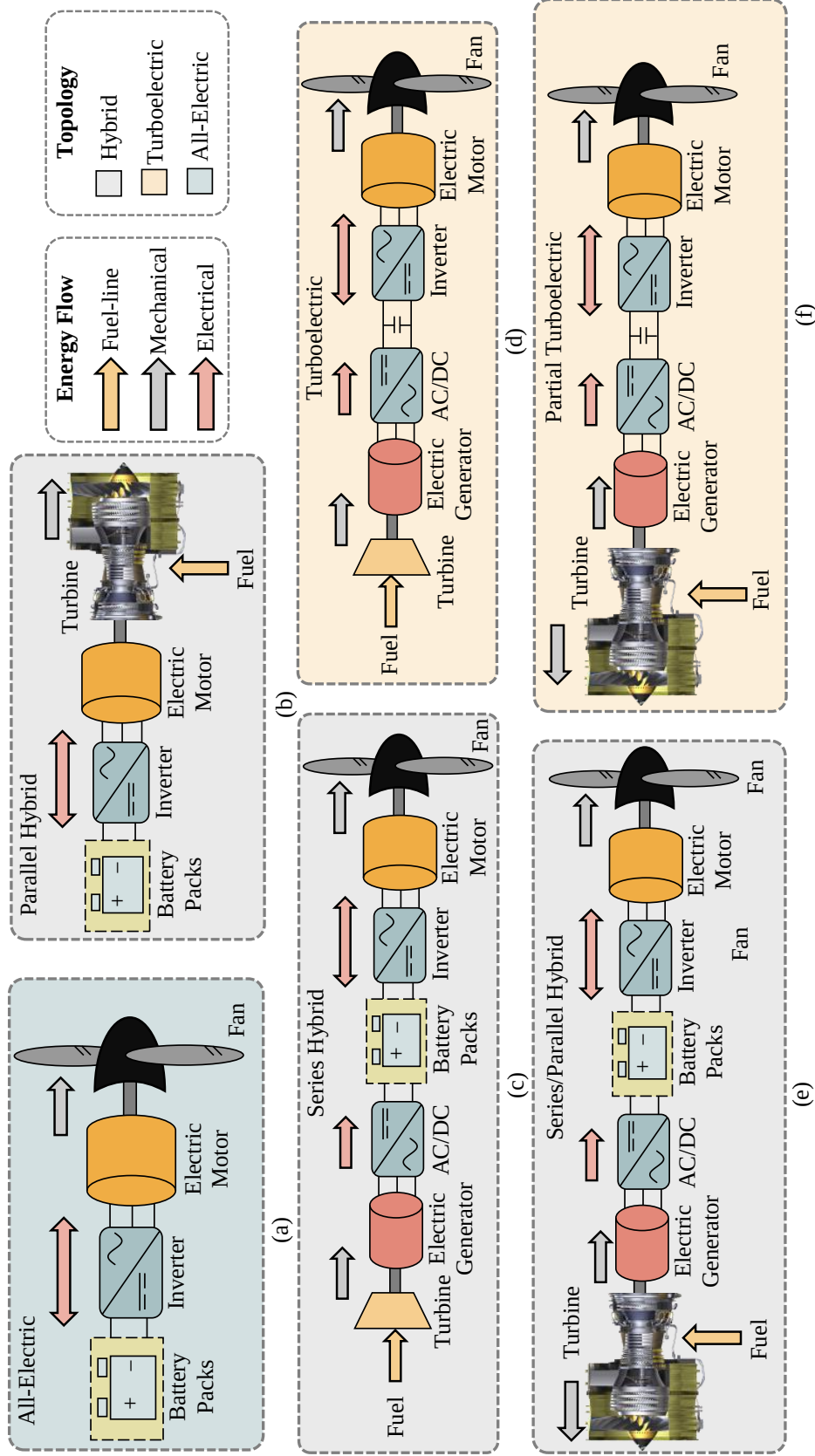


Figure 2.6: Aircraft powertrain architecture; (a) all-electric, (b) parallel hybrid, (c) series hybrid, (d) turboelectric, (e) series/parallel hybrid, and (f) partial turboelectric [5, 13].

series/parallel hybrid configuration, see Fig. 2.6(e), the propulsion of the aircraft can be achieved using one or multiple fans. The fans can be driven by a turbine or an electric motor; these motors can be powered by the battery and turbine-driven electric generator.

2.1.3.3 Turboelectric

A turboelectric powertrain in its full or partial configuration, as shown in Figs. 2.6(d) and 2.6(f), does not rely on any electric energy storage system to power the electric propulsion motor. Similar to the series hybrid architecture, a full turboelectric system uses a turbine to drive an electric generator that powers the propulsion electric motor. In contrast, the partial turboelectric system uses both a turbine and an electric motor to provide propulsive power. Due to the lack of combined energy sources, a turboelectric propulsion system with no additional aircraft developments is less efficient than a conventional gas turbine engine. In this configuration, the fuel-to-propulsor efficiency drops almost 10% compared to modern gas turbine engine due to the losses of the electric devices. However, with the implementation of distributed propulsion and boundary-layer ingestion, a turboelectric system can achieve higher efficiencies [112].

As of now, both hybrid and all-electric powertrains are discarded as immediate transitions stages from the MEA. Nonetheless, turboelectric architectures are not dependent on the advancements of energy storage systems and are therefore considered to be the most viable configuration to begin the more electric powertrain aircraft era. By itself, a turboelectric architecture is less efficient than a conventional gas turbine engine due to the lack of additional energy storage devices. However, when implemented with distributed propulsion and boundary-layer ingestion (BLI), it can improve the overall efficiency of the propulsion system, reducing fuel consumption [45].

Distributed electric propulsion (DEP) consists of implementing efficient and high-power density electric motors to drive distributed electric fans that can be strategically placed on the aircraft to achieve a very high effective bypass ratio, which results in higher propulsion efficiency. The onboard electric generators and gas turbines can be placed in different loca-

Table 2.1: Performance matrix for commercial electric aircraft powertrain architectures studies [113, 114]

Aircraft Study	Propulsion	Passengers No.	Speed (Mach)	Electric Power (MW)	Fuel Burn Reduction	TRL* 4-6 Timeline
NASA STARC-ABL	Partial Turboelectric	154	0.80	2 to 3	7-12%	2025
Boeing SUGAR Freeze	Partial Turboelectric (Fuel Cell)	154	0.70	N/A	56%	2030
NASA N3-X	Turboelectric	300	0.84	50	70%	2030
Boeing SUGAR Volt	Parallel Hybrid Electric	154	0.70	1.3 to 5.3	60%	2025
Rolls-Royce	Parallel Hybrid Electric	154	0.70	1 to 2.6	28%	N/A
UTRC (Raytheon)	Parallel Hybrid Electric	154	0.70	2.1	6%	N/A
Airbus/Siemens/Rolls Royce/E-Fan X	Series Hybrid	100	N/A	2	N/A	2030
WrightElectric/EasyJet	All-Electric	120	N/A	0.78	N/A	2027

* TRL: Technology Readiness Level

tions with respect to the fan. Hence, allowing an improved aerodynamic and weight profile for the aircraft. This feature can only be accomplished due to the simplicity of transporting electric energy efficiently onboard. In addition, with BLI, the propulsion efficiency can be further increased by reducing the energy usage to ingest the thin, low-momentum air region at the aircraft structure. Some industrial MEA examples, including the N3-X designed by NASA for 300 passengers with a fully superconducting turboelectric powertrain that can provide up to 70% fuel savings [115], are shown in Table 2.1. Whereas single-aisle aircraft are presented in [116,117] with fuel savings of up to 10%. Lastly, numerous turboelectric aircraft concepts and reviews can be consulted in [118–120]. Among all the consulted designs, the implementation of a successful turboelectric powertrain is tied to high power density electric machines, and smart power converters.

2.2 Three-Phase AC-DC Power Converter Modeling

In this section, the mathematical modeling utilized in this dissertation for the three-phase ac-dc power converter is presented. The modeling is developed for a 2L-VSC topology consisting of six fully-controlled semiconductor-based bidirectional switches consisting of a transistor and anti-parallel diode. Moreover, the models presented in this section are based on the external filter dynamics of the converter and does not take into consideration the switching states for any average modeling [14].

2.2.1 Primitive Coordinates or *abc* Modeling

The schematic diagram of the three-phase ac-dc converter utilized in this dissertation is presented in Fig. 2.7. The input to the converter is a variable-frequency/constant-amplitude voltage, which emulates a wound-field synchronous generator. The output terminals of the variable-frequency/constant-amplitude input source are referred to in this dissertation as the point-of-common-coupling (PCC). The PCC and the converter are interconnected by a grid-interfacing inductive L filter of parameters R_f and L_f , where L_f is the inductance

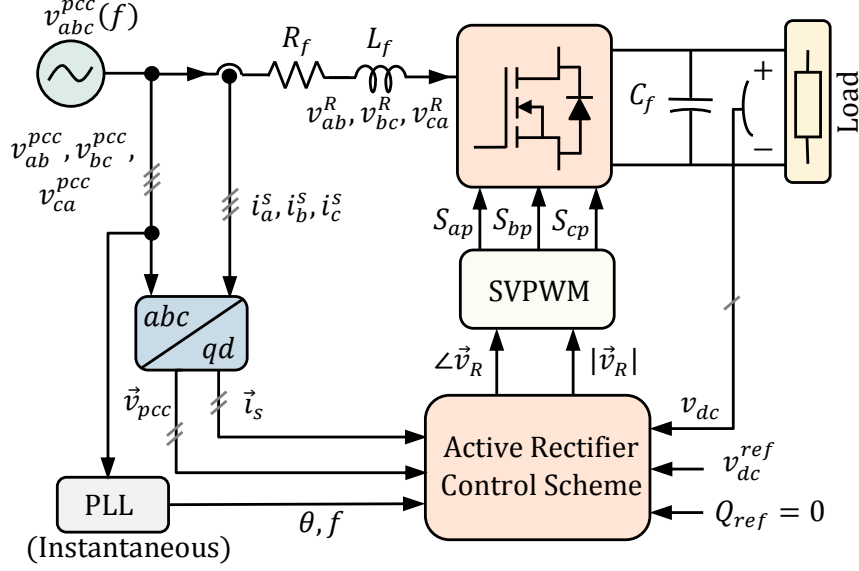


Figure 2.7: Block diagram of the three-phase ac-dc power converter implemented in this dissertation [103].

per-phase of the filter, and R_f is its the per-phase parasitic resistance.

It is assumed that the control schemes developed in this dissertation have access to the current and voltage measurements at the PCC, as well as the dc-bus voltage measurement. The dc-bus or output of the converter is connected to a resistive load and contains a dc-bus filtering electrolytic capacitor C_f .

That being said, the line current dynamic equations in the abc -coordinate system considering only the voltage drop across the inductive filter can be written as

$$\begin{bmatrix} v_a^R \\ v_b^R \\ v_c^R \end{bmatrix} = \begin{bmatrix} v_a^{pcc} \\ v_b^{pcc} \\ v_c^{pcc} \end{bmatrix} - \begin{bmatrix} R_f & 0 & 0 \\ 0 & R_f & 0 \\ 0 & 0 & R_f \end{bmatrix} \begin{bmatrix} i_a^s \\ i_b^s \\ i_c^s \end{bmatrix} - \begin{bmatrix} L_f & 0 & 0 \\ 0 & L_f & 0 \\ 0 & 0 & L_f \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_a^s \\ i_b^s \\ i_c^s \end{bmatrix} \quad (2.1)$$

where, $\begin{bmatrix} v_a^R & v_b^R & v_c^R \end{bmatrix}$ is the phase-to-neutral voltage vector at the terminals of the power converter, $\begin{bmatrix} v_a^{pcc} & v_b^{pcc} & v_c^{pcc} \end{bmatrix}$ is the phase-to-neutral voltage vector at the PCC, and $\begin{bmatrix} i_a^s & i_b^s & i_c^s \end{bmatrix}$ is the line current vector.

Alternatively, (2.1) can be written using line-to-line quantities as

$$\begin{bmatrix} v_{ab}^R \\ v_{bc}^R \\ v_{ca}^R \end{bmatrix} = \begin{bmatrix} v_{ab}^{pcc} \\ v_{bc}^{pcc} \\ v_{ca}^{pcc} \end{bmatrix} - \begin{bmatrix} R_f & 0 & 0 \\ 0 & R_f & 0 \\ 0 & 0 & R_f \end{bmatrix} \begin{bmatrix} i_{ab}^s \\ i_{bc}^s \\ i_{ca}^s \end{bmatrix} - \begin{bmatrix} L_f & 0 & 0 \\ 0 & L_f & 0 \\ 0 & 0 & L_f \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{ab}^s \\ i_{bc}^s \\ i_{ca}^s \end{bmatrix} \quad (2.2)$$

where, $\begin{bmatrix} v_{ab}^R & v_{bc}^R & v_{ca}^R \end{bmatrix}^T$ is the line-to-line voltage vector at the terminals of the active rectifier, $\begin{bmatrix} v_{ab}^{pcc} & v_{bc}^{pcc} & v_{ca}^{pcc} \end{bmatrix}^T$ is the line-to-line voltage vector at the PCC, and $\begin{bmatrix} i_{ab}^s & i_{bc}^s & i_{ca}^s \end{bmatrix}^T$ are the so-called virtual line-to-line currents, i.e., $i_{ab}^s = i_a^s - i_b^s$, $i_{bc}^s = i_b^s - i_c^s$, and $i_{ca}^s = i_c^s - i_a^s$ [14, 121].

2.2.2 dq -Arbitrary Reference Frame Modeling

The systems in (2.1) and (2.2) can be transformed to the $dq0$ reference frame by implementing the

$$\vec{x} = \frac{2}{3} (x_a + \alpha x_b + \alpha^2 x_c) e^{-j\delta} = x_q - jx_d \quad (2.3)$$

space-vector transformation for any given time-varying phase-variable $x(t)$; where $\alpha = e^{j\frac{2\pi}{3}}$, and δ is an arbitrary phase [122]. In particular, if $\delta = 0$, (2.1) can be expressed in the $dq0$ stationary reference frame as

$$\vec{v}_R = \vec{v}_{pcc} - R_f \vec{i}_s - L_f \frac{d\vec{i}_s}{dt}, \quad (2.4)$$

where $\vec{v}_R$ is the rectifier phase-to-neutral voltage space-vector, $\vec{i}_s$ is the line current space-vector, and $\vec{v}_{pcc}$ is the PCC phase-to-neutral voltage space-vector. Conversely, if $\delta = \theta = \omega t$, (2.1) can be transformed to the $dq0$ synchronous reference frame as [14]

$$v_q^R = v_q^{pcc} - R_f i_q - L_f \frac{di_q}{dt} - L_f \omega i_d, \quad (2.5)$$

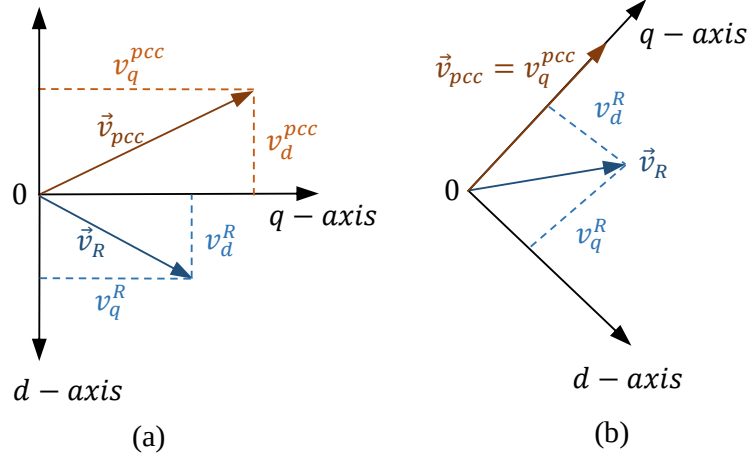


Figure 2.8: Space-vector diagrams representing the PCC and active rectifier voltages in the qd -plane; (a) stationary reference frame, and (b) synchronous reference frame [63, 99].

$$v_d^R = v_d^{pcc} - R_f i_d - L_f \frac{di_d}{dt} + L_f \omega i_q. \quad (2.6)$$

Similarly, (2.2) can be written in the same reference frame as

$$v_q^R = v_q^{pcc} - R_f i_{Lq} - L_f \frac{di_{Lq}}{dt} - L_f \omega i_{Ld}, \quad (2.7)$$

$$v_d^R = v_d^{pcc} - R_f i_{Ld} - L_f \frac{di_{Ld}}{dt} + L_f \omega i_{Lq}. \quad (2.8)$$

Lastly, Fig. 2.8 shows the space-vector diagrams of the transformations used in this dissertation. Notice that in Fig. 2.8(b), the voltage at the PCC is locked to the q -axis [122]. A detailed explanation of the reference frame theory implemented in this dissertation is presented in Appendix A.

2.3 Phase-Angle and Frequency Detection Schemes

Grid-tied and stand-alone power converters are typically equipped with phase-angle and frequency detection schemes as part of their control scheme [14, 63, 121]. In many controllers, the phase-angle is used as input for the dq -synchronous frame of reference transformation,



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A classical procedure to estimate the phase-angle and frequency of a three-phase system consists of implementing a phase-locked loop (PLL) algorithm [123]. The most common PLL structure is formed by two second-order closed-loop transfer functions, which require tuning to adjust the bandwidth of the PLL [125]. More advanced adaptive algorithms [126] are equipped with dynamic parameter tuning algorithms in order to improve the performance of the PLL, while other complex algorithms such as the real-time discrete Fourier transformation (DFT) [124] and steady-state linear Kalman filter [125] exhibit an improvement in the dynamic performance when estimating the phase-angle and frequency at the expense of increasing the computational burden of their implementation.

2.3.1 Classic and DFT-based PLLs

The classic configuration of a PLL, as shown in Fig. 2.9(a), consists of: (i) a phase detector, (ii) a PID control loop with automatic gain control, (iii) a controlled oscillator, and (iv) a low-pass filter. In this PLL scheme, the frequency is assumed constant or quasi-constant, which simplifies the tuning of the controllers. Conversely, Fig. 2.9(b) depicts DFT-based PLL that uses advanced signal processing algorithms to improve the bandwidth of the PLL and its dynamic response. The DFT-based PLL estimates the frequency using a PID control loop that minimizes the error between the estimated and measured DFT frequency. This PLL scheme is investigated for variable-frequency applications in [124,125].

2.3.2 Direct Phase-Angle Detection

In this dissertation, both the benchmark and the developed control schemes use a direct phase-angle technique, as developed in [121,127]. The direct phase-angle detection (DPD) technique, applicable only to three-phase systems, utilizes two of the three line-to-line voltage setting one as the reference. For balanced three-phase system represented as

$$\begin{cases} v_{ab} = V_{LL} \cos(\theta_{ab}) \\ v_{bc} = V_{LL} \cos\left(\theta_{ab} - \frac{2\pi}{3}\right) \\ v_{ca} = V_{LL} \cos\left(\theta_{ab} - \frac{4\pi}{3}\right) \end{cases}, \quad (2.9)$$

where V_{LL} is the line-to-line rms value, and θ_{ab} is the phase-angles of the line-to-line voltages v_{ab} . This phase-angle, initially assumed unknown, can be directly estimated by first expanding the expression for v_{bc} in (2.9) as

$$v_{bc} = -\frac{1}{2}V_{LL} \cos(\theta_{ab}) + \frac{\sqrt{3}}{2}V_{LL} \sin(\theta_{ab}). \quad (2.10)$$

Re-arranging (2.10) and dividing both sides of the equation by $v_{ab} = V_{LL} \cos(\theta_{ab})$ leads to:

$$\frac{v_{bc}}{v_{ab}} = -\frac{1}{2} + \frac{\sqrt{3}}{2} \tan(\theta_{ab}). \quad (2.11)$$

From (2.11), $\theta_{ab} \in [0, 2\pi]$ can be computed as:

$$\theta_{ab} = \tan^{-1} \left[\frac{v_{ab} + 2v_{bc}}{\sqrt{3}v_{ab}} \right]. \quad (2.12)$$

For its digital implementation, the phase-angle estimation computed with (2.12) employs a four-quadrant inverse tangent (atan2), which outputs angles from $[-\pi, \pi]$. As a result, (2.12) is programmed as $\theta_{ab} = \text{atan2}(v_{ab} + 2v_{bc}, \sqrt{3}v_{ab})$. An additional, “if” condition is added to the algorithm to sum 2π when $\theta_{ab} < 0$. Fig. 2.10 provides a graphical visualization of the DPD method for detecting θ_{ab} [127].

Once the phase-angle is computed, the frequency f can be estimated by taking the time derivative of (2.12). Fig. 2.9(c) shows the block diagram utilized to estimate the frequency. Herein, a first-order low-pass filter $H_f(s) = 1/(1 + s/\omega_c)$ with a cut-off angular frequency of ω_c is used as part of the time derivative transfer function block to attenuate the high-frequency noise present in the voltage measurements. In addition, a rate limiter block is implemented to suppress the undesired frequency spikes produced by the discontinuous nature of the θ_{ab} signal.

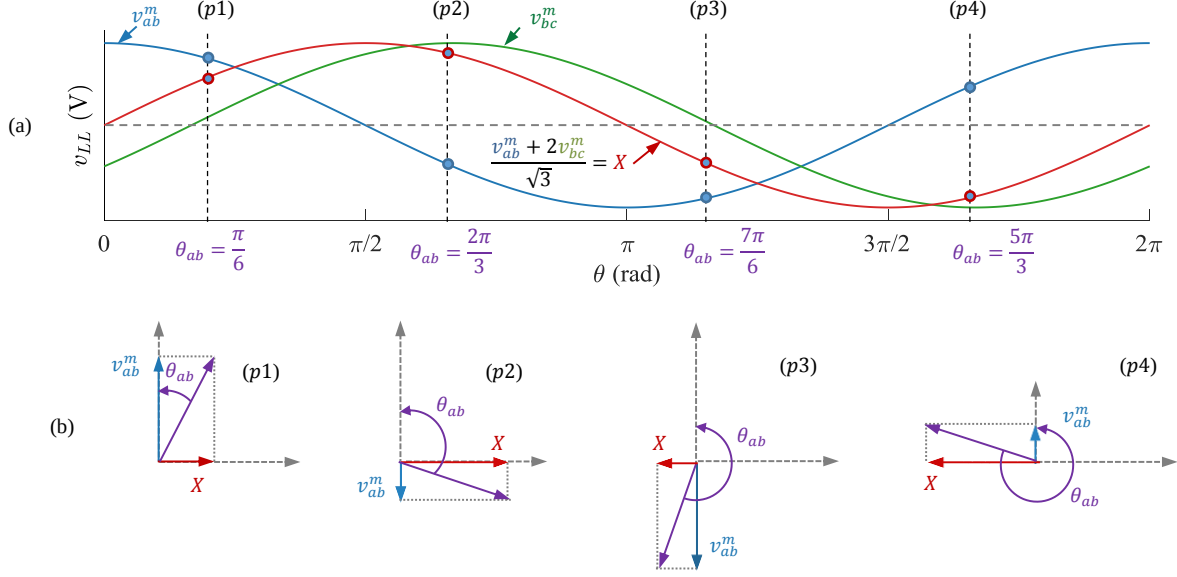


Figure 2.10: Graphical visualization of the four-quadrant inverse tangent formula used for DPD method. (a) Three-phase signal, (b) Phase-angle, θ_{ab} ranging as (p1) $0 \leq \theta_{ab} < \pi/2$, (p2) $\pi/2 \leq \theta_{ab} < \pi$, (p3) $\pi \leq \theta_{ab} < 3\pi/2$, and (p4) $3\pi/2 \leq \theta_{ab} < 2\pi$ [127].

2.3.3 Direct-Phase Angle Detection and Classic PLL Comparison

To validate the effectiveness of the phase-angle detection compared to classic PLL in terms of frequency tracking, a Matlab/Simulink simulation was performed using acceleration and deceleration frequency ramps from 360 Hz to 800 Hz and vice-versa. The results are presented in Fig. 2.11, which shows the advantages of implementing the direct phase-angle detection scheme as opposed to a classic PI-based PLL.

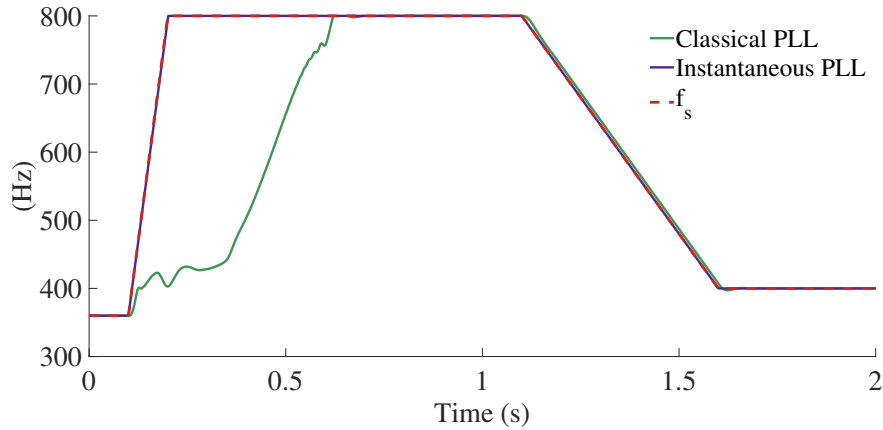


Figure 2.11: Simulated PLL comparison for a variable-frequency profile [100].

Given its PI-based structure, the classic PLL can be re-tuned to improve its performance based on the frequency ramp slopes. However, as with any PI controller, the tuning is unique to a range of frequency variations, and the performance of the PLL cannot be guaranteed for any frequency change. Conversely, the direct phase-angle detection algorithm does not require any parameter tuning aside from its low-pass filter. Furthermore, its computational burden is lower due to its reduced number of integrators compared to other PLLs.

2.4 Classic Model-based dq –Control Scheme

The model-based dq –control scheme is a derivation of the electric machine vector control scheme that uses the model an RL branch to regulate the current. Since then, the model-based dq –controller has been extensively implemented by researchers and engineers in three-phase ac-dc power converters [99, 128–130]. The main objective of the model-based dq –control scheme is to regulate the dc-bus and control the reactive power flow in the ac-side. Typically, the reactive reference power is set to zero to operate the system at unity power factor efficiently.

That being said, the model-based dq –control scheme is formulated in the $dq0$ synchronous reference frame. Herein, the active P and reactive Q power at the PCC are defined as

$$P = \frac{3}{2} \left(v_d^{pcc} i_d + v_q^{pcc} i_q \right); \quad (2.13)$$

$$Q = \frac{3}{2} \left(v_q^{pcc} i_d - v_d^{pcc} i_q \right). \quad (2.14)$$

Once the q –axis is locked to the a –axis or ab –axis of the three-phase system, the d –component of the PCC voltage is null, i.e., $v_d^{pcc} = 0$. As a result, (2.13) and (2.14) are simplified to $P = 3v_q^{pcc}i_q/2$ and $Q = 3v_q^{pcc}i_d/2$. This formulation allows the active power to be controlled with i_q and the reactive power with i_d . Furthermore, since the dc-bus voltage v_{dc} is proportional to P , v_{dc} can also be regulated with i_q . Consequently, the dc-bus and reactive power can be controlled using two decoupled PI-controllers, as depicted in Fig.

2.5 Summary

In summary, this chapter covered the system description of a more electric powertrain. Starting with hybrid and all-electric powertrains for vehicle and aircraft applications, followed by three-phase active rectifier modeling and phase-angle and frequency detection schemes; finalizing with the classic model-based dq –controller. Based on the modeling concepts articulated in this chapter, the following chapter presented the three-phase ac-dc power converter controller schemes developed in this dissertation for more electric powertrains.

Chapter 3

Smart Power Converter Control Schemes

for More Electric Powertrains

More electric powertrains exhibit rapid variations in the input and output variables of the system. Furthermore, power converters, working as part of a more electric powertrain, experience important temperature and pressure fluctuations, which translate in the variation of the parameters of the system.

Both input/output-variable and parameter variations impose new challenges for power converter controllers. Consequently, this chapter presents the formulation of three novel control schemes for smart power converters applicable to more electric powertrains. The controllers are designed to regulate the dc-bus voltage and reactive power of the powertrain electric generator ac-dc converter. Firstly, a model predictive controller (MPC) with a direct frequency tracking algorithm is presented [63,99,100,104,131]. Secondly, an online adaptive parameter estimation technique is added to the previously mentioned controller to address performance degradation due to parameter mismatch [101,102]. Lastly, a seamless dynamics direct model reference adaptive (MRAC) control scheme is presented [103].

3.1 Step-Ahead Model Predictive Control Scheme

The main objective of the developed step-ahead model predictive control scheme is to regulate the dc-bus voltage and power factor at the ac mains while maintaining a low THD in the generator-side currents. Consequently, the two control variables are v_{dc} and the reactive power Q .

3.1.1 Formulation

The developed step-ahead model predictive control scheme [99, 100], as shown in Fig. 3.1, can be categorized as a fixed switching frequency deadbeat controller [132–135]. The control scheme is designed in the discrete-time domain. Herein, two line-to-line voltages at the PCC, v_{ab}^{pcc} and v_{bc}^{pcc} , and three line currents, i_a^s , i_b^s , and i_c^s , are sampled at a time step T_s and transformed to the $dq0$ stationary frame of reference using the transformation depicted in Appendix A. Once in the $dq0$ stationary frame of reference, (2.4) can be discretized by implementing the Forward Euler method for every instant k as

$$\vec{v}_R[k] = \vec{v}_{pcc}[k] - \left(R_f - \frac{L_f}{T_s}\right) \vec{i}_s[k] - \frac{L_f}{T_s} \vec{i}_s[k+1]. \quad (3.1)$$

Herein, the rectifier voltage space-vector $\vec{v}_R[k]$ is a function of $\vec{v}_{pcc}[k]$ and $\vec{i}_s[k]$ (measured at the PCC), the parameters of the inductive filter, i.e., R_f and L_f , and the line current space-vector at the next control cycle, $\vec{i}_s[k+1]$. Moreover, $\vec{i}_s[k+1]$ can be related to the instantaneous active, P , and reactive power, Q , at instant $k+1$ by:

$$P + jQ = (3/2) \vec{v}_{pcc}[k+1] \left(\vec{i}_s[k+1]\right)^*, \quad (3.2)$$

where, $P = P_{ref} = g(v_{dc}^{ref}, v_{dc})$ is the amount of watts the rectifier must set to regulate the dc-bus to v_{dc}^{ref} . This quantity is determined by a PI-loop that minimizes the steady-state dc-bus error, as shown in Fig. 3.1. In contrast, to minimize the copper losses on the ac-side, the reactive power is set to zero, i.e., $Q_{ref} = 0$, which translates into a unity power factor. Consequently, $\vec{i}_s[k+1]$ can be isolated from (3.2) resulting in

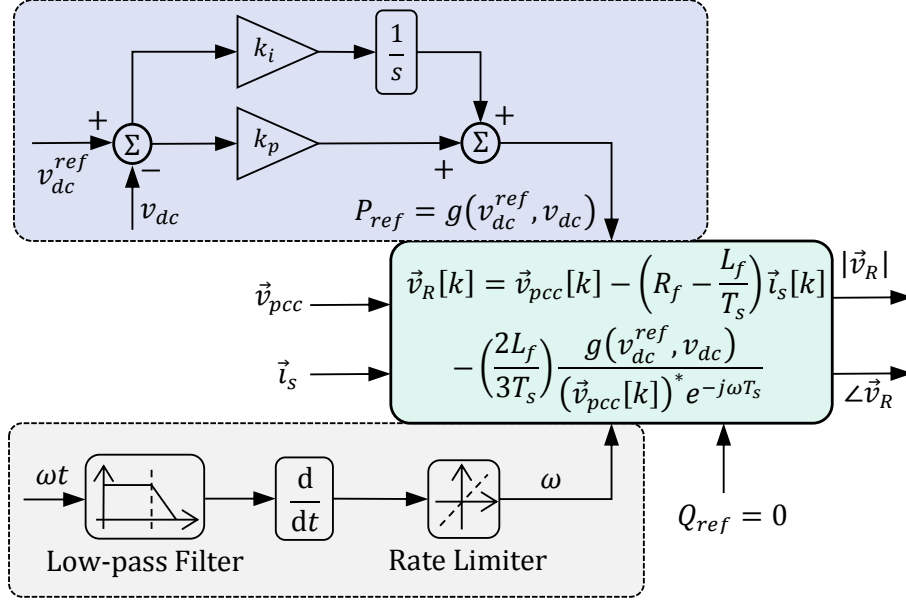


Figure 3.1: Block diagram of the step-ahead control scheme [63].

$$\vec{i}_s[k+1] = \left(\frac{2}{3}\right) \frac{g(v_{dc}^{ref}, v_{dc})}{(\vec{v}_{pcc}[k+1])^*}. \quad (3.3)$$

Herein, $\vec{v}_{pcc}[k+1]$ can be computed for a three-phase voltage supply by shifting $\vec{v}_{pcc}[k]$ by ωT_s radians:

$$\vec{v}_{pcc}[k+1] = \vec{v}_{pcc}[k] e^{j\omega T_s}, \quad (3.4)$$

where, $\omega = 2\pi f$ is the angular frequency at the PCC and it is estimated by the direct phase-angle detection algorithm described in Section 2.3 and shown in Fig. 2.9(c). Finally, by replacing (3.3) and (3.4) into (3.1), the targeted rectifier voltage $\vec{v}_R[k]$, which is required to control the dc-bus voltage at unity power factor, can be computed at each sampling time, k , as

$$\vec{v}_R[k] = \vec{v}_{pcc}[k] - \left(R_f - \frac{L_f}{T_s}\right) \vec{i}_s[k] - \left(\frac{2L_f}{3T_s}\right) \frac{g(v_{dc}^{ref}, v_{dc})}{(\vec{v}_{pcc}[k])^* e^{-j\omega T_s}}. \quad (3.5)$$

Once the targeted space-vector is estimated using (3.5), $\vec{v}_R[k]$ is normalized by v_{dc} and used to form SVPWM switching pattern. Lastly, although given the high-switching frequency

implemented in this work is enough to achieve low current THDs, Appendix B describes the formulation and implementation of the step-ahead controller in power control mode for an *LCL* filter [131].

3.1.2 Delay Compensation

To implement the step-ahead control scheme in a hardware microprocessor unit, the processing time of the algorithm and updating PWM reference must be taken into account. Under this context, the rectifier voltage $\vec{v}_R[k]$ computed by (3.5) will be available for the PWM module only after the calculations of the controller and all the measurement A/D conversions are completed. Fig. 3.2 illustrates the operation of the step-ahead controller and the computation time in comparison to the sampling time. Herein, if the $\vec{v}_R[k]$ computation takes longer than half of the sampling time, the duty-cycles generated by the SVPWM scheme will not be updated until the next sampling time. As a result, $\vec{v}_R[k]$ will be applied at $k+1$. To address this inherent computational delay, (3.5) is shifted one step forward [136]; thus generating a step delay between the controller's inputs and its implementation [137]. Moreover, when applying the control action at $k+1$, the control variables are expected to reach their set-points values approximately at $k+2$. However, in most cases, the control variables will not be regulated exactly at $k+2$, as illustrated in Fig. 3.2. It may take one or two more control time steps to reach their set-points depending on the gains of the controller [99].

Lastly, the main steps taken by the step-ahead control technique can be summarized as follows [99]:

1. Apply $\vec{v}_R[k]$ from the previous step, $k-1$, to the PWM generator module.
2. Measure the voltage and current values: v_{dc} , v_{ab}^{pcc} , v_{bc}^{pcc} , i_a^s , i_b^s , and i_c^s .
3. Compute $g(v_{dc}^{ref}, v_{dc})$ from PI control loop, and the frequency f from the direct-phase angle detection algorithm.
4. Compute $\vec{v}_R[k]$ from (3.5) and apply it in the next step control step, i.e., $k+1$.

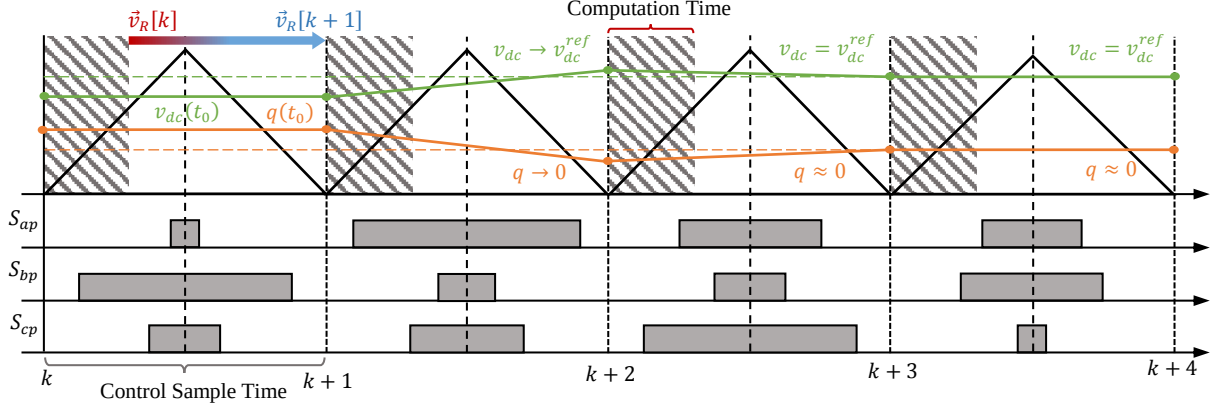


Figure 3.2: Operation of the step-ahead control scheme from k to $k+4$ [99].

3.1.3 Variable-Frequency Steady-State Error

The performance of an MPC scheme relies heavily on the correct prediction of the control variables at the next time step to compute the switching states of the converter. In model-based controllers, the predicted variables are known to be dependent on the accuracy of the parameters of the model, e.g., grid-interfacing filter parameters. This principle is known as parameter mismatch or model uncertainty, and it has been investigated in multiple reports [75, 138]. Nonetheless, in all of the previous works, the input frequency of the converter is considered constant and has not been included in the model uncertainty analysis. In fact, in the majority of the algorithms, the frequency is not measured since the voltage at the PCC is assumed constant from one control step to the other, i.e., $\vec{v}_{pcc}[k+1] \approx \vec{v}_{pcc}[k]$.

This subsection investigates the effects of frequency mismatch between the actual system and the step-ahead controller. As mentioned in the previous subsection, the control variables tend to reach their set-points at $k+2$, for a control action applied at $k+1$, based on data measured at k . Under this premise, the error in frequency measurement appears mainly on the reactive power. Notice that the dc-bus voltage is regulated by a PI-loop that minimizes the steady-state error of v_{dc} with its integral component. Accordingly, the reactive power tracking error at $k+2$, $\Delta Q[k+2]$, can be accounted as follows:

$$(3/2) \vec{v}_{pcc}[k+2] \left(\vec{i}_s[k+2] \right)^* = g(v_{dc}^{ref}, v_{dc}) \pm j \Delta Q[k+2], \quad (3.6)$$

where $\vec{v}_{pcc}[k+2] = \vec{v}_{pcc}[k] e^{j2\omega T_s}$, and $\omega = 2\pi f$. Based on (3.6), the reactive power tracking error $\Delta Q[k+2]$ when the frequency f is inaccurately estimated can be derived using the Error Propagation formula as:

$$\Delta Q[k+2] = \left| \frac{\partial Q[k+2]}{\partial f} \right| \Delta f, \quad (3.7)$$

where $\Delta f = \Delta\omega/2\pi$ is the difference between the frequency used in the step-ahead algorithm and the actual frequency of the system. In addition, the reactive power at $k+2$ is:

$$Q[k+2] = \Im m \left\{ (3/2) \vec{v}_{pcc}[k] e^{j2\omega T_s} (\vec{i}_s[k+2]) \right\} \quad (3.8)$$

According to (3.6) and (3.7), the frequency error is proportional to the reactive power tracking error. Moreover, these equations proof that for the step-ahead algorithm, assuming $\vec{v}_{pcc}[k+2] \approx \vec{v}_{pcc}[k]$ is only valid for low- and fixed-frequency applications, which is not the case in more electric powertrains. Consequently, the main contribution of the step-ahead controller resides in its integration with the direct phase-angle detection, which enables fast and accurate tracking of the frequency, thus minimizing the tracking error in the reactive power.

3.2 Indirect Adaptive Step-Ahead Control Scheme

Model predictive control schemes are known to suffer from performance degradation under model parameter mismatch or model uncertainty [75, 76], which can be an important factor in variable-frequency applications, e.g., more electric powertrains, since the parameters of the system may change with the frequency [77–80, 102]. Furthermore, due to temperature and pressure variations in electric/hybrid vehicles or aircraft, the passive filter properties can be affected. For the specific case of an inductive L filter, as shown in Fig. 2.7, the inductance and parasitic resistance of the filter can change their values depending on several external operating conditions.

As demonstrated in (3.5), the step-ahead control scheme requires the accurate values of R_f and L_f to correctly select the switching duty ratios. Consequently, in this section, an adaptive filter parameter estimation algorithm is developed to address the performance degradation suffered by the step-ahead controller under system parameter variations [101, 102]. The addition of the aforementioned estimation technique enables the power converter to dynamically adapt to external changes that may occur during the vehicle operation, thus introducing for the first time, in this work, the concept of a *smart* power converter.

3.2.1 Adaptive Filter Parameter Estimation

The developed adaptive parameter estimation technique implements a first-order parallel-series system identification scheme [139], see Fig. 3.3(a), to estimate the parameters of the generator-side L filter, i.e., R_f and L_f . Based on Fig. 3.3(a), the output of a first-order system with unknown parameters, a and b , defined as

$$\dot{x} = ax + bu, \quad (3.9)$$

is compared to the output of a reference model with adjustable parameters, a_m and b_m , defined as

$$\dot{x}_m = a_mx_m + \beta e + b_mu; \quad (3.10)$$

where $e = x - x_m$ is the state variable error, and $\beta > 0 \in \mathbb{R}$ is the error weight factor.

For a balanced system, the parameter estimation technique can be applied for any of the phases in (2.1). In particular, for phase a , the first row of (2.1) can be expressed in a state-space notation, i.e., $\dot{x} = ax + bu$, as

$$\frac{di_a}{dt} = (-R_f/L_f)i_a + (1/L_f)(v_a^{pcc} - v_a^R); \quad (3.11)$$

where the input $u = v_a^{pcc} - v_a^R$ is the voltage drop on the L filter, and the state variable $x = i_a$ is the line current. Both v_a^{pcc} and i_a are measured at the PCC by the step-ahead

controller. Whereas, v_a^R can be estimated with (3.12) using the duty cycle signals outputted by the step-ahead algorithm, i.e., S_{ap} , S_{bp} , and S_{cp} , in conjunction with v_{dc} [140]. Notice that this formulation can be extended to multi-phase systems with the proper sensor count, whereas the use of ancillary services can compensated asymmetries in the 2L-VSC [141].

$$\begin{bmatrix} v_a^R \\ v_b^R \\ v_c^R \end{bmatrix} = \left(\frac{v_{dc}}{3\sqrt{3}} \right) \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_{ap} \\ S_{bp} \\ S_{cp} \end{bmatrix}. \quad (3.12)$$

From (3.11), the unknown system parameters are $a = -R_f/L_f$ and $b = 1/L_f$. Similarly, the adjustable parameters of the reference model are $a_m = -R_{f0}/L_{f0}$ and $b_m = 1/L_{f0}$, where R_{f0} and L_{f0} are the parameters to be estimated by the adaptive algorithm. Using the previously described notation, the parameter errors Δa and Δb are defined as follows

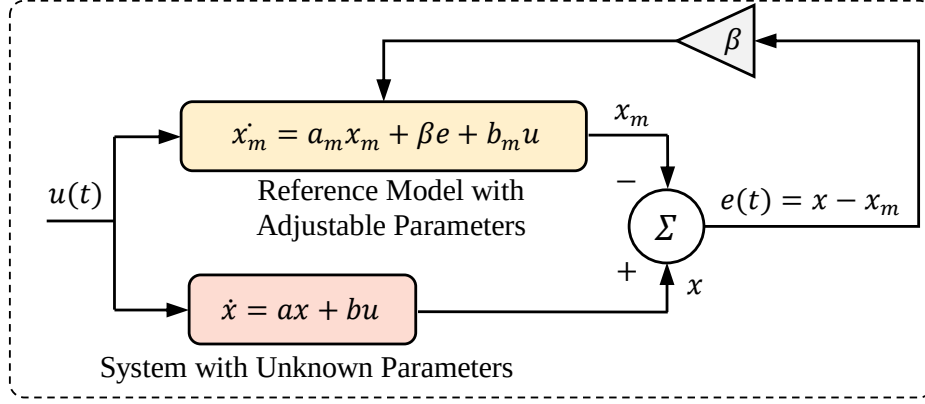
$$\begin{aligned} \Delta a &= a - a_m = (-R_f/L_f) - (-R_{f0}/L_{f0}), \\ \Delta b &= b - b_m = (1/L_f) - (1/L_{f0}). \end{aligned} \quad (3.13)$$

To estimate the desired parameters, the errors Δa and Δb must be minimized. Furthermore, the state error $e = x - x_m$ must also be minimized to guarantee that the system and reference model match. For such purposes, a positive definite Lyapunov function V defined as

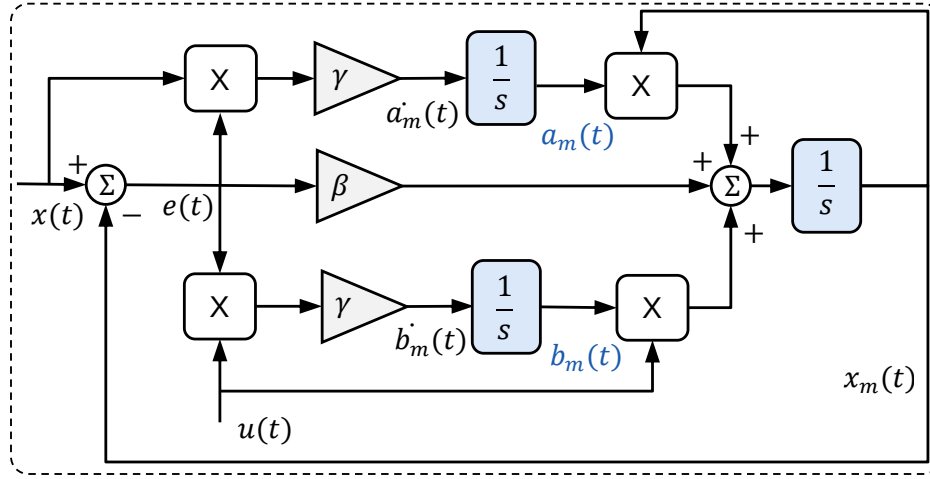
$$V = \frac{1}{2} \left(e^2 + \frac{1}{\gamma} \Delta a^2 + \frac{1}{\gamma} \Delta b^2 \right), \quad (3.14)$$

can be used. The convergence of a_m and b_m to the parameters of the system a and b can be achieved by formulating a set of adaptation laws that force $dV/dt \leq 0$ for $t \rightarrow \infty$. With this in mind, the time derivative of V can be obtained by taking the derivative of (3.14), which results in

$$\frac{dV}{dt} = e \frac{de}{dt} + \frac{1}{\gamma} \Delta a \frac{d\Delta a}{dt} + \frac{1}{\gamma} \Delta b \frac{d\Delta b}{dt}. \quad (3.15)$$



(a)



(b)

Figure 3.3: Parallel-series system identification technique; (a) general block diagram, and (b) detailed block diagram [101, 102].

In (3.15), the time derivative of the state error e is

$$\begin{aligned} \frac{de}{dt} &= (a_m - \beta)(x - x_m) + (a - a_m)x + (b - b_m)u \\ &= (a_m - \beta)e + \Delta a x + \Delta b u. \end{aligned} \quad (3.16)$$

Substituting (3.16) in (3.15) results in

$$\frac{dV}{dt} = (a_m - \beta)e^2 + \Delta a \left(e x + \frac{1}{\gamma} \frac{d\Delta a}{dt} \right) + \Delta b \left(e u + \frac{1}{\gamma} \frac{d\Delta b}{dt} \right). \quad (3.17)$$

Based (3.17), $dV/dt \leq 0$ when $a_m < 0$, $\gamma > 0 \in \mathbb{R}$, $d\Delta a/dt = -\gamma e x$, and $d\Delta b/dt = -\gamma e u$;

which translates into the following adaptation laws:

$$\begin{aligned} a_m &= \int \gamma e(t) x(t) dt \\ b_m &= \int \gamma e(t) u(t) dt \end{aligned} \quad (3.18)$$

where $\gamma > 0 \in \mathbb{R}$ is the adaptation coefficient. Notice that if $dV/dt \leq 0$, the Lyapunov function V is bounded, which results in Δa , Δb , and $e = x - x_m$ being bounded as well. Moreover, since the reference model, $\dot{x}_m = a_m x_m + \beta e + b_m u$, is stable and bounded ($a_m < 0$), then $x = e + x_m$ is bounded. As a result, the convergence of (3.18) intrinsically guarantees the stability of the system by ensuring that the state variable $x = i_a$ is bounded and $e \rightarrow 0$ for $t \rightarrow \infty$ [142]. Finally, a detailed block diagram of the developed adaptive parameter estimation scheme is depicted in Fig. 3.3(b) [101, 102].

3.2.2 Controller Design

The general procedure to design the step-ahead controller in conjunction with the developed adaptive parameter estimation technique is provided in this subsection:

1. **DC-Bus PI controller design:** The proportional and integral gains, i.e., k_p and k_i , of the v_{dc} PI control loop, see Fig. 3.1, are set to achieve a settling time of 1s and 10% voltage overshoot for a 50% load step change at 100 Hz. Consequently, $k_p = 15$ and $k_i = 10$ at 1 kW base power, as shown in Fig. 3.4.
2. **Adaptive parameter estimation gains (γ and β):** The adaptation coefficient γ and error weight factor β directly affect the dynamic performance of estimated filter parameters. Fig. 3.5(a) shows the estimation of the filter inductance L_{f0} for different γ set-points. Higher values of γ reduce the settling time, with the trade-off of increasing the overshoot. In this work, γ is set to achieve a settling time of 20ms and 5% overshoot for a 1 mH inductance step increase at 100 Hz. Based on the aforementioned criteria, γ is set to 10^4 ; which produces a rapid reactive power tracking, as depicted in Fig. 3.5(b). Lastly, the error weight factor β is set to ten times less than γ , i.e., $\beta = 10^3$.

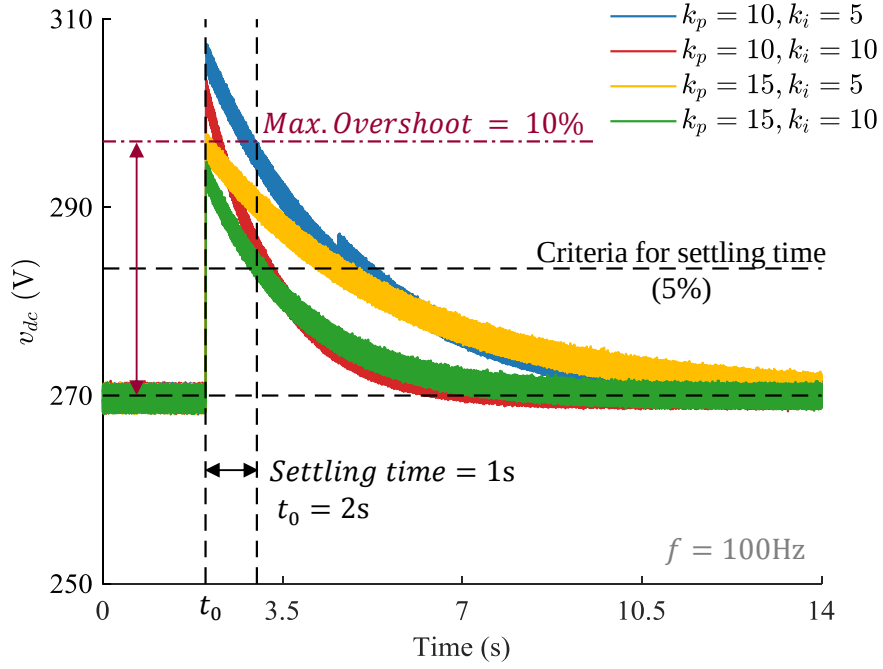


Figure 3.4: Experimentally obtained transient response of the dc-bus voltage for a 50% load step decrease for several k_p and k_i set-points.

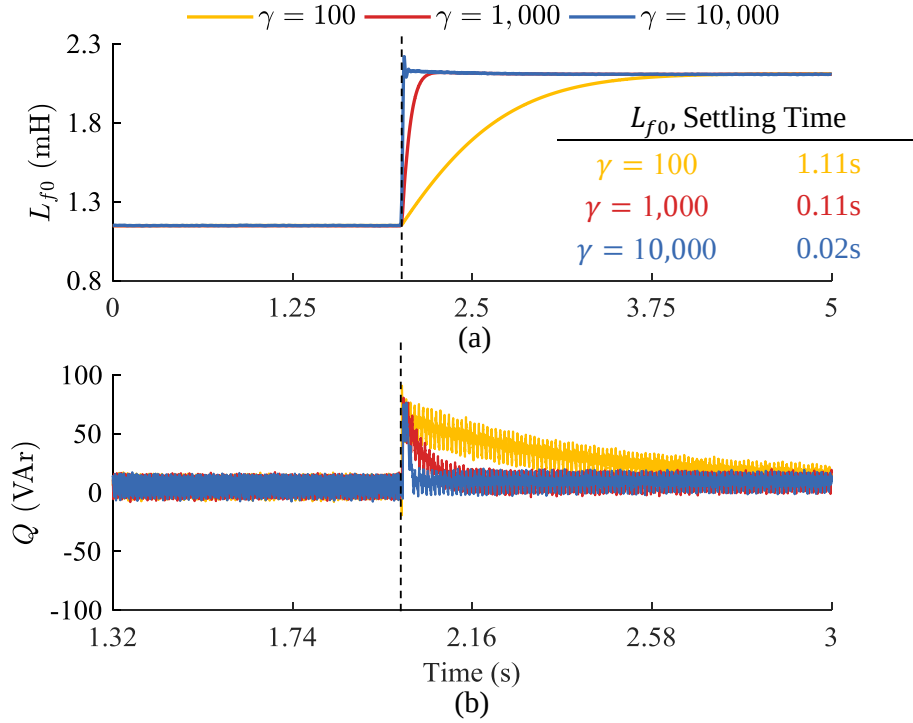


Figure 3.5: Experimentally obtained adaptation coefficient γ at 100 Hz based on the performance of the system; (a) estimated filter inductance L_{f0} , and (b) reactive power transient [101,102].

3.3 Seamless Dynamics Direct Model Reference

Adaptive Control (MRAC) Scheme

In this section, the seamless dynamics direct model reference adaptive control (MRAC) scheme for more electric powertrain smart converters is developed [103]. The controller is designed in the $dq0$ synchronous reference frame with (2.7)-(2.8) as the principal modeling equations. The objectives of this controller are to regulate the dc-bus voltage to its desired set-point while controlling the reactive power flow on the ac-side of the converter. In addition, it is expected that by implementing the control scheme, the ac-side currents exhibit a low THD in the desired load range.

The detailed block diagram of the developed MRAC scheme is shown in Fig. 3.6. Each subsection of this section details the modules or subsystems of the developed MRAC scheme. First, the dc-bus and reactive power PI-control loops are explained based on the active and reactive power decoupled control strategy in the $dq0$ synchronous reference frame. Second, the inner current control loops are developed based on the Command Generator Tracker Concept (CGT) [98]. The CGT provides an adaptive control law whose gains automatically updated using a specific set of adaptation laws. Lastly, the stability of the developed MRAC algorithm is investigated using a Lyapunov approach that involves: (i) finding a positive definite Lyapunov candidate V , and (ii) evaluating the closed-loop stability of the system by analyzing the sign of the time derivative of V , i.e., $dV/dt = \dot{V}$ [103].

3.3.1 DC-bus and Reactive Power Outer PI Control Loops

The developed MRAC scheme is based on (2.7) and (2.8) in the $dq0$ synchronous reference frame. Herein, the model of the system can be written in a state-space representation as:

$$\begin{aligned} \dot{x}(t) &= Ax(t) + Bu(t) \\ y(t) &= Cx(t) \end{aligned}, \tag{3.19}$$

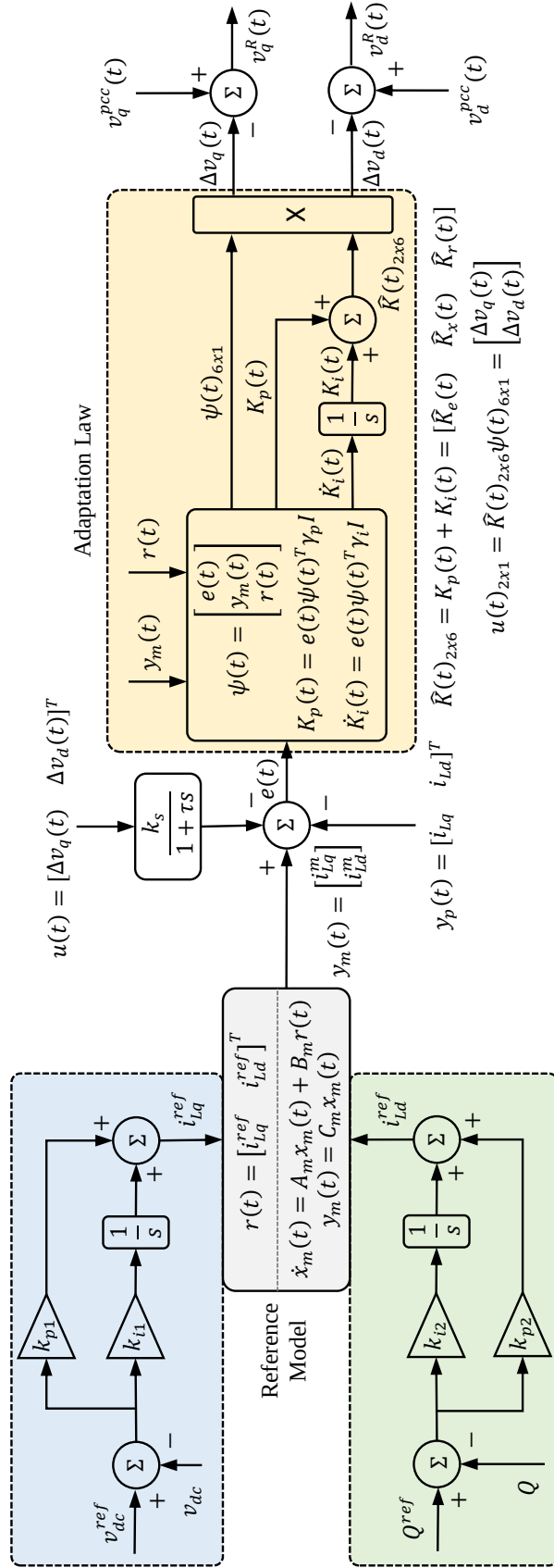


Figure 3.6: Detailed block diagram of the developed MRAC for more electric powertrain smart converters [103].

where $x(t) = \begin{bmatrix} i_{Lq} & i_{Ld} \end{bmatrix}^T$ is the state vector, $u(t) = \begin{bmatrix} \Delta v_q & \Delta v_d \end{bmatrix}^T$ is the input vector, and $y(t) = x(t)$ is the output vector. In particular, the input vector $u(t)$ is defined using the dq components of the voltage drop in the grid-interfacing inductive filter as

$$u(t) = \begin{bmatrix} \Delta v_q \\ \Delta v_d \end{bmatrix} = \begin{bmatrix} v_q^{pcc} - v_q^R \\ v_d^{pcc} - v_d^R \end{bmatrix}. \quad (3.20)$$

Furthermore, the state and output matrices are:

$$A = \begin{bmatrix} -\frac{R_f}{L_f} & -\omega \\ \omega & -\frac{R_f}{L_f} \end{bmatrix}, \quad B = \begin{bmatrix} \frac{1}{L_f} & 0 \\ 0 & \frac{1}{L_f} \end{bmatrix}. \quad (3.21)$$

The developed MRAC technique can only be designed to control the currents or states of the system (i_{Lq} and i_{Ld}), see (3.19). However, since $P = v_q^{pcc}i_{Lq}/2$ and $Q = v_q^{pcc}i_{Ld}/2$ (when $v_d^{pcc} = 0$), the active power or dc-bus voltage can be regulated with i_{Lq} , whereas the reactive power can be controlled with i_{Ld} . With this in mind, two PI-control loops are implemented to minimize the v_{dc} and Q errors. The outputs of these control loops are the dq reference currents, i_{Lq}^{ref} and i_{Ld}^{ref} , required to achieve the desired v_{dc} and Q set-points. Subsequently, i_{Lq}^{ref} and i_{Ld}^{ref} can serve as the input reference vector $r(t) = \begin{bmatrix} i_{Lq}^{ref} & i_{Ld}^{ref} \end{bmatrix}^T$ to the MRAC algorithm, which acts as the inner current control loop, as shown in Fig. 3.6.

3.3.2 Adaptive Current Control Loop

The main objective of the adaptive current control is to regulate the currents i_{Lq} and i_{Ld} such that they match the reference currents i_{Lq}^{ref} and i_{Ld}^{ref} . More specifically, the dynamic response of i_{Lq} and i_{Ld} should match the response of a reference model whose inputs are i_{Lq}^{ref} and i_{Ld}^{ref} , i.e., $r(t) = \begin{bmatrix} i_{Lq}^{ref} & i_{Ld}^{ref} \end{bmatrix}^T$. The reference model is designed as a second-order system such that a step input to the model would cause its outputs to respond with a specific overshoot, rise time, and settling time. Typically, the reference model is designed to have a critically damped response. Notice that the reference model can be arbitrarily designed, and no information on the parameters of the system is required in its design. In contrast, a

specific control law is directly adjusted to minimize the error between the system and model outputs. For a reference model defined as

$$\begin{aligned}\dot{x}_m(t) &= A_m x_m(t) + B_m r(t) \\ y_m(t) &= C_m x_m(t)\end{aligned}, \quad (3.22)$$

where $y_m(t) = x_m(t) = \begin{bmatrix} i_{Lq}^m & i_{Ld}^m \end{bmatrix}^T$ is the desired response for $x(t) = \begin{bmatrix} i_{Lq} & i_{Ld} \end{bmatrix}^T$ to the current reference set-point command $r(t)$. Then the adaptive control law based on the CGT approach is given as [98]

$$u(t) = \hat{K}_e(t)e(t) + \hat{K}_x(t)y_m(t) + \hat{K}_r(t)r(t) \quad (3.23)$$

where $e(t) = y_m(t) - y(t)$ is the output tracking error and $\hat{K}_e(t)$, $\hat{K}_x(t)$, and $\hat{K}_r(t)$ are the adaptive gains. These gains can be written in a vector as

$$\hat{K}(t) = \begin{bmatrix} \hat{K}_e(t) & \hat{K}_x(t) & \hat{K}_r(t) \end{bmatrix}. \quad (3.24)$$

Similarly, $e(t)$, $y_m(t)$, and $r(t)$ can be expressed in vector form as

$$\psi(t) = \begin{bmatrix} e(t) \\ y_m(t) \\ r(t) \end{bmatrix}, \quad (3.25)$$

which allows the control law (3.23) to be written in a compact form as

$$u(t) = \hat{K}(t)\psi(t). \quad (3.26)$$

Herein, the vector gain $\hat{K}(t)$ is defined as the sum of a proportional gain $K_p(t)$ and an integral gain $K_i(t)$:

$$\hat{K}(t) = K_p(t) + K_i(t), \quad (3.27)$$

Each gain is computed using the following adaptations laws:

$$\begin{aligned} K_p(t) &= e(t)\psi(t)^T \bar{T} \\ K_i(t) &= \int e(t)\psi(t)^T T dt \end{aligned} \quad , \quad (3.28)$$

where $\bar{T} \geq 0$ and $T > 0$ are positive definite and positive semi-definite adaptation coefficient matrices. For simplicity, $\bar{T}$ and T can be set to $\bar{T} = \gamma_p I$ and $T = \gamma_i I$, where I is the identity matrix. Following this train of thought, $\gamma_p \geq 0$ and $\gamma_i > 0$ can be conceptually seen as adaptation coefficients, whose tuning process is indicated in Section 3.3.4. Finally, the sufficient conditions for asymptotic tracking are [103]:

- i. The parameters of the system (3.22) are assumed to be bounded
- ii. All possible pairs of A and B are controllable
- iii. All possible pairs of A and C are observable
- iv. The reference model is assumed to be bounded-input/bounded-state stable
- v. There exists a solution for the CGT problem
- vi. The system is almost strictly positive real (ASPR) and a positive definite constant gain matrix exists such that the closed-loop transfer function of the system is strictly positive real (SPR) [98].

For the specific case of the system presented in Fig. 2.7; condition (i) is satisfied since all the parameters of the system are positive and bounded, given their physical properties. Conditions (ii) and (iii) are also satisfied as the system can be designed with the appropriate number of sensors and outputs such that it is controllable and observable. Furthermore, (iv) can be accomplished by selecting a stable system as a reference model. Moreover, (v) is guaranteed by the existence of matrices $B^{-1}(A_m - A)$ and $B^{-1}B_m$ [98]. Finally, since the SPR condition can only be satisfied under certain parameter variation scenarios, (vi) is satisfied by incorporating an augmented plant structure with feedforward compensator, as

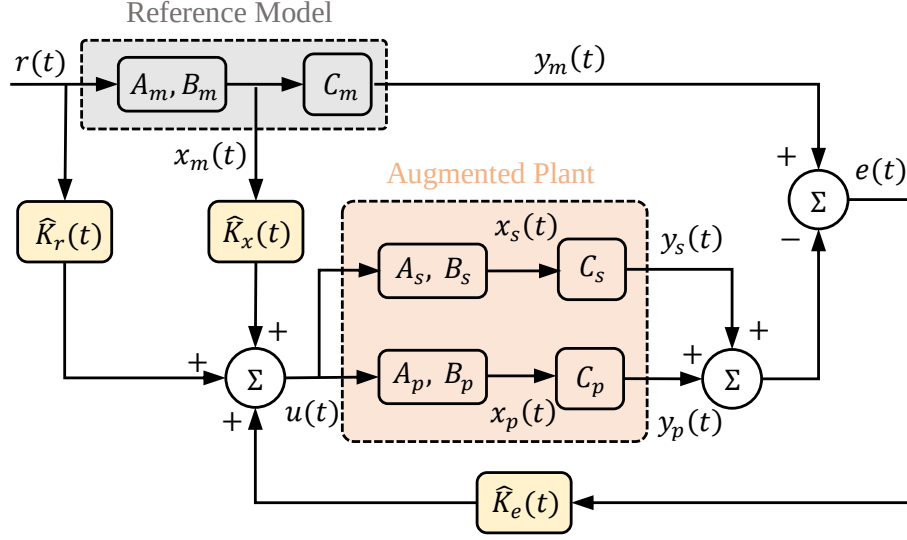


Figure 3.7: General block diagram of a direct model reference adaptive controller with plant feedforward augmentation [103].

seen in Fig. 3.7, to guarantee stability and asymptotic tracking [98]. Herein, the augmented system described by matrices A_s , B_s , and C_s can be implemented as a first-order transfer function $k_s/(1 + \tau s)$, as shown in Fig. 3.6.

3.3.3 Stability Analysis

The stability of the MRAC scheme can be proven by analyzing the sign of the time derivative of the following quadratic Lyapunov function:

$$V(t) = e(t)^T \Psi e(t) + \text{tr} \left\{ \left[K_i(t) - \tilde{K} \right] T^{-1} \left[K_i(t) - \tilde{K} \right]^T \right\} \quad (3.29)$$

where Ψ is a positive definite symmetric matrix, “tr” denotes trace, and $\tilde{K}$ is dummy gain matrix only used in $V(t)$ to study the stability of the controller. If the system is considered to be ASPR, then the derivative of $V(t)$ can be written as [98]

$$\dot{V}(t) = \dot{e}(t)^T \Psi e(t) + e(t)^T \Psi \dot{e}(t) + 2e(t)^T \left[K_i(t) - \tilde{K} \right] r(t) - 2e(t)^T e(t) r(t)^T \bar{T} r(t). \quad (3.30)$$

Herein, $\dot{V}(t)$ is not necessarily negative definite or semi-definite. However, by applying the Gronwall-Bellman Lemma [98] it can be proven that the Lyapunov function $V(t)$ formulated in (3.29) is bounded, which translates into $e(t)$ and $K_i(t)$ being bounded as well. Consequently, the Modified Invariance Principle for non-linear non-autonomous systems gives the following theorem of stability for the developed adaptive algorithm:

Theorem: Assuming the system to be controlled is ASPR, which guarantees the existence of two positive definite matrices Ψ and Φ , with

$$A_m^T \Psi + \Psi A_m = -\Phi; \quad (3.31)$$

and considering $\bar{T}$ and T are positive definite and positive semi-definite matrices. Then all states and gains of the adaptive system defined by (3.23)-(3.28) are bounded, and the output tracking error $e(t)$ vanishes asymptotically [98, 103].

3.3.4 Controller Design

The general design procedure for the developed MRAC scheme, applicable to a three-phase ac-dc smart power converters in a more electric powertrain is as follows:

1. **Feedforward Compensator Design:** Advanced optimization methods based on Kharitonov polynomials can be utilized to determine the exact values of k_s and τ as described in [143]. Nonetheless, for practical implementation τ can be set to $\tau = 1/(2\pi f_c)$, where f_c is the maximum input frequency of the system, e.g., 800 Hz for MEA applications. On the other hand, the gain k_s can be set as $k_s = 1/(\sqrt{2}V_{base})$, where V_{base} is the rated line-to-line input voltage of the system, e.g., 230/115 V for a MEA.
2. **Adaptation Gains Selection:** For multi-input-multi-output (MIMO) systems, the derivation of an analytical expressions to compute the adaptation gains can be a difficult task. A practical approach is to initially set $\gamma_p = 0$ and $\gamma_i = k_s$, then increase γ_i until the system achieves the desired dynamic performance, i.e., settling time and

overshoot. Once the value of γ_i is set, γ_p can be computed as $\gamma_p = \gamma_i/\tau_s$, where τ_s is the settling time of the reference model [98].

3. **Reference Model Selection:** The order of the reference model described by A_m , B_m , and C_m can be selected to be equal or lower than the order of the system. Subsequently, the components of A_m , B_m , and C_m are selected such that the reference model achieves the desired dynamic step response in terms of overshoot and settling time. One design approach is to set the reference model to exhibit a critically damped response, thus A_m and B_m are defined as:

$$A_m = \begin{bmatrix} -\frac{1}{\tau_m} & 0 \\ 0 & -\frac{1}{\tau_m} \end{bmatrix}, B_m = \begin{bmatrix} \frac{1}{\tau_m} & 0 \\ 0 & \frac{1}{\tau_m} \end{bmatrix}, \quad (3.32)$$

where, τ_m is the time constant of the model, and C_m is a 2×2 identity matrix, i.e., $C_m = I_{2 \times 2}$. Using these definitions for A_m , B_m , and C_m results in a model whose steady-state values are unitary and have a settling time of $4\tau_m$ [144].

3.4 Summary

In summary, this chapter covered the mathematical formulation, development, and implementation of the step-ahead controller, the indirect adaptive step-ahead controller, and the seamless dynamics direct MRAC scheme. In the next chapter, the experimental results validating and assessing the steady-state and dynamic performance of the aforementioned controllers are presented.

Chapter 4

Experimental Results

In this chapter, the experimental results validating the smart power converter control schemes developed in Chapter 3 are presented. The controllers were tested under variable-frequency/constant-amplitude and variable-frequency/variable-amplitude input voltage to emulate a more electric powertrain generator operation. The variable-frequency/constant-amplitude study case emulates a MEA variable-frequency electric power generation system; whereas, the variable-frequency/variable-amplitude represents a more generic powertrain topology in which the onboard generators could be a PM synchronous machine. In all cases, the model-based dq -control scheme, as described in Section 2.4, was used as the benchmark controller.

This chapter contains five sections. In Section 4.1, the hardware setup implemented to verify the feasibility and performance of the smart power converter control schemes is described; and the results are presented in Sections 4.2, 4.3, and 4.4 for the step-ahead controller, step-ahead controller with adaptive online parameter estimation, and the seamless dynamics direct MRAC, respectively. Lastly, Section 4.5 demonstrates the performance of the aforementioned controller under a variable-frequency/variable-amplitude input voltage study case.

4.1 Hardware Setup

The performance of the controllers was verified experimentally using the custom-built laboratory-scale SiC-MOSFET-based 2L-VSC shown in Fig. 4.1. The converter is formed by three main print circuit boards (PCB):

- **Top PCB – Power Stage:** the top board contains a grid interfacing inductive filter with $L_f = 1$ mH and $R_f = 0.25 \Omega$, and two $200 \mu\text{F}/360$ V dc-bus electrolytic capacitors connected in series. A dc-bus pre-charge circuit consisting of a MOSFET with a ceramic 10 W resistor is equipped as part of the converter dc-bus circuit to provide a smooth charging of the dc-bus electrolytic capacitors. Lastly, the pre-charge circuit MOSFET is switched using an external control signal. This control signal allows the bypassing of the pre-charge resistor when the dc-bus voltage exceeds 45 V. Once the ≥ 45 V condition is reached, the pre-charge resistor will stay bypassed until the voltage drops below 10 V. This logic is implemented via software.
- **Middle PCB – Measurements and Sensors:** the middle board is equipped with three voltage sensors and three current sensors. The measured voltages are v_{ab} , v_{bc} , and v_{dc} ; whereas the measured currents are i_a , i_b , and i_c . Each voltage is measured with a resistor-based voltage divider circuit and an instrumentation amplifier. On the other hand, the currents are sensed using Hall Effect sensors. The six measurements are directly connected with BNC cables to the A/D acquisition ports of the controller board.
- **Bottom PCB – Gate Driver and Power Module:** the bottom board houses a CREE (CCS020M12CM2) SiC-MOSFET three-phase module driven by a CGD15FB45P gate driver, which receives the gate signals directly from the controller board.

The control schemes were digitally implemented using a dSPACE MicroLabBox 1202 with a PWM switching frequency of 20 kHz. All the controllers used a $50 \mu\text{s}$ sampling time, i.e., $T_s = 1/20$ kHz, except for the adaptive parameter estimation technique which was set to run at half the sampling time of the controller, i.e., $25 \mu\text{s}$ [145].

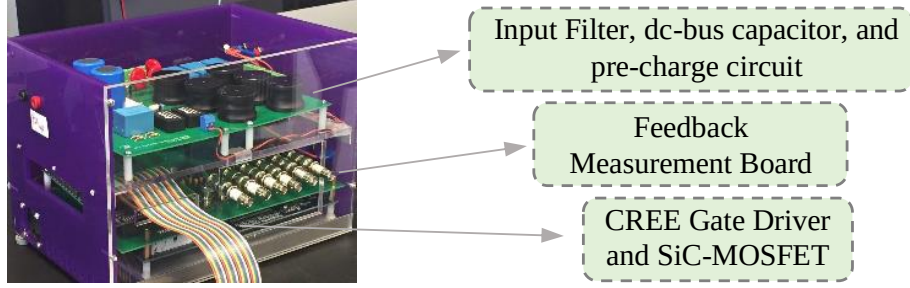


Figure 4.1: Proof-of-concept laboratory-scale SiC-MOSFET-based 2L-VSC [99].

Table 4.1: Direct phase-angle detection parameter settings [63]

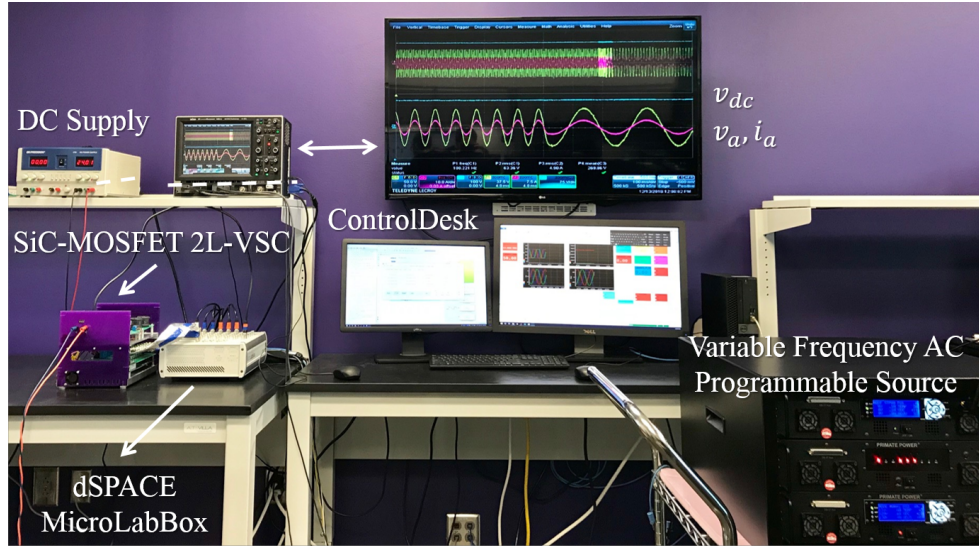
	Parameter	Value
	Filter Order, n	2
Low-Pass Filter	Frequency, $\omega_n = 2\pi f_n$	$f_n = 800$ Hz
	Damping Ratio, ζ	0.707
Rate Limiter	Rising slew rate	1,500 Hz/s
	Falling slew rate	-1,500 Hz/s

In addition, the 2L-VSC was connected to a 12 kW Primate Power programmable grid emulator from Abstract Power Electronics with an input line-to-line voltage of 115 V at a variable-frequency between 100 – 400 Hz, being 400 Hz the maximum output frequency at which the hardware can operate. The converter was loaded at 1 kW using a resistive load with a dc voltage set-point of 270 V and unity power factor, i.e., $Q_{ref} = 0$. Notice that the results and conclusions drawn at 1 kW and maximum frequency of 400 Hz can be extended to higher power ratings and frequency levels as only the dynamic performance of the controllers is being studied.

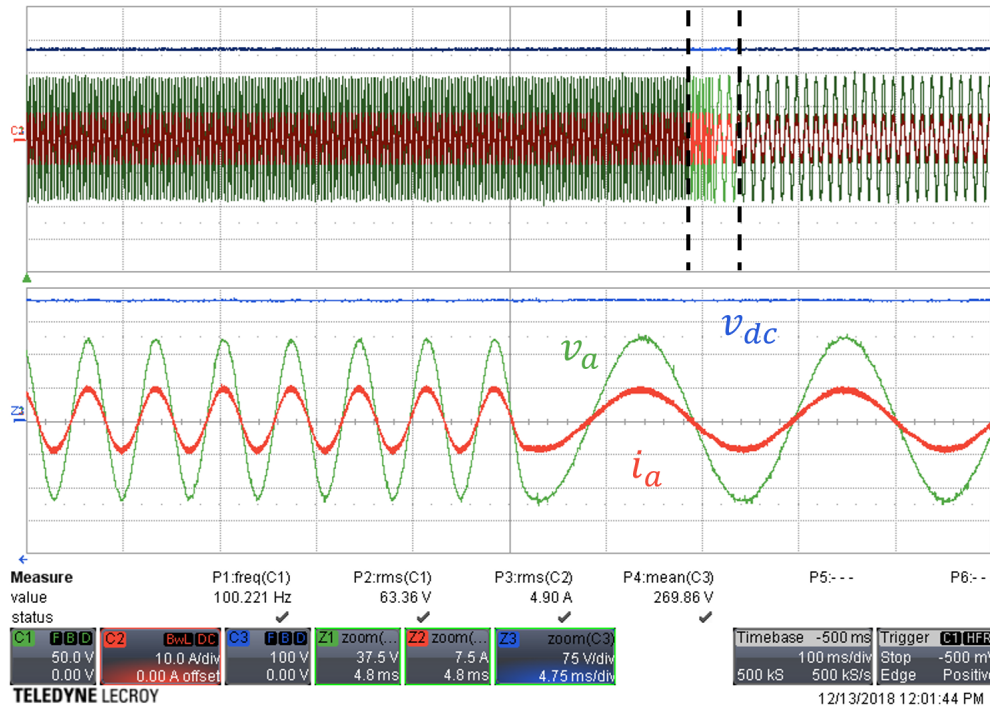
Finally, the PLL scheme implemented for phase-angle detection and frequency tracking is the direct phase-angle technique as described in Fig. 2.9(c). The parameters of the PLL are outlined in Table 4.1.

4.2 Step-Ahead Control Scheme

The step-ahead control scheme was experimentally validated using the hardware setup described in Section 4.1. For performance evaluation and benchmarking, the model-based



(a)



(b)

Figure 4.2: Experimental validation; (a) hardware setup, and (b) oscilloscope screenshot of the step-ahead control scheme under a frequency step increase [63].

Table 4.2: Parameters used in the experimental implementation of the step-ahead controller [63]

Block Description	Parameter	Value
v_{dc} , PI controller gains	k_p	15
	k_i	10
Q , PI controller gains	k_p	0.01
	k_i	0.1

dq –controller explained in Section 2.4 was implemented with the same hardware setup and used as a benchmark. The proportional and integral gains of the dc-bus voltage PI control loop in both controllers were set to the same values as depicted in Table 4.2. These values were selected based on the stability margin of the model-based dq –controller at the maximum operating frequency of the grid emulator, i.e., 400 Hz. The data shown in this section was captured using a Teledyne LeCroy WaveRunner 64Xi-A 600 MHz oscilloscope at a rate of 500 kS/s, and then re-plotted for enhanced visibility and presentation clarity.

4.2.1 Steady-State Waveforms and Line Current THD

Fig. 4.3 shows the phase-to-neutral voltages and line currents for the strap-ahead and model-based dq –controllers at 100 Hz and 300 Hz. In terms of the line current THDs, both controllers show similar harmonic distortion at the two tested input frequencies. In all cases, the THD is less than 2.5%, thus complying with the IEEE Std. 519 [146]. Notice that the line current THD is mainly a function of the converter topology, switching frequency, filter design, and the THD of the PCC voltage. Consequently, the THD is presented only as a power quality indicator to verify the correct operation of the compared controllers. In this sense, the main comparison between the controllers is based on their dynamic performance, which is presented in the following subsection.

4.2.2 Dynamic Performance

In this subsection, the transient performance of the step-ahead and model-based dq –controllers is analyzed for three distinct perturbations: (i) load step-changes, (ii) frequency ramp changes, and (iii) frequency step-changes. Here, cases (ii) and (ii) are considered directly

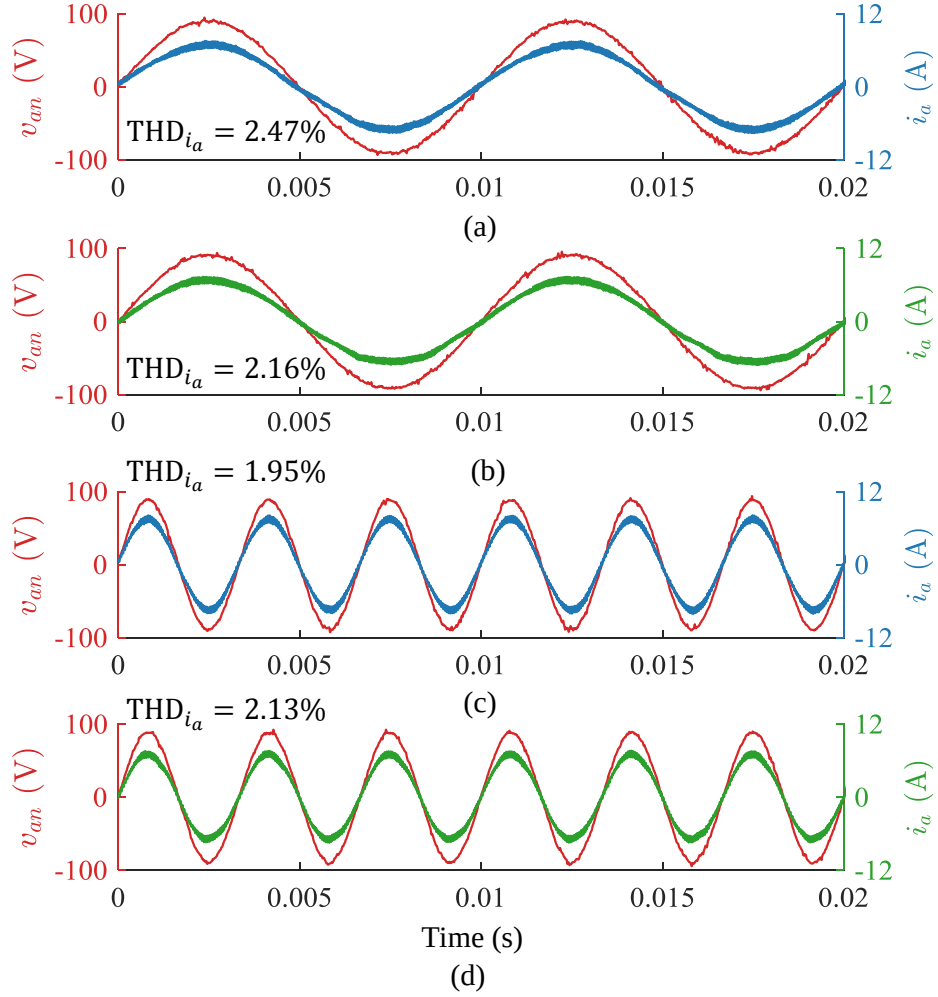


Figure 4.3: Experimentally obtained steady-state phase-to-neutral voltages and line currents waveforms at 1kW for the step-ahead and model-based dq -controllers at 100 Hz and 300 Hz; (a) model-based dq -controller at 100 Hz, (b) step-ahead controller at 100 Hz, (c) model-based dq -controller at 300 Hz, (d) step-ahead controller at 300 Hz [63].

related to more electric powertrain applications and are therefore part of the main contributions of this work. More specifically, the aforementioned cases mimic the dynamics of a MEA electric power generation system, in which the voltage of the generators varies in frequency at a constant voltage amplitude.

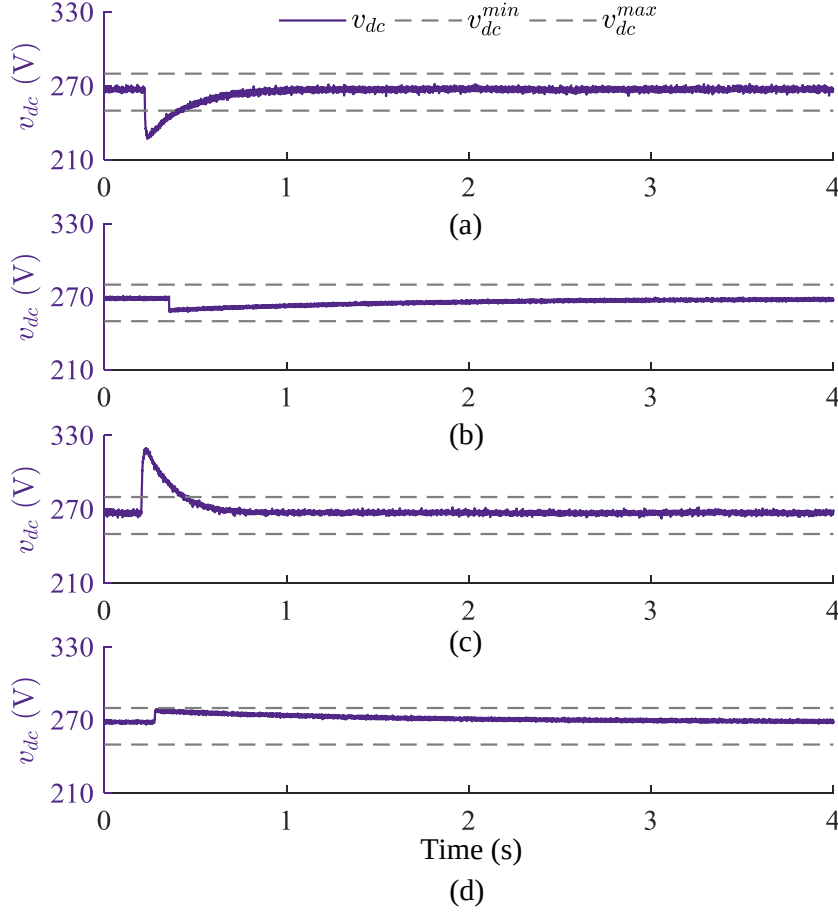


Figure 4.4: Experimentally obtained voltage regulation of the dc-bus for a 50% load step change; model-based dq -control: (a) load increase and (c) decrease; step-ahead controller: (b) load increase and (d) decrease [63].

4.2.2.1 Load Step-Changes

The transient performance of the step-ahead and model-based dq -controllers under load step-changes is depicted in Fig. 4.4. Here the red dashed lines represent the minimum and maximum voltage limits required by the MIL-STD-704F standard [147]. For both controllers, the load was suddenly increased from 500 W to 1 kW and vice versa for the load decrease. Table 4.3 summarizes the maximum and minimum voltage variations, i.e., voltage dips and swells, as well as the settling times for the step-ahead and model-based dq -controllers taken from Fig. 4.4. In both, the load increase and decrease, the step-ahead controller outperforms the model-based dq -controller in terms of the voltage dip and swell generated by the

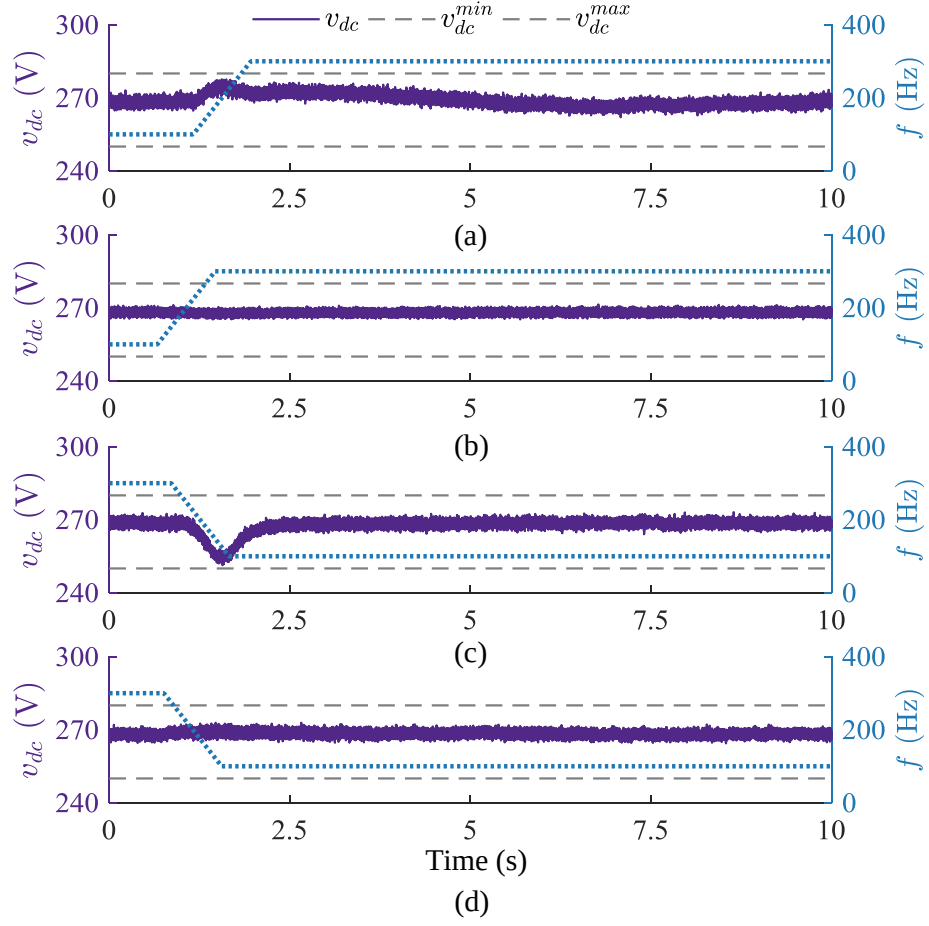


Figure 4.5: Experimentally obtained regulation of the dc-bus voltage for a ± 250 Hz/s input frequency ramp; model-based dq -control: (a) acceleration and (c) deceleration; step-ahead controller: (b) acceleration and (d) deceleration [63, 99].

load transients. Conversely, model-based dq -control scheme shows a better performance in regards to the settling time of the dc-bus voltage.

Table 4.3: Experimentally obtained transient performance indicators for load step-changes [63]

Control Scheme	Load Increase		Load Decrease	
	Voltage Dip	Settling Time	Voltage Swell	Settling Time
Model-based dq	-42 V	1.09 s	+50 V	0.96 s
Step-ahead	-12 V	1.66 s	+10 V	1.60 s

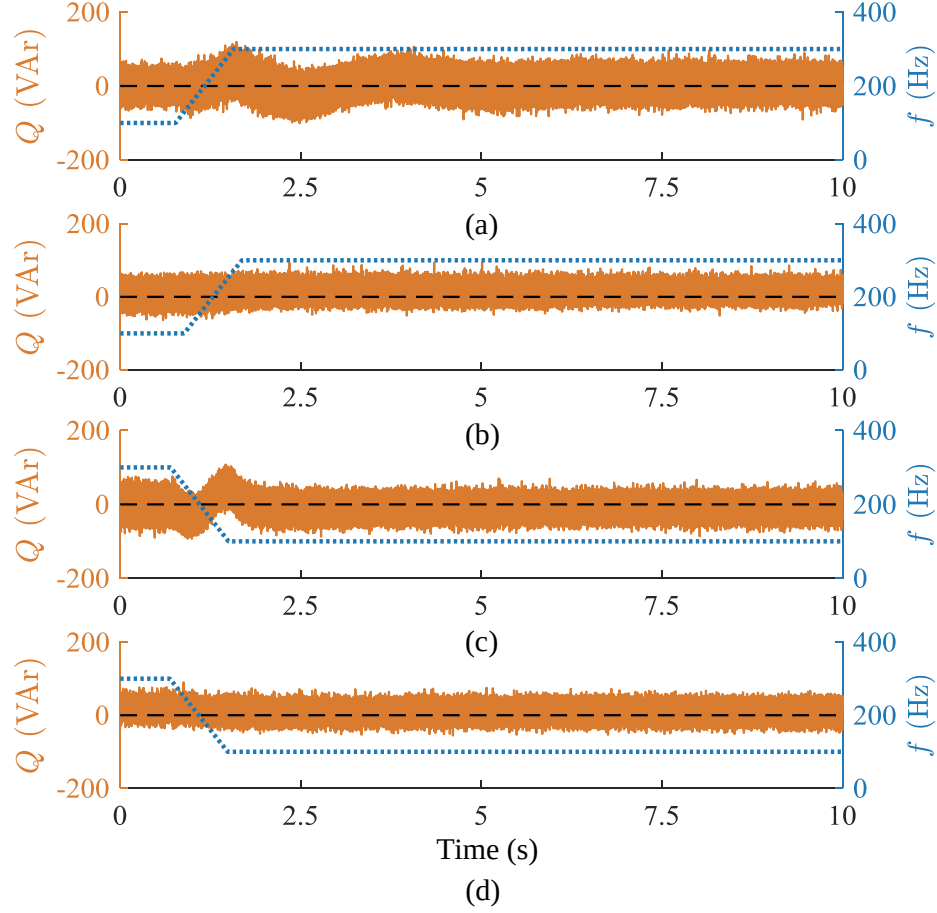


Figure 4.6: Experimentally obtained regulation of the reactive power for a ± 250 Hz/s input frequency ramp; model-based dq -control: (a) acceleration and (c) deceleration; step-ahead controller: (b) acceleration and (d) deceleration [63, 99].

4.2.2.2 Input-Frequency Ramps

As recommended by the Aircraft Electric Power Characteristics MIL-STD-704 standard [147], the maximum frequency rate of change in a MEA is limited to ± 250 Hz/s. Figs 4.5 and 4.6 show the dynamic performance of the step-ahead and model-based dq -controllers when the frequency is increased and decreased at ± 250 Hz/s. As can be seen in Fig. 4.5, the step-ahead control scheme effectively regulates v_{dc} in both the acceleration and deceleration frequency ramps. When compared to the step-ahead controller, the model-based dq -control scheme exhibits noticeable voltage swells and dips. Similarly, the reactive power regulation follows the same pattern as shown in Fig. 4.6. Here, both controllers maintain Q around

zero; however, the step-ahead controller offers a more stable dynamic response with a lower settling time and overshoot.

4.2.2.3 Frequency Step-Changes

To evaluate the dynamic performance of the developed controller at the fastest possible frequency variation rate, two frequency step-changes were applied to both the model-based dq -controller and the step-ahead controller. The first step change corresponding to an increase of the frequency from 100 to 300 Hz is depicted in Figs. 4.7(a) and 4.7(c), while Figs. 4.7(b) and 4.7(d) correspond to a 300 to 100 Hz step decrease. In general, Fig. 4.7 shows the dynamic response of the phase-to-neutral voltage v_a , the line current of the same phase i_a , and the dc-bus voltage v_{dc} . Here, it is clear how the step change in frequency negatively affects the dc voltage regulation of the model-based dq -controller, being the worst case observed when decreasing the frequency. Conversely, the step-ahead control scheme exhibits an *ultrafast* dynamic response to both frequency step-changes, which translates into a successful dc voltage regulation. Moreover, with the developed controller, v_a and i_a are always in phase during the frequency transients, thus maintaining a unity power factor at the ac mains at all times. In fact, this successful and accurate unity power factor regulation is a direct consequence of combining the step-ahead algorithm with the direct phase-angle detection technique as demonstrated in Fig. 4.8. Here, the performance of the step-ahead controller was studied with and without the instantaneous direct phase-angle detection technique for a 100 – 300 Hz frequency step change. As can be seen from Fig. 4.8, as quick as the first zero-crossing of the voltage and current after the frequency step change, the step-ahead control scheme maintains a phase difference of 1.2° , whereas without the direct phase-angle detection, this phase difference nearly doubles. This tendency is also maintained for the second positive slope zero-crossing.

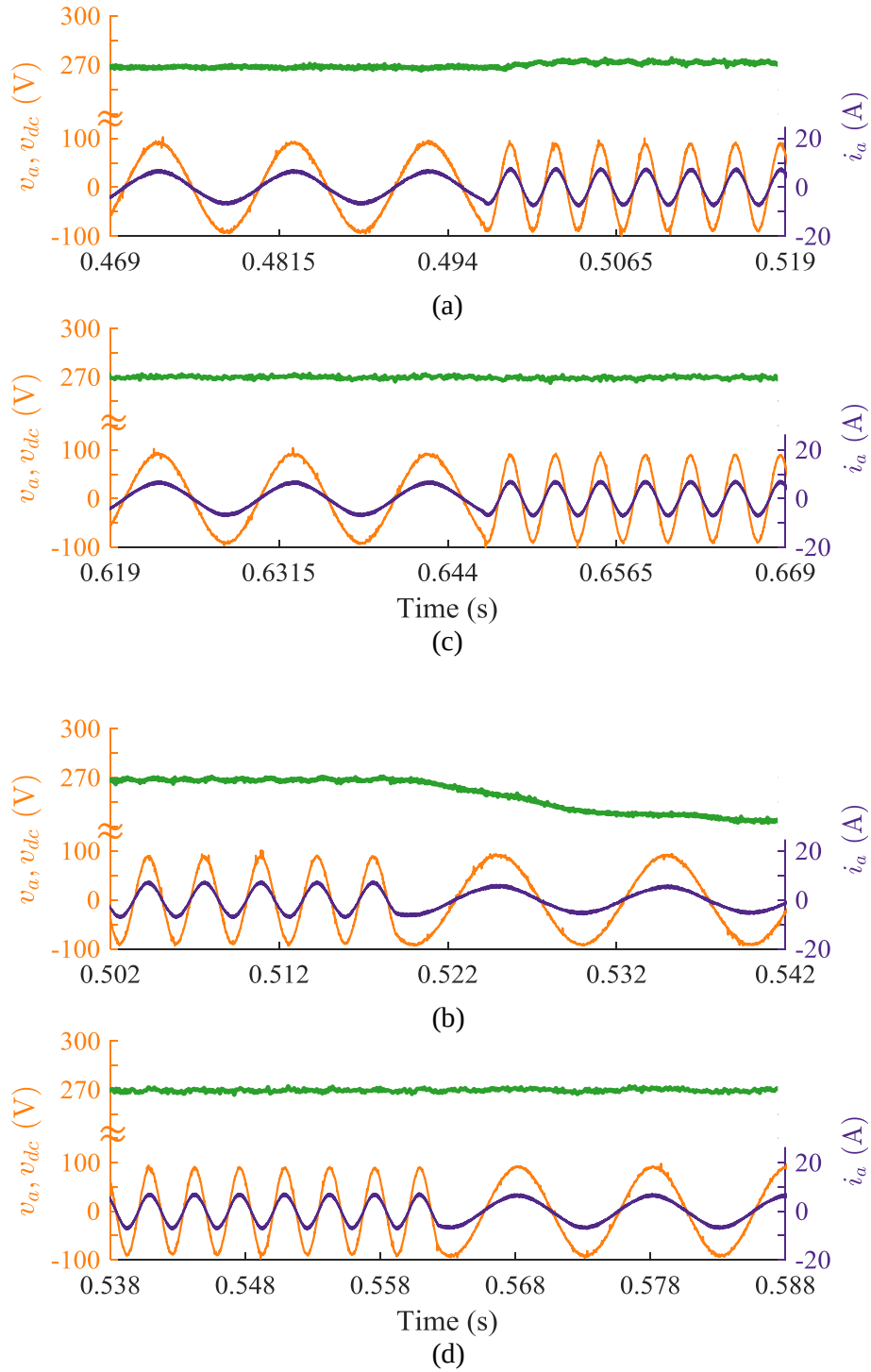


Figure 4.7: Experimentally obtained dynamic performance under a step change in the supply frequency; (a) model-based dq -controller, 100 to 300 Hz, (b) model-based dq -controller, 300 to 100 Hz, (c) step-ahead controller, 100 to 300 Hz, and (d) step-ahead controller, 300 to 100 Hz [63,99].

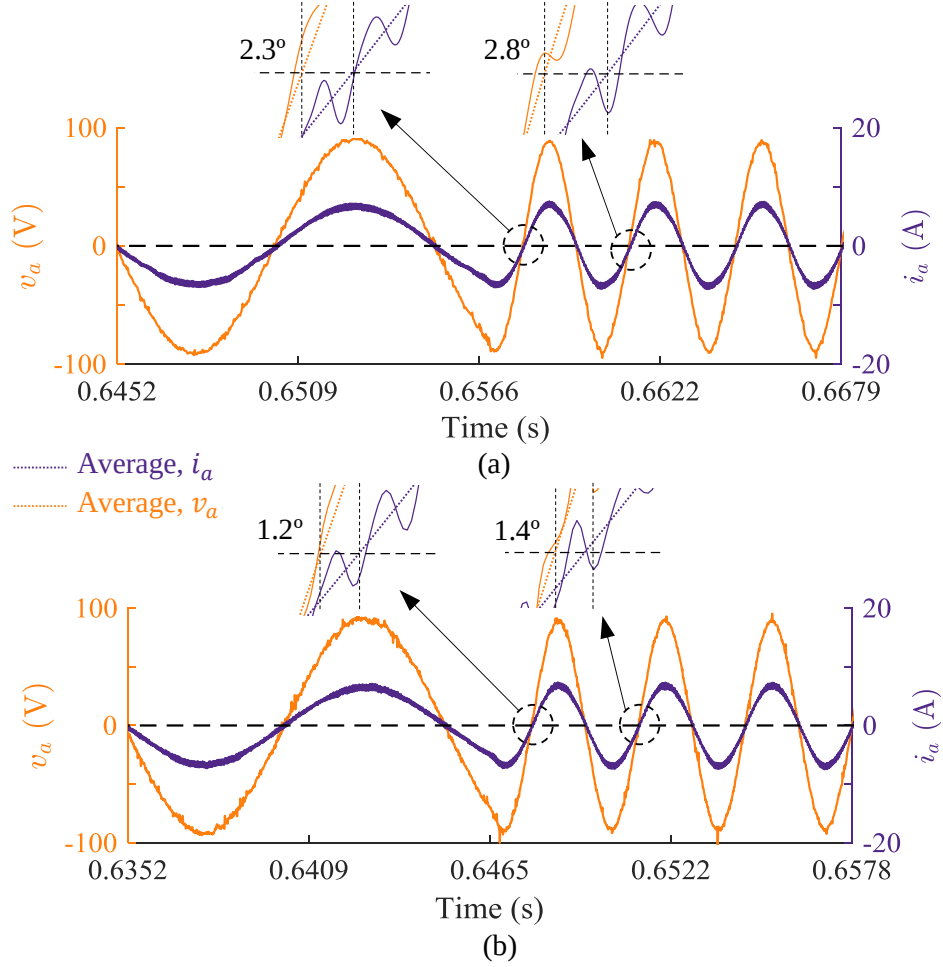


Figure 4.8: Experimentally obtained dynamic performance of the step-ahead controller under a 100 to 300 Hz step change in the supply frequency; (a) no phase-angle detection, and (b) with direct phase-angle detection [99].

4.2.3 Parameter Mismatch

A well-known disadvantage of MPC schemes is the performance degradation they suffer under parameter mismatch [68]. The step-ahead controller belongs to the MPC category. As such, it is important to analyze and compare its robustness to parameter mismatch under the context of variable-frequency applications. Fig. 4.9 depicts the dynamic performance of the step-ahead and model-based dq -controllers under a step-change in the supply frequency with 50% parameter mismatch. In the case of the model-based dq -controller, the parameter mismatch was induced by reducing the L_f parameter in the inner PI-loops, see Fig. 2.12, to half of the actual hardware value of L_f . In a similar way, the step-ahead controller was

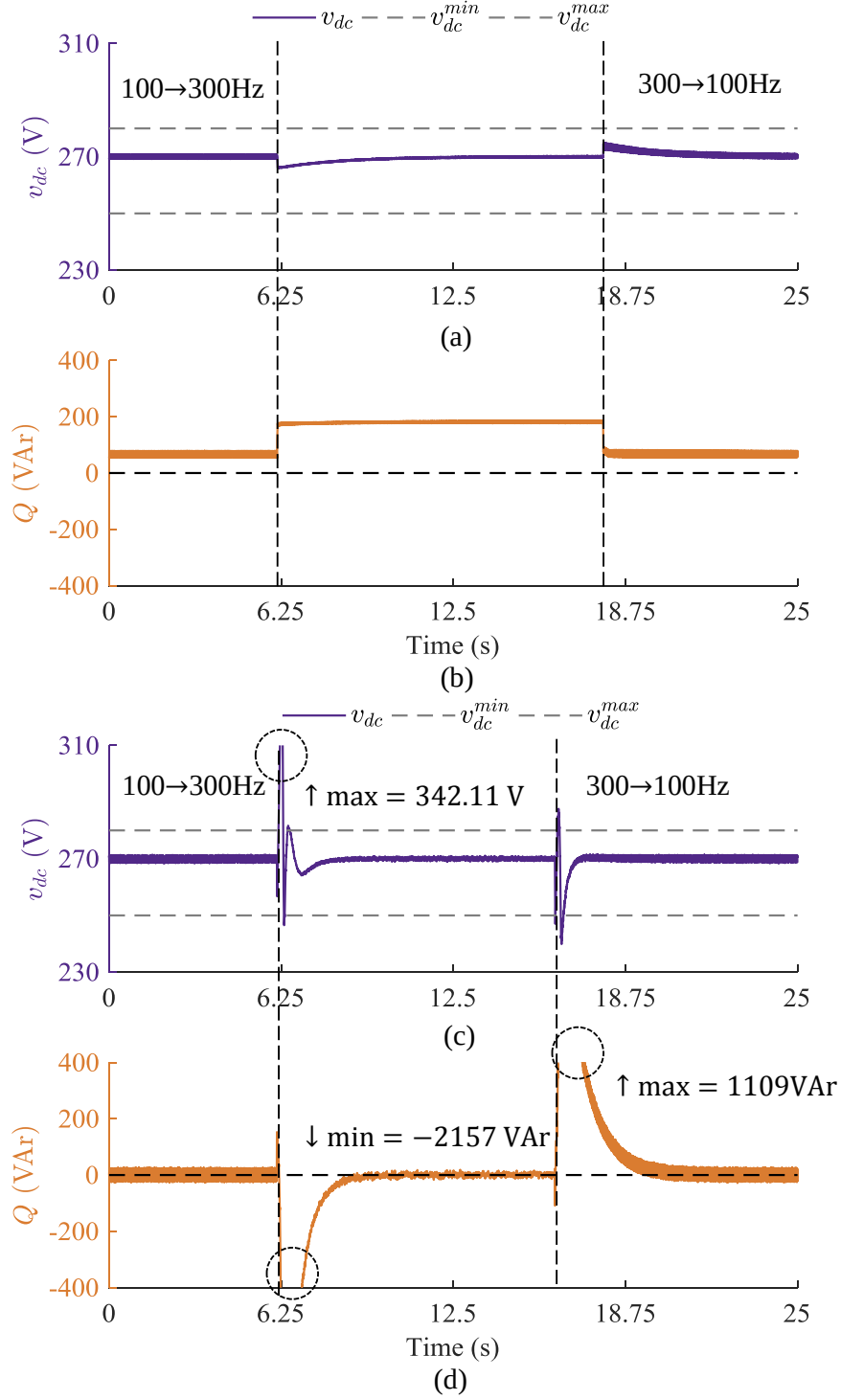


Figure 4.9: Experimentally obtained dynamic performance under a step change in the supply frequency with 50% parameter mismatch in L_f . Regulation of v_{dc} : (a) step-ahead controller and (c) model-based dq -controller. Regulation of Q : (b) step-ahead controller and (d) model-based dq -controller [63, 101].

Table 4.4: Experimentally obtained steady-state error under 50% parameter mismatch

Controller	Steady-State Error	
	v_{dc} (V)	Q (VAr)
Model-based dq	≈ 0	≈ 0
Step-ahead	≈ 0	70 VAr at $f = 100$ Hz 185 VAr at $f = 300$ Hz

Table 4.5: Experimentally obtained dc-bus overshoot under 50% parameter mismatch

Controller	DC-Bus Overshoot			
	100 $\rightarrow$ 300 Hz	300 $\rightarrow$ 100 Hz	100 $\rightarrow$ 300 Hz	300 $\rightarrow$ 100 Hz
Model-based dq	11.16%	-26.71%	-215.7%	110.9%
Step-ahead	1.49%	-1.81%	7%	18.5%

altered so that it uses half of L_f in (3.5). As can be seen in Fig. 4.9, the step-ahead control scheme exhibits a noticeable steady-state performance degradation under parameter mismatch, with the reactive power control being the most affected. Furthermore, Fig. 4.9(c) shows how the steady-state error in Q increases with the frequency reaching up to 185 VAr at 300 Hz. Notice that v_{dc} does not have a steady-state error as it is regulated by a PI-control loop, see Fig. 3.1. Conversely, the model-based dq -controller has no steady-state error due to its PI-based structure. That being said, the model-based dq -controller produces large overshoots when compared to the step-ahead control scheme. Tables 4.4 and 4.5 summarize the performance indicators for the compared controllers.

4.2.3.1 Reactive Power Error at Variable-Frequency

The error in the reactive power due to the mismatch between the actual frequency and the frequency used in the step-ahead algorithm was validated experimentally by measuring the reactive power with and without the direct phase-angle detection algorithm. Table 4.6 outlines how the error in the reactive power is reduced and maintained constant when implementing the direct phase-angle detection algorithm, whereas the without phase-angle detection, the error increases proportionally with frequency as predicted by (3.7).

Table 4.6: Experimentally measured per unit reactive power error as a function of frequency ($S_{base} = 1$ kW) [99]

f (Hz)	No Phase-Angle Detection Error (%)	Phase-Angle Detection Error (%)	Reduction (%)
100	4.41	1.14	3.27
150	6.62	1.38	5.24
200	8.82	1.66	7.16
250	11.03	2.05	8.98
300	13.24	2.44	10.80

Table 4.7: Experimentally measured sensitivity of the step-ahead controller to parameter mismatch (R_f) as a function of frequency [101, 102]

f (Hz)	Error ΔQ when $Q_{ref} = 0$ VAr				
	R_f used in (3.5)				
	0.2500 Ω	0.1875 Ω	0.1250 Ω	0.3125 Ω	0.3750 Ω
100	16.30	16.03	15.94	16.59	16.80
200	28.34	27.94	27.50	28.80	29.22
300	41.70	41.17	40.61	42.48	43.01
400	54.50	53.60	52.80	55.21	56.19

4.2.3.2 Reactive Power Steady-State Error and Parameter Mismatch

To analyze the reactive power error, ΔQ , when $Q_{ref} = 0$, the values of R_f and L_f used in (3.5) were changed from 75% to 125% of their rated values. Tables 4.7 and 4.8 outline the experimentally measured reactive power error ΔQ for the aforementioned parameter mismatch scenarios.

As can be seen from Tables 4.7 and 4.8, the ΔQ magnitudes when R_f is changed are negligible compared to the errors generated by the L_f variations. Furthermore, Table 4.8 demonstrates how $|\Delta Q|$ increases proportionally with the input frequency, hence proving experimentally that for variable-frequency systems an accurate parameter estimation technique is required to avoid any performance degradation on the controller.

Table 4.8: Experimentally measured sensitivity of the step-ahead controller to parameter mismatch (L_f) as a function of frequency [101, 102]

Error ΔQ when $Q_{ref} = 0$ VAr					
f (Hz)	L_f used in (3.5)				
	1 mH	0.75 mH	0.50 mH	1.25 mH	1.50 mH
100	16.30	37.73	78.61	3.21	-5.72
200	28.34	69.50	146.40	3.16	-13.97
300	41.70	101.70	213.20	4.70	-20.60
400	54.50	132.01	275.20	5.48	-28.03

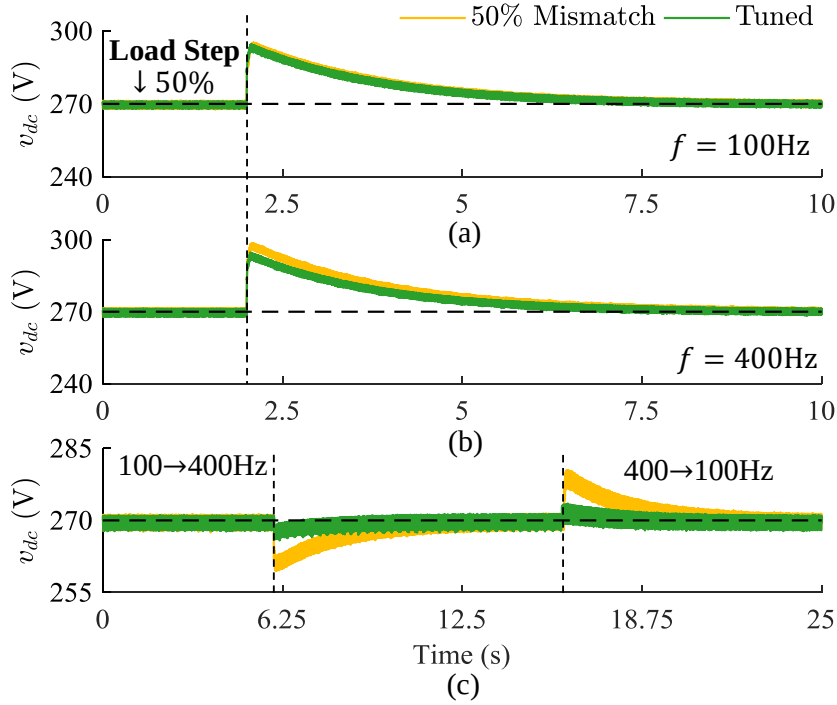


Figure 4.10: Experimentally obtained dc-bus voltage dynamic performance under parameter mismatch; (a) 50% load step decrease at 100 Hz, (b) 50% load step decrease at 400 Hz, and (c) frequency step-change of ± 300 Hz [101, 102].

4.2.3.3 DC-Bus Voltage Dynamic Performance under Parameter Mismatch

Fig. 4.10 shows the dynamic performance of v_{dc} with a 50% parameter mismatch, i.e., the inductance of the filter in the hardware is 50% more than the inductance used in (3.5) to compute $\vec{v}_R[k]$. Figs. 4.10(a) and 4.10(b) depict the dynamic response of the dc-bus for a 50% (500W) load step decrease at 100 Hz and 400 Hz, respectively. At both input

frequencies, the parameter mismatch increases the v_{dc} overshoot by 3.13% at 100 Hz, and 16.23% at 400 Hz. On the other hand, Fig. 4.10(c) displays the dc-bus dynamics when the input frequency is stepped from 100 Hz to 400 Hz and vice-versa. Here, the overshoot increases by approximately 2.5% under parameter mismatch. In summary, a parameter mismatch in the filter inductance deteriorates the dynamic performance of the step-ahead controller in terms of the dc-bus voltage regulation. This phenomenon is amplified at higher input frequencies, and could even worsen in some converter designs due to the presence of the dc-bus circuit parasitic inductance.

4.3 Indirect Adaptive Step-Ahead Control Scheme

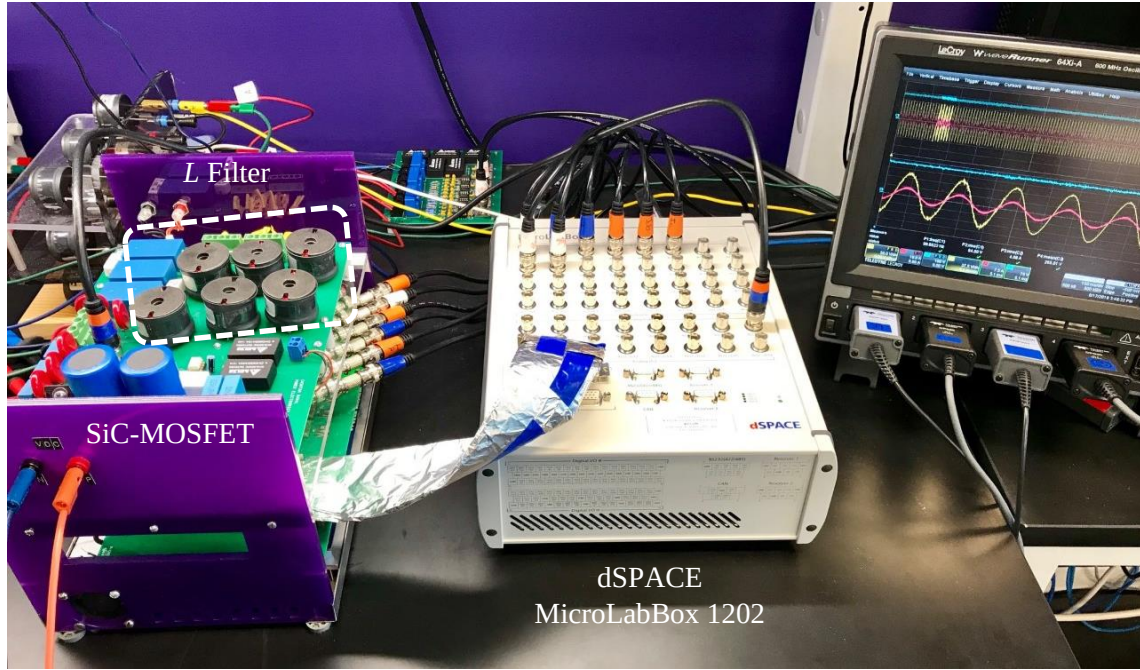
The indirect adaptive step-ahead control scheme was experimentally validated using the hardware setup described in Section 4.1. In addition to the SiC-MOSFET-based 2L-VSC, an additional set of 1 mH inductors were included with a bypass circuit breaker as part of the inductive filter as depicted in Fig. 4.11(a). The operation of the previously mentioned circuit breaker allowed a 100% increase in the filter inductance and was used to evaluate the dynamic performance of the adaptive parameter estimation technique.

The steady-state and dynamic performance of the indirect adaptive step-ahead control scheme was analyzed using the step-ahead controller as a benchmark. More specifically, the performance comparisons were performed with and without the adaptive parameter estimation algorithm.

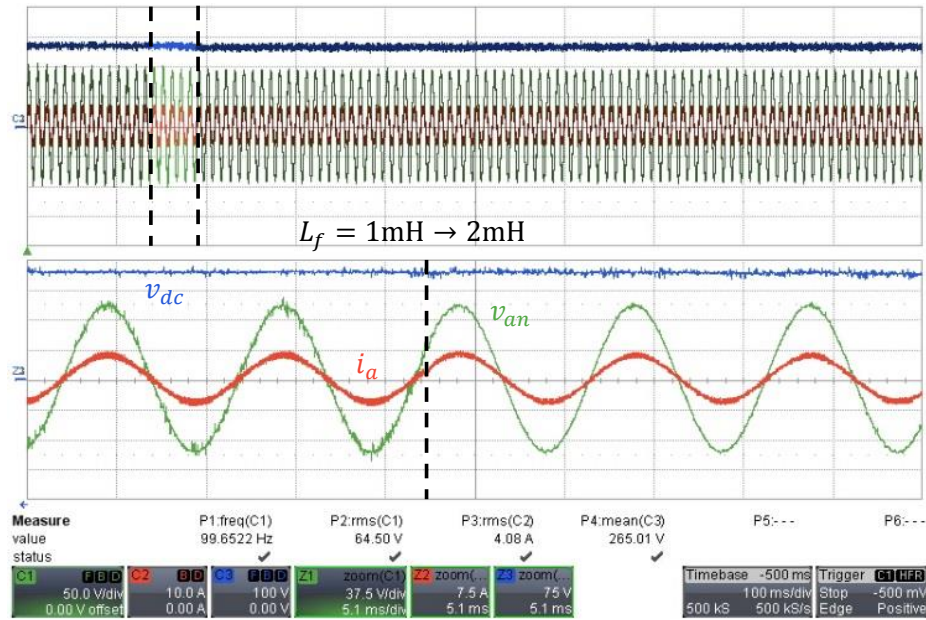
The controller was digitally implemented using a dSPACE MicroLabBox 1202 with a:

- PWM switching frequency of 20 kHz
- 25 μ s sampling time for the adaptive parameter estimation
- 50 μ s for the step-ahead controller.

In all the tested scenarios, the gains for the PI dc-bus voltage control loop were set to $k_p = 15$ and $k_i = 10$. Finally, to account for any asymmetries present in the inductive filter, the



(a)



(b)

Figure 4.11: Experimental verification; (a) hardware setup, and (b) oscilloscope screen capture when the inductance of the filter is increased by 1mH in the hardware [101].

adaptive parameter estimation algorithm was applied to all three phases. Consequently, the parasitic resistance R_{f0} and inductance L_{f0} of the filter were computed using the average of the per-phase values. Lastly, the ac-voltages and currents shown in this section were

Table 4.9: Steady-state experimental estimation of the average R_{f0} and L_{f0} [101,102]

f (Hz)	R_{f0} (Ω)	L_{f0} (mH)	e (%), $I_{base} = 5$ A
100	0.5981	1.1437	0.70%
200	0.6041	1.1004	0.46%
300	0.6240	1.0931	0.54%
400	0.6645	1.0901	0.56%

captured using a Teledyne LeCroy WaveRunner 64Xi-A 600 MHz oscilloscope at a rate of 500 kS/s, whereas the reactive power, dc-bus, and estimated parameters were recorded at a sampling frequency of 20 kHz using the dSpace ControlDesk acquisition toolbox. In both cases, the collected data was re-plotted to enhance its visibility and presentation clarity.

4.3.1 Adaptive Parameter Estimation Steady-State Performance

Sub-sections 4.2.3.2 and 4.2.3.3 demonstrate the detrimental effects that parameter mismatch can cause to the step-ahead controller. Furthermore, it was also experimentally demonstrated that the reactive power error ΔQ is proportional to the input frequency. Table 4.9 outlines the experimentally estimated parameters of the inductive filter, i.e., R_{f0} and L_{f0} . As can be seen from the previously mentioned table, both parameters vary as a function of the frequency. The parasitic resistance of the inductive filter R_{f0} increases with frequency, while the inductance L_{f0} decreases.

In the case of R_{f0} , the skin and proximity effects become more relevant as the input frequency increases raising the value of R_{f0} . Notice that R_{f0} is considered to be the ac-resistance of the filter since it is estimated based on ac voltages and currents. In contrast, the increase in the input frequency magnetic reduces the flux penetration in the core of the inductors, which translates into a higher effective reluctance. Consequently, the estimated inductance of the filter L_{f0} tends to reduce its magnitude as the frequency increases [77, 78, 148]. Lastly, the far-right column of Table 4.9 outlines the percent state error, e , using a base current of 5 A for every input frequency operating point. In all cases, the error is below 0.70%, with an average error of 0.56% for the tested input frequency range, which demonstrates the accuracy of the developed adaptive parameter estimation technique.

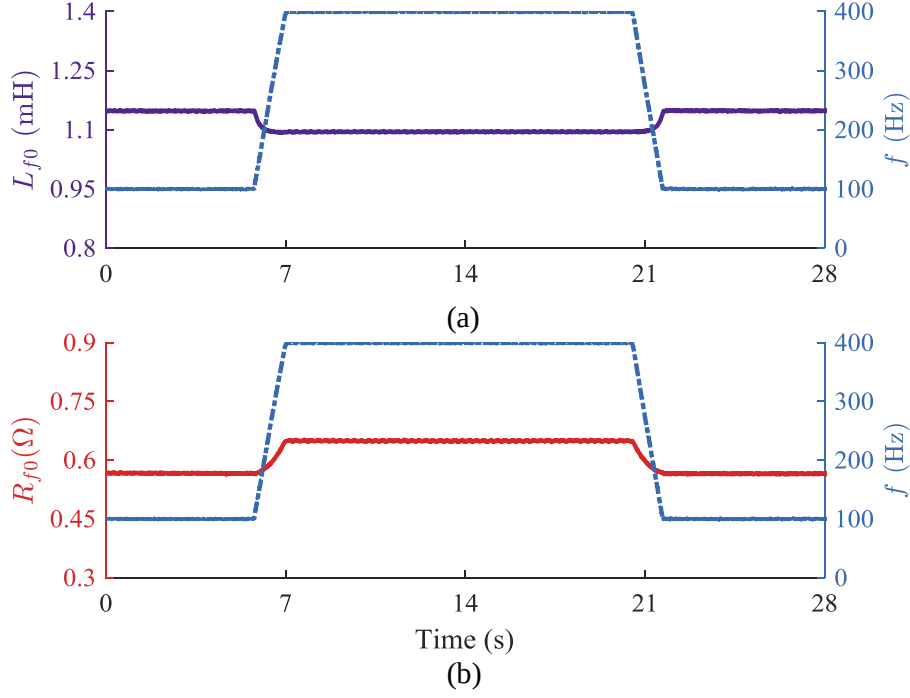


Figure 4.12: Experimentally obtained dynamic adaptive parameter estimation during a ± 250 Hz/s input frequency ramp; (a) estimated filter inductance L_{f0} and (b) estimated filter parasitic resistance R_{f0} [101, 102].

4.3.2 Adaptive Parameter Estimation Dynamic Performance

Fig. 4.12 shows the dynamic estimation of R_{f0} and L_{f0} when the input frequency is increased/decreased linearly. The slope of the frequency ramps was set to ± 250 Hz/s based on the Aircraft Electric Power Characteristics MIL-STD-704 [147] standard. Fig. 4.12 is consistent with the steady-state parameter estimation outlined in Table 4.9, where R_{f0} increases with frequency and L_{f0} decreases. Furthermore, Fig. 4.12 demonstrates how R_{f0} and L_{f0} can be tracked dynamically using the adaptive parameter estimation technique developed in Section 3.2 with minimum delay; hence maintaining the inherently rapid dynamic performance of the step-ahead control scheme.

To further demonstrate the efficacy of the adaptive parameter estimation technique, the developed method is compared with the least-square parameter estimation algorithm developed in [76]. Fig. 4.13 shows the filter inductance L_{f0} estimation for a 250 Hz/s slope frequency ramp from 100 Hz to 400 Hz, and when a 1 mH is purposely switched in between

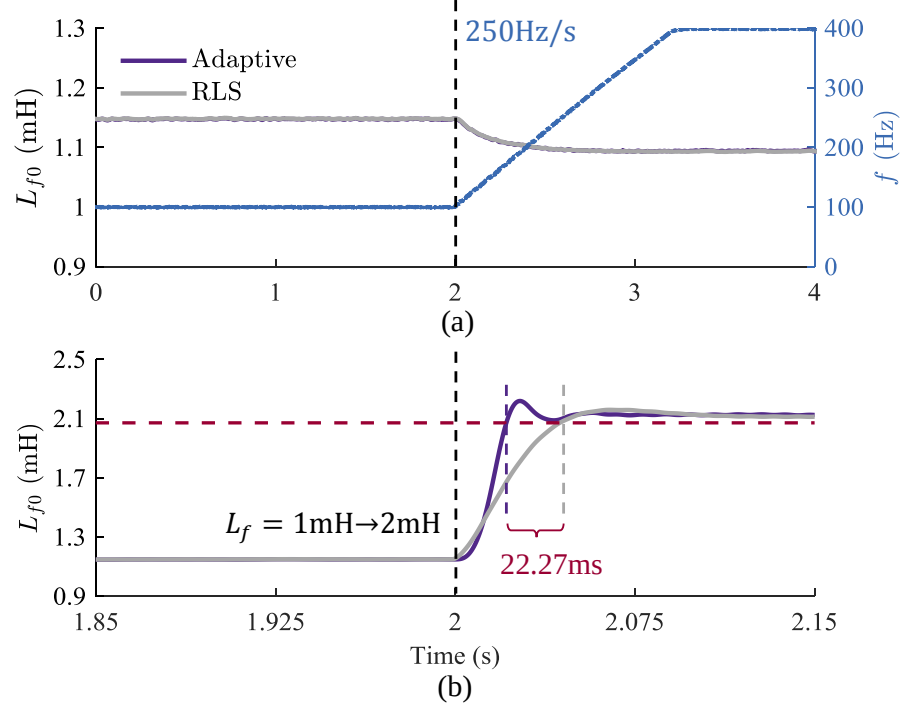


Figure 4.13: Experimentally obtained dynamic performance comparison between the developed adaptive parameter estimation technique and least-square estimation [76]; (a) L_{f0} estimation for a 250 Hz/s slope frequency, and (b) 1 mH inductance step increase [101, 102].

the converter and the PCC. Fig. 4.13(a) proves that both parameter estimation techniques accurately track the changes in L_{f0} under frequency variations. Nonetheless, Fig. 4.13(a) illustrates that the developed adaptive parameter estimation technique achieves a settling time of approximately half, i.e., 51.05% reduction, compared to the least-square estimation; which demonstrates its dynamic performance superiority.

On the other hand, Fig. 4.14 displays the reactive power regulation when the inductance of the filter is intentionally increased in 1 mH in the hardware at different input frequencies, i.e., 100 Hz and 400 Hz. An external circuit breaker was implemented to introduce an additional 1 mH inductor between the converter and the PCC. This abrupt 100% variation of the filter inductance is intended to evaluate the robustness of the developed parameter estimation technique with the controller, and not to replicate actual variations that may occur on a physical inductive filter. That being said, Figs. 4.14(b) and 4.14(d) demonstrate the dynamic performance of the adaptive parameter estimation algorithm in terms of

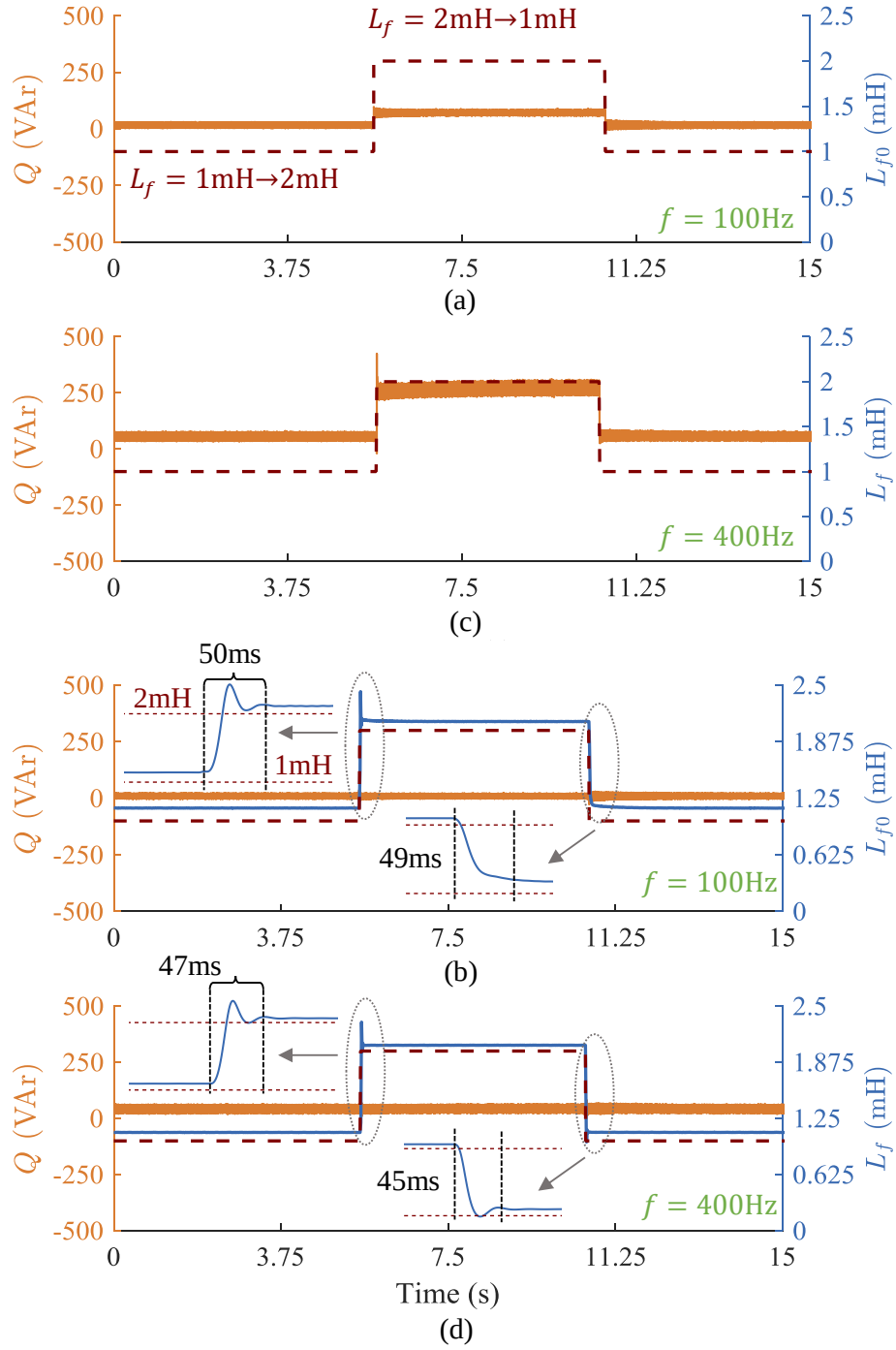


Figure 4.14: Experimentally obtained reactive power regulation when the L filter is increased by 1 mH in the hardware at different input frequencies; (a) adaptive parameter estimation algorithm disabled at 100 Hz, (b) adaptive parameter estimation algorithm enabled at 100 Hz, (c) adaptive parameter estimation algorithm disabled at 400 Hz, and (d) adaptive parameter estimation algorithm enabled at 400 Hz [101,102].

the estimation of the filter inductance. It can be seen in both figures that once the 1 mH inductance is inserted in the system, the algorithm adjusts the estimated inductance in a maximum time of 50 ms at both 100 Hz and 400 Hz. In addition, Figs. 4.14(b) and 4.14(d) depict the reactive power regulation when the adaptive parameter estimation is enabled. Here, the reactive power remains regulated even after the filter inductance is increased and decreased by 1 mH. Conversely, Figs. 4.14(a) and 4.14(c) reveal the negative effects on the step-ahead controller when the inductance is changed in the hardware and not adjusted in the model; which translates into an important reactive power tracking error that amplifies proportionally with the input frequency.

Finally, Fig. 4.11(b) depicts the phase-to-neutral voltage, line current, and dc-bus voltage waveforms when the inductance of the filter is increased by 1 mH, and the adaptive parameter estimation is enabled. This oscilloscope screen capture verifies the quality of the voltage and current waveforms as well as the ability of the controller to maintain the phase-to-neutral voltage and line current in phase at all times, thus guaranteeing a unity power factor operation.

4.3.3 Performance Improvements when Implementing the Adaptive Parameter Estimation

In a more electric powertrain generation system, the input frequency can vary rapidly due to the acceleration and deceleration of the prime movers, e.g., jet engines and ICE. At the same time, the temperature inside the vehicle or aircraft can range from -80°C to $+180^{\circ}\text{C}$ [149,150]. Under these temperature and frequency variations, the core of the filter inductors change their permeability and saturation point, which leads to dramatic changes in the inductance of the filter [79,80]. Consequently, without the help of a parameter estimation technique, the step-ahead control scheme could operate under severe parameter mismatch.

That being said, and considering the effectiveness of the developed adaptive parameter estimation technique has been demonstrated in Figs. 4.12 and 4.14 by tracking the parameter of the grid-interfacing filter during frequency ramp changes and the sudden increase of the

Table 4.10: Experimentally obtained dynamic performance indicators taken from Fig. 4.15 when the adaptive parameter estimation algorithm is enabled [101, 102]

Control Scheme	Max. Value		Min. Value	
	Q (VAr)	v_{dc} (V)	Q (VAr)	v_{dc} (V)
Step-Ahead	312.62	277.96	49.52	260.85
Adaptive Step-Ahead	66.70	272.25	-6.09	266.71
Diff. (%)	78.66%	2.05%	112.30%	2.25%

filter inductance, this subsection presents the improvements in the dynamic response of the step-ahead controller when it operates in conjunction with the developed adaptive parameter estimation technique.

Fig. 4.15 depicts the dynamic response of the step-ahead controller when the input frequency is linearly increased and decreased at ± 250 Hz/s rate, and the inductance of the filter is 50% more than the inductance used in (3.5) to compute $\vec{v}_R[k]$. More specifically, Figs. 4.15(a) and 4.15(c) demonstrate the performance degradation suffered by the step-ahead controller when the developed parameter estimation technique is disabled. In contrast, Figs. 4.15(b) and 4.15(d) exhibit how the same control variable, i.e., v_{dc} and Q are effectively regulated when the parameter estimation is engaged. Finally, Table 4.10 outlines the minimum and maximum values or overshoots achieved by v_{dc} and Q during the frequency ramp transients. Herein, the v_{dc} overshoot is reduced in 2.15% by activating the adaptive parameter estimation technique; whereas, the Q is reduced by 95.48% on average. Notice, that the reactive power reduction is substantially higher than the dc-bus since the control of this variable directly depends on the accurate computation of (3.5), while v_{dc} is regulated by an external PI-control loop.

4.4 Seamless Dynamics Direct Model Reference

Adaptive Control (MRAC) Scheme

The performance of the MRAC scheme was verified experimentally using the hardware setup described in Section 4.1. For performance evaluation and benchmarking, the model-

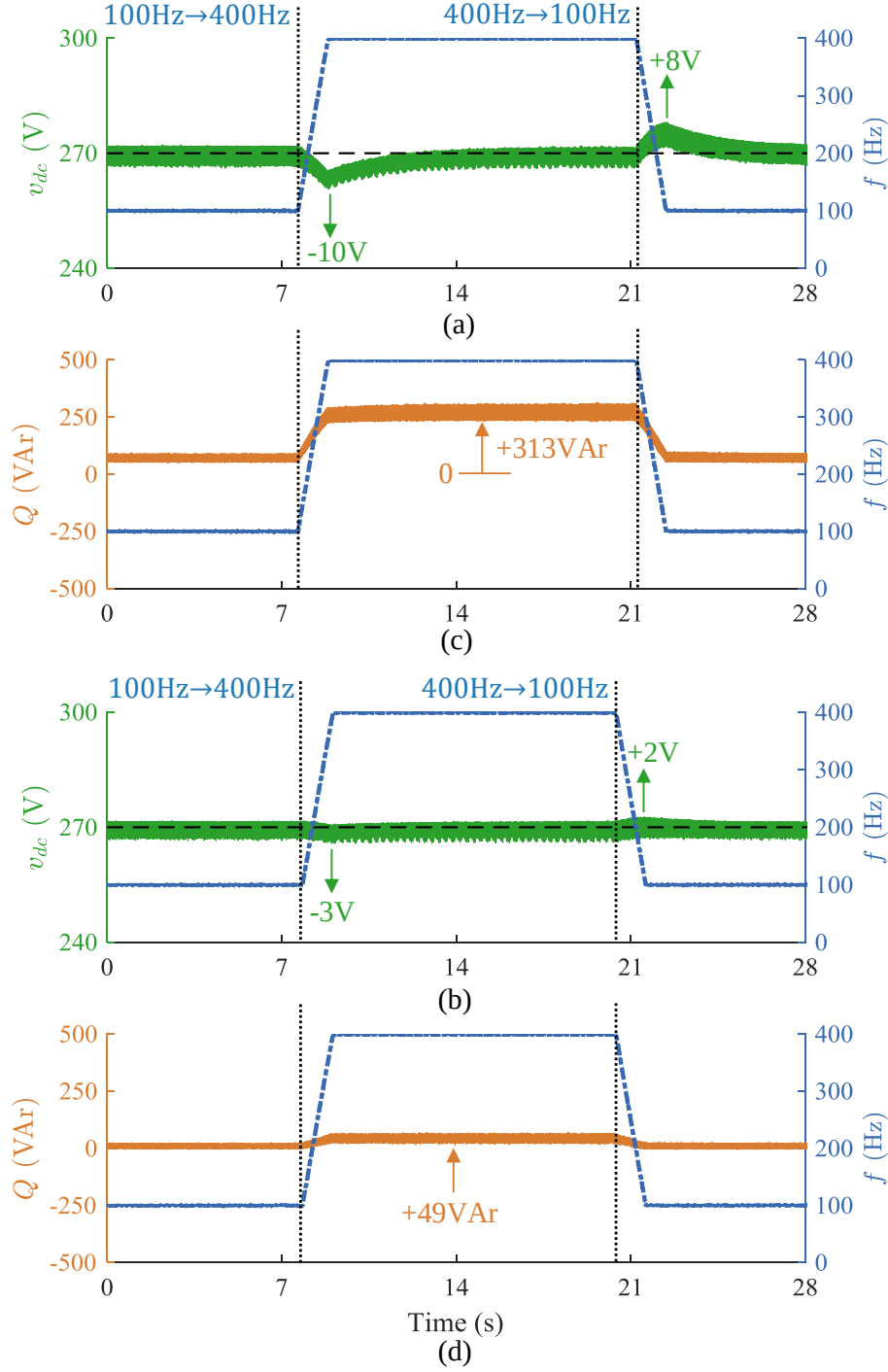


Figure 4.15: Experimentally obtained dc-bus voltage and reactive power regulation during a ± 250 Hz/s input frequency ramp with a 50% parameter mismatch; (a) dc-bus voltage with the adaptive parameter estimation algorithm disabled, (b) dc-bus voltage with the adaptive parameter estimation enabled, (c) reactive power with the adaptive parameter estimation algorithm disabled, and (d) reactive power with the adaptive parameter estimation enabled [101, 102].

based dq -controller explained in Section 2.4 was implemented with the same hardware setup and used as a benchmark. The control schemes were digitally implemented using a dSPACE MicroLabBox 1202 with a PWM switching frequency of 20 kHz, and sampling time of 50 μ s. The execution time of the developed MRAC scheme is 28.08 μ s. Lastly, the ac-voltages and currents shown in this section were captured using a Teledyne LeCroy WaveRunner 64Xi-A 600 MHz oscilloscope at a rate of 500 kS/s, whereas the reactive power, dc-bus, and controller gains were recorded at a sampling frequency of 20 kHz using the dSpace ControlDesk acquisition toolbox. In both cases, the collected data was re-plotted to enhance its visibility and presentation clarity.

4.4.1 Gain Selection and Benchmarking

To establish fair benchmarking criteria, the gains of the dc-bus and reactive power outer PI-control loops, i.e., k_{p1} , k_{i1} , k_{p2} , and k_{i2} , for the MRAC and model-based dq -control schemes were tuned such that both controllers exhibit similar dynamic performance, i.e., overshoot and settling time, for a specific perturbation at a constant input frequency. In the dc-bus voltage PI-loops, the gains were adjusted using a 50% load step decrease, as shown in Fig. 4.16(a). In contrast, the reactive power PI-loops gains were tuned by implementing a 250 VAr step-change in the reactive power reference q_{ref} , as depicted in Fig. 4.16(b). In both cases, the gains of the controllers were adjusted to achieve similar overshoots and settling times t_s . The resulting gains are outlined in Table 4.11. These gains remained unaltered throughout the experiments in this section for both control schemes.

Table 4.11: Proportional and integral gains for the dc-bus and reactive power PI-control loops [103]

Control Loop	Model-based dq		MRAC	
	k_p	k_i	k_p	k_i
v_{dc}	0.15	4.00	0.15	1.00
Q	0.01	0.20	0.01	0.10

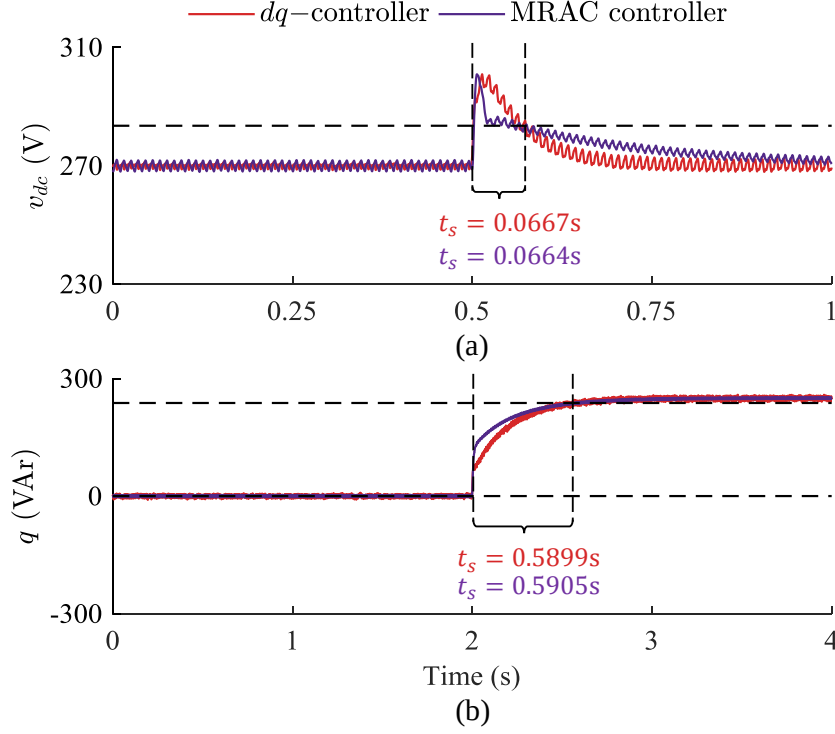


Figure 4.16: Experimentally obtained transients responses used to tune the PI outer-loop controller gains; (a) load step-change and (b) reactive power reference step-change [103].

4.4.2 Steady-State Waveforms and Frequency Spectrum

Fig. 4.17 depicts the steady-state waveforms (v_{an} and i_{an}), line current frequency spectra, and phase-to-neutral voltage harmonic spectra for the MRAC and model-based dq -controllers. Herein, all the figures in the left column demonstrate the capability of the aforementioned control schemes to maintain a unity power factor successfully. Furthermore, when supplying a 1 kW load at regulated v_{dc} , the MRAC and model-based dq -controllers exhibit similar line current THDs ($< 3\%$) at 100 Hz and 400 Hz. Notice that these low THDs levels are achieved with approximately 2% THD in the input voltage.

For partial loading conditions from 0.4 kW to 1 kW, the line current THD and PF for both controllers at 100 Hz and 400 Hz are outlined in 4.12. Here, the controllers perform similarly in terms of the line current THD, and PF steady-state values. As expected for an inductive filter, the THD of the line current increases as the load decreases, registering maximum values of approximately 5%. This tendency remains valid with the increase of the

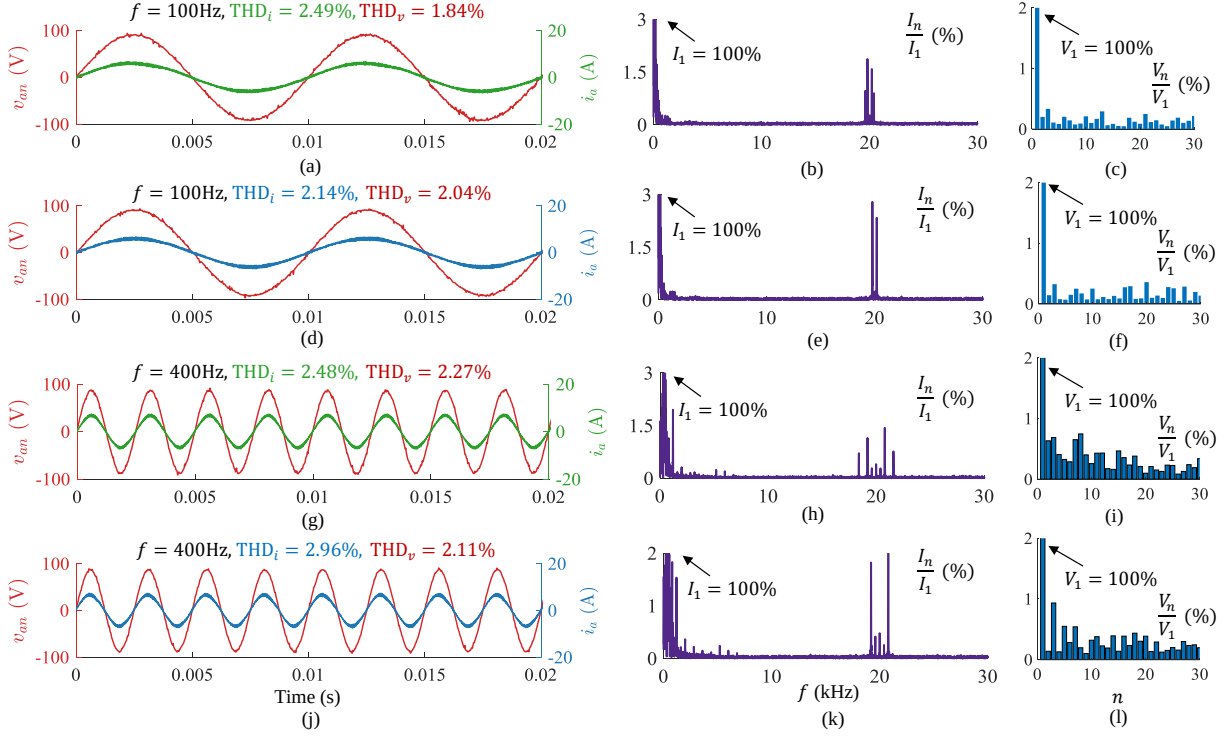


Figure 4.17: Experimentally obtained steady-state phase-to-neutral voltages, line currents, line current frequency spectra, and low order harmonic spectra of the phase-to-neutral voltages at 1 kW. (a), (b), and (c) model-based dq -controller at 100 Hz; (d), (e), and (f) MRAC at 100 Hz; (g), (h), and (i) model-based dq -controller at 400 Hz; (j), (k), and (l) MRAC at 400 Hz [103].

input frequency. On the other hand, the load variation does not affect the PF steady-state error.

In summary, Fig. 4.17 and Table 4.12 demonstrate that the compared controllers exhibit similar line current THDs and achieve unity power factor operation at the tested frequencies. In fact, the current THD is independent of the control scheme implemented, as it mainly depends on the topology of the converter, switching frequency, filter design, and load level. For this reason, the THD data presented in this subsection is considered only as a power quality index and not as a performance indicator. The comparison of the controllers is primarily focused on their dynamic performance, which is presented in the next subsections.

Table 4.12: Experimentally obtained line current THD and power factor (PF) for the developed MRAC scheme and the model-based dq -controller as a function of the load

Load (W)	model-based dq -controller			
	100 Hz		400 Hz	
	PF	THD	PF	THD
400	1.00	4.77%	1.00	5.52%
600	1.00	3.95%	1.00	4.10%
800	1.00	2.88%	1.00	3.38%
1,000	1.00	2.49%	1.00	2.48%
Load (W)	MRAC			
	100 Hz		400 Hz	
	PF	THD	PF	THD
400	1.00	4.41%	1.00	5.94%
600	1.00	3.25%	1.00	4.50%
800	1.00	2.46%	1.00	3.55%
1,000	1.00	2.14%	1.00	2.96%

4.4.3 Dynamic Performance Under Frequency Step-Changes

To verify the dynamic performance of the developed MRAC in comparison to the model-based dq -controller, two $\Delta f = \pm 300$ Hz frequency step-changes, i.e., $100 \rightarrow 400$ Hz and $400 \rightarrow 100$ Hz, were applied to the system. In the particular case of the model-based dq -controller, the current cross-coupling gains ωL_f and $-\omega L_f$, see Fig. 2.12, were tuned using three distinct criteria: (i) the gains are updated online using the frequency f computed by the direct phase-angle detection scheme, (ii) the gains are assumed constant with the frequency set to the minimum expected value, i.e., $f = 100$ Hz, and (iii) the gains are assumed constant with the frequency set to the maximum expected value, i.e., $f = 400$ Hz. That being said, all the v_{dc} graphs shown in this subsection contain two dotted lines that indicate the minimum and maximum voltages permitted by the MIL-STD-704 standard [147].

The dynamic performance of the model-based dq -controller is shown in Fig. 4.18. More specifically, 4.18(a) and 4.18(b) depict the transient response of the dc-bus voltage and reactive power when the frequency is stepped from $100 \rightarrow 400$ Hz and $400 \rightarrow 100$ Hz, and the gains ωL_f and $-\omega L_f$ are updated online with ω taken from the direct phase-angle detection scheme. Herein, the model-based dq -controller exhibits large overshoots in v_{dc}

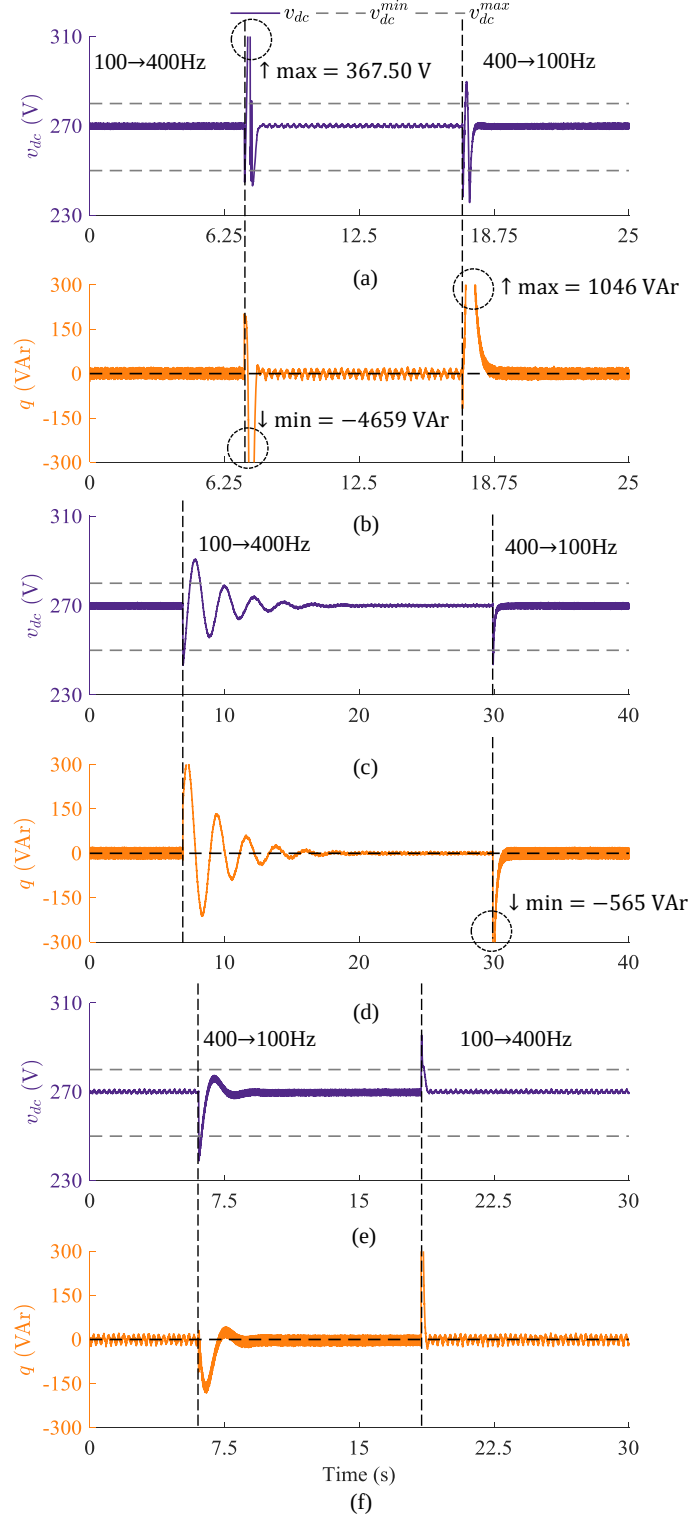


Figure 4.18: Experimentally obtained dynamic performance of the model-based dq -controller under a $\Delta f = \pm 300$ Hz step-change in the supply frequency; (a) v_{dc} and (b) reactive power Q regulation with variable frequency tuning, (c) v_{dc} and (d) reactive power Q regulation with the controller tuned for 100 Hz, (e) v_{dc} and (f) reactive power Q regulation with the controller tuned for 400 Hz [103].

and Q as a consequence of including the PLL dynamics into the inner current control loops. On the other hand, the fixed frequency gain settings produce better results. However, the overshoots for these constant gain cases are still unsatisfactory. Notice that when the $\pm\omega L_f$ gains are tuned to the minimum frequency, i.e., $\omega = 2\pi f$ with $f = 100$ Hz, the controller performs worst when the frequency changes from 100 to 400 Hz than when it varies in the opposite direction, see Figs. 4.18(c) and 4.18(d). This observation also applies when the $\pm\omega L_f$ gains are tuned to the maximum frequency, as shown in Figs. 4.18(e) and 4.18(f).

In general, Fig. 4.18 demonstrates that for a variable frequency system, a constant or variable frequency tuning for ωL_f and $-\omega L_f$ results in a poor dynamic performance of the model-based dq -controller. Based on the previous analysis, it is clear that for a variable frequency system such as MEA, an adaptive controller is required to achieve acceptable dynamic performance. Consequently, with the implementation of the developed MRAC scheme, both v_{dc} and Q remain completely regulated and with seamless dynamics under variable frequency operation, as depicted in Figs. 4.19(a) and 4.19(b). Furthermore, Figs. 4.19(c) and 4.19(d) show the dynamic characteristic of two of the twelve adaptive gains in matrix $\hat{K}(t)$, which demonstrates how the gains adjust their values in accordance with the frequency changes in the system.

A summary of the dynamic performance indicators, i.e., overshoot and settling time for the MRAC and model-based dq -controller, are outlined in Tables 4.13 and 4.14. As can be seen, the MRAC reduced the v_{dc} overshoot from 36% to 2% for a frequency step increase, and from 13% to 1% for a step decrease when compared to the variable frequency tuning model-based dq -control scheme. Moreover, the Q overshoot was reduced by over 1 kVAr for a frequency step increase, and over 4.6 kVAr for a step decrease. On the other hand, in terms of the settling time, the developed MRAC shows practically a null settling time for all the analyzed transients, thus offering a seamless transition between frequencies, which is crucial for MEA applications.

Finally, Fig. 4.20 depicts a more detailed comparison of the seamless dynamics capability of the developed MRAC when compared to the model-based dq -controller with variable frequency ωL_f and $-\omega L_f$ gain adjustment. Here, the line-to-neutral voltage v_a , line current

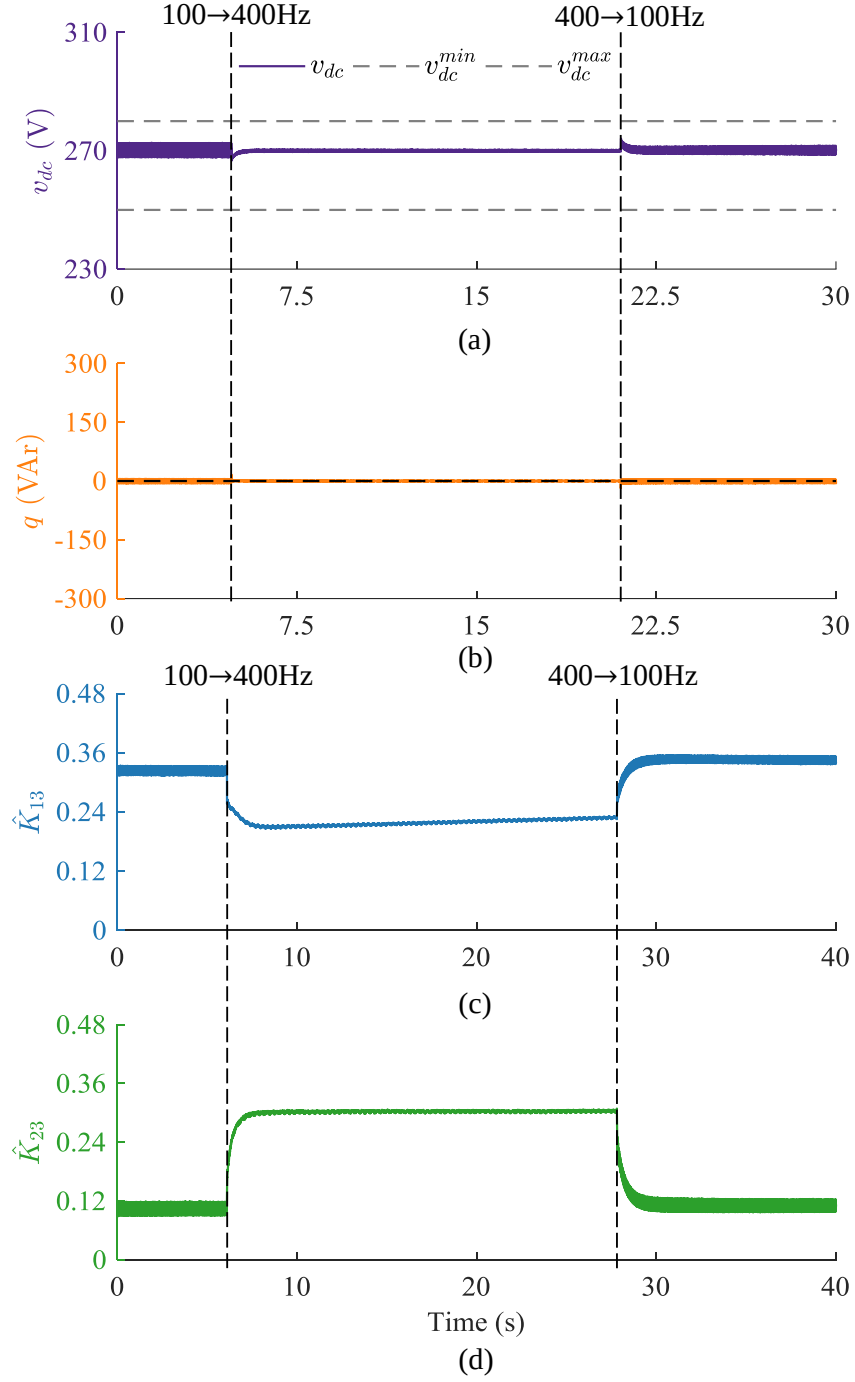


Figure 4.19: Experimentally obtained dynamic performance of the developed MRAC under a 300 Hz step-change in the supply frequency; (a) v_{dc} , (b) reactive power Q , (c) adaptive gains $\hat{K}_{13}$, and (d) $\hat{K}_{23}$ [103].

i_a , and v_{dc} are shown for the 100 $\rightarrow$ 400 Hz and 400 $\rightarrow$ 100 Hz frequency step-changes. It can be seen from Fig. 4.20 that once again the developed MRAC outperforms the model-

Table 4.13: Experimentally obtained minimum and maximum overshoot values for the developed MRAC scheme and the model-based dq -controller [103]

Control Scheme	Tuning	Max. Value		Min. Value	
		Q (VAr)	v_{dc} (V)	Q (VAr)	v_{dc} (V)
Model-based dq -controller	Constant ω	309.95	290.87	-565.28	243.30
	$f = 100$ Hz				
	Constant ω	886.21	295.50	-181.86	238.92
	$f = 400$ Hz				
	Variable ω	1046	367.50	-4659	235.71
MRAC	Adaptive	15.81	274.08	-18.38	266.72

Table 4.14: Experimentally obtained settling time for the developed MRAC scheme and the model-based dq -controller [103]

Control Scheme	Tuning	Settling Time (s)	
		Reactive Power, Q	DC-Bus Voltage, v_{dc}
Model-based dq -controller	Constant ω	5.56	2.14
	$f = 100$ Hz		
	Constant ω	1.02	0.27
	$f = 400$ Hz		
	Variable ω	0.57	0.53
MRAC	Adaptive	≈ 0	≈ 0

based dq -controller in terms of dc-bus regulation and unity power factor. Notice that, as demonstrated in Figs. 4.18(a) and 4.18(b), v_{dc} returns to its reference value after the frequency transient finalizes.

4.4.4 Performance Under Abnormal Grid Conditions

The performance of the MRAC scheme under distorted ac voltage was investigated by intentionally adding a 5% 5th and 7th harmonic components to the PCC voltage. Fig. depicts the steady-state and dynamic performance of the developed MRAC scheme under abnormal grid conditions. More specifically, Figs. 4.21(a) and 4.21(b) depict the dynamic performance for a frequency step-change from 100 Hz to 400 Hz and vice-versa, while Figs. 4.21(c) to 4.21(n) show the steady-state performance at 100 Hz and 400 Hz.

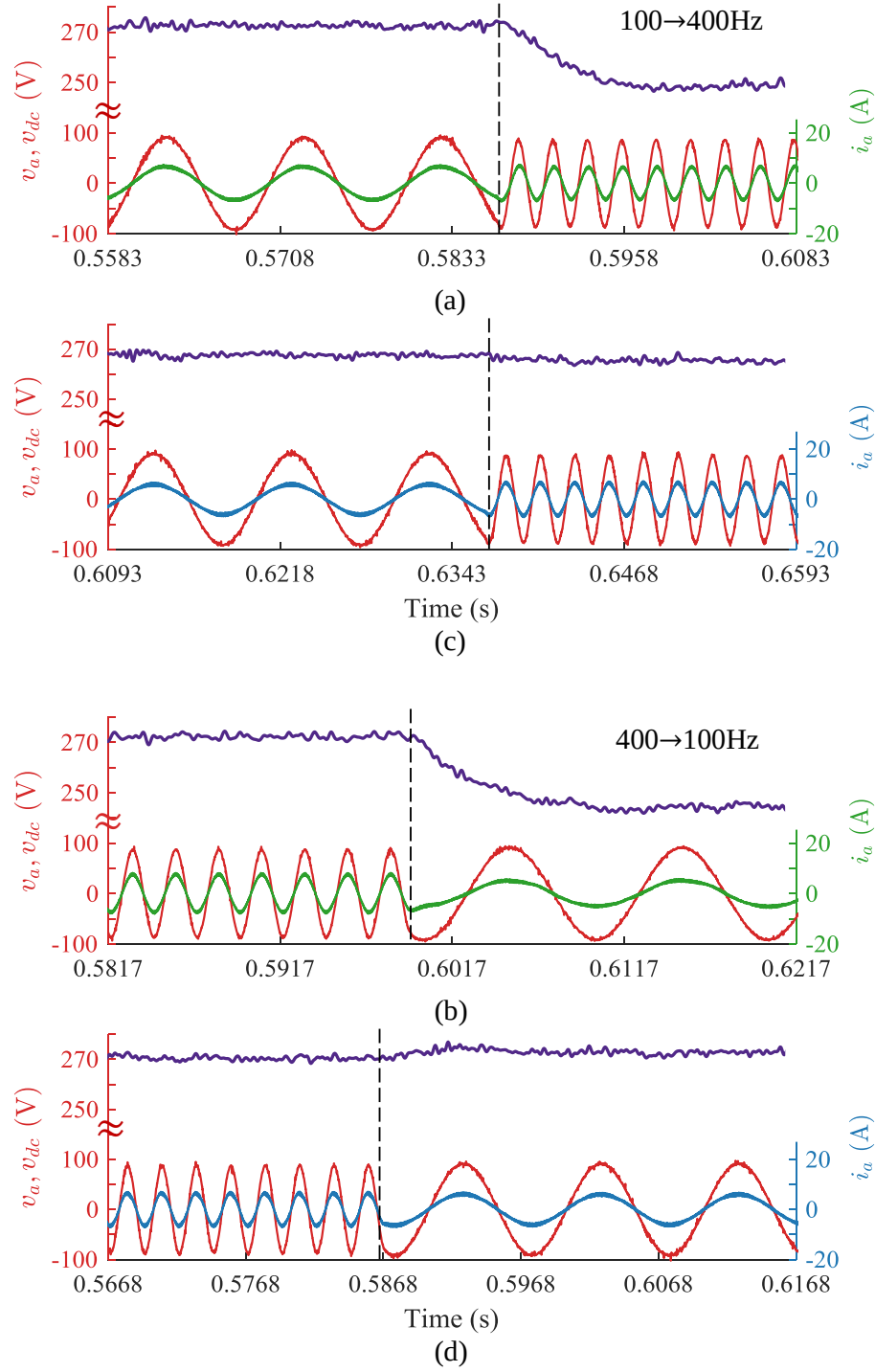


Figure 4.20: Experimentally obtained dynamic performance under a step-change in the supply frequency; (a) model-based dq -controller, 100 to 400 Hz, (b) model-based dq -controller, 400 to 100 Hz, (c) MRAC, 100 to 400 Hz, (d) MRAC, 400 to 100 Hz [103].

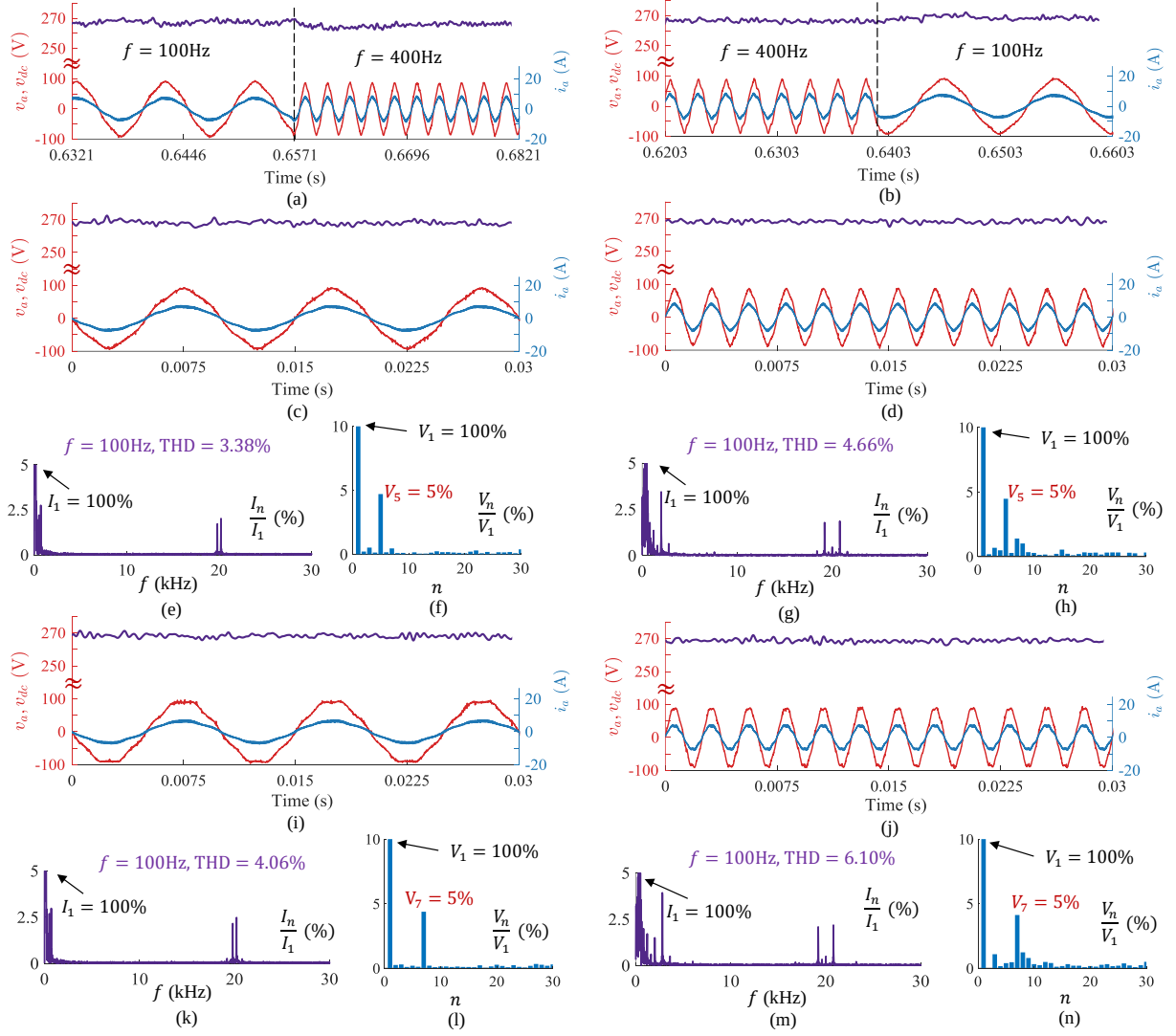


Figure 4.21: Experimentally obtained dynamic and steady-state performance of the developed MRAC under distorted PCC conditions. Dynamic response for a frequency step-change with 5% content of 5th harmonic in the PCC voltage: (a) from 100 to 400 Hz and (b) from 400 to 100 Hz. Steady-state performance with 5% content of 5th harmonic in the PCC voltage: (c) and (d) v_a , i_a , and v_{dc} at 100 Hz and 400 Hz, respectively; (e) and (g) i_a frequency spectrum at 100 Hz and 400 Hz, respectively; (f) and (h) low-order harmonic content of v_a at 100 Hz and 400 Hz, respectively. Steady-state performance with 5% content of 7th harmonic in the PCC voltage: (i) and (j) v_a , i_a , and v_{dc} at 100 Hz and 400 Hz, respectively; (k) and (m) i_a frequency spectrum at 100 Hz and 400 Hz, respectively; (l) and (n) low-order harmonic content of v_a at 100 Hz and 400 Hz, respectively [103].

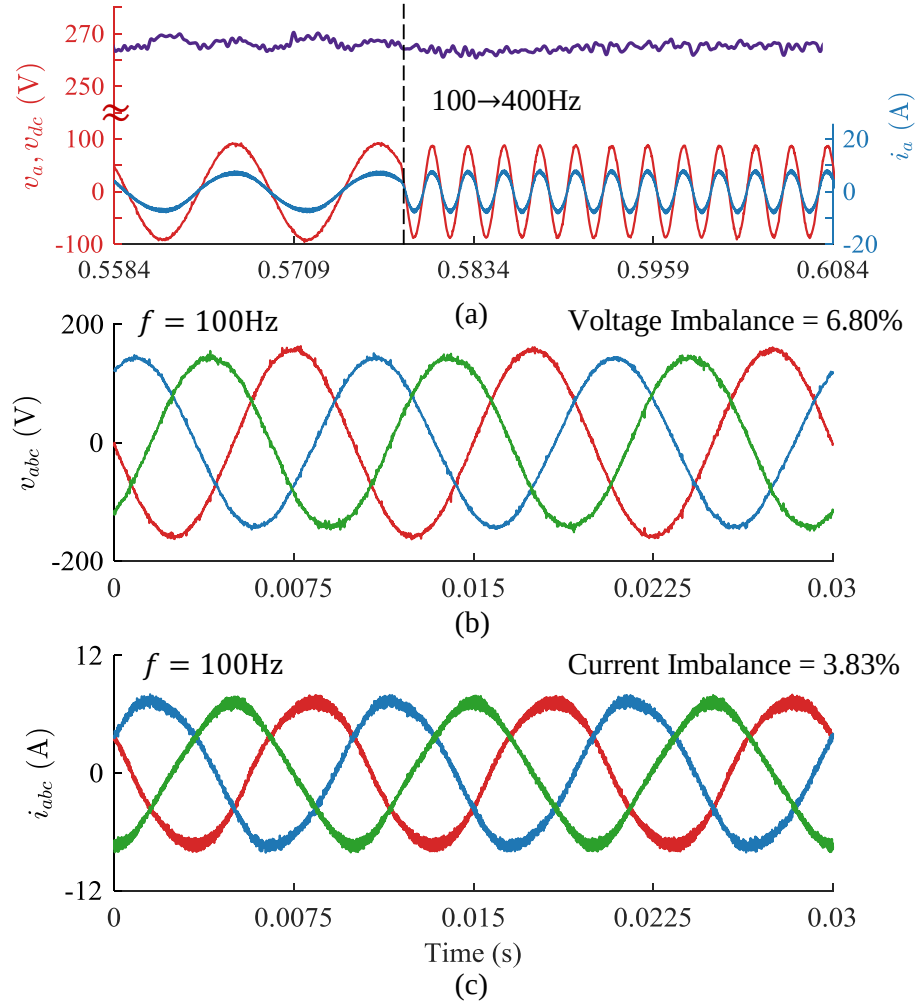


Figure 4.22: Experimentally obtained MARC steady-state and dynamic performance with unbalanced input voltage; (a) dynamic performance for a frequency step-change from 100 $\rightarrow$ 400 Hz, (b) steady-state line-to-line voltage waveforms, and (c) steady-state line current waveforms [103].

As demonstrated in Fig. 4.21, the developed MRAC is able to regulate the dc-bus voltage and maintain a unity power factor under distorted grid voltages. Notice that the THD levels are expected to increase compared to the non-distorted grid case, see Fig. 4.3, due to the high harmonic component introduced in the input voltages. Nonetheless, the currents exhibit relatively low THD values demonstrating the robustness of the developed controller. Similarly, Fig. 4.21 corroborated the seamless dynamics characteristic of the MRAC showing minimum overshoots and settling times for v_{dc} while maintaining a unity power factor with high harmonic distorted input voltages.

Finally, as can be seen in Fig. 4.22, the developed MRAC scheme is able to achieve sound steady-state performance and seamless dynamic under a severe 6.80% input voltage imbalance ($> 5\%$, see IEEE Std. 1159 [151]). More specifically, Fig. 4.22(a) demonstrates the dynamic performance of the controller when the frequency is increased from 100 Hz to 400 Hz in a step-change. On the other hand, Figs. 4.22(b) and 4.22(c) depicts the steady-state performance of the controller. Herein, the input current is maintained below 4%, even when the voltage has a 6.80% imbalance, thus complying with international power quality standards such as the IEEE Std. 1159 [151] and NEMA MG-1 [152].

4.5 Variable-Voltage/Frequency Study Case

The variable-amplitude/frequency voltage input study case emulates a hybrid architecture more electric powertrain whose electric generator is a permanent magnet synchronous machine (PMSG). Herein, the ac-dc power converter or active rectifier is connected to the stator of the PMSG. Consequently, if the electric generator is directly coupled to its prime mover, e.g., jet engine or an internal combustion engine, the PMSG will output a voltage whose amplitude and frequency are proportional to the mechanical speed. More specifically, the amplitude and frequency of the stator voltage maintain a constant V/f ratio given by the physical properties of the machine, such as the number of poles, number of turns per slot, magnet volume, among others.

As described in Section 1.2.1 (1.2.1.1), in some applications, such as MEA, the powertrain electric power generation outputs a voltage with constant amplitude but variable-frequency. However, in the automobile industry, PM synchronous machines are implemented as both generators and motors. Furthermore, it is expected that, due to their superior efficiency and power density, PMSG will replace other machines in modern more electric powertrains. As a result, in this section, the dynamic performance of all the developed control schemes described in Chapter 3 was evaluated for a variable-amplitude/frequency voltage input compared to the benchmark model-based dq -controller. The objective of this section is to experimentally validate the controllers developed in this work under a variable-

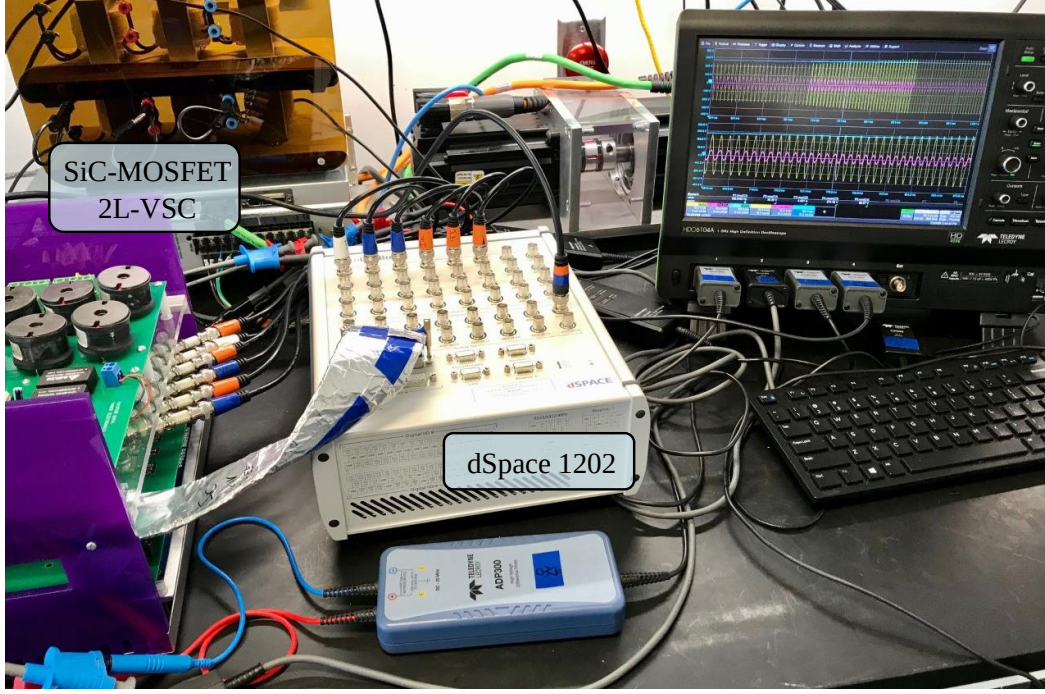
amplitude/frequency voltage input, which is set to be the case for future more electric power-trains that may adopt a PM synchronous machines as their electric generators. With this in mind, each controller was tested with a voltage/frequency ramp and step increase/decrease. Notice that although a voltage/frequency step-change is not physically achievable by a primer mover, this scenario was considered to assess the converter's dynamic performance under an extreme case. Lastly, it is important to mention that the steady-state performance and current THD were not studied in this section since only the dynamic performance will be affected when the amplitude and frequency of the input voltage are changed simultaneously.

4.5.1 Hardware Setup

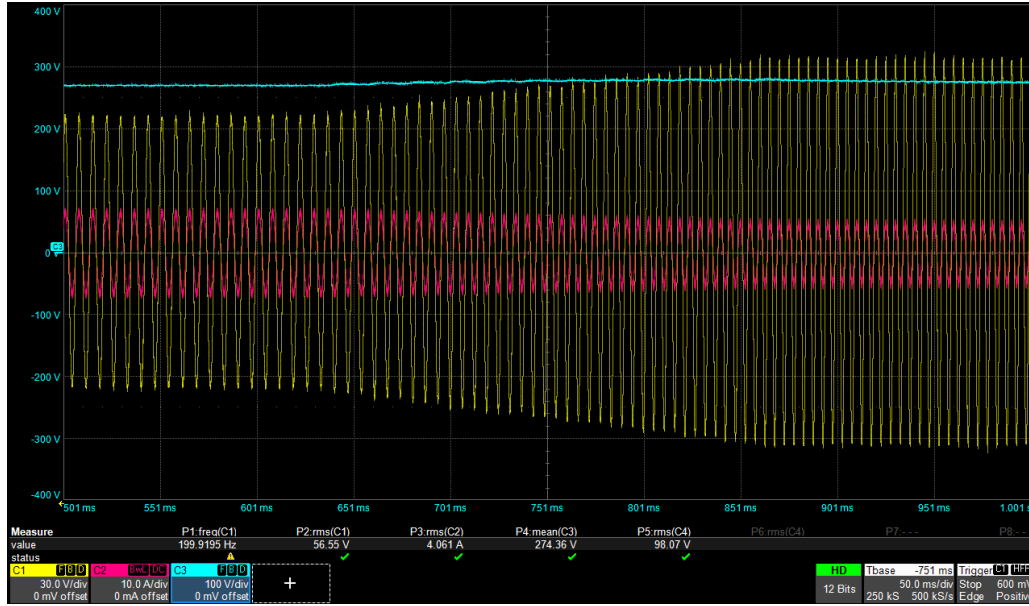
The hardware setup implemented for the variable-amplitude/frequency voltage input study case is depicted in Fig. 4.23. Due to hardware limitations, all the test cases were tested between 140 – 200 Hz / 80.5 – 115 V with a 500 W equivalent resistive dc-load. The voltage amplitude/frequency variations were performed at a constant ratio of $V/f = 115/200 = 0.575 \text{ V/Hz}$ by a grid emulator. The converter was set to operate at unity power factor, i.e., $Q_{ref} = 0 \text{ VAR}$ and $V_{dc}^{ref} = 270 \text{ V}$. In all cases, the gains and parameters of the tested controllers, as well as their hardware implementation were set to the same values described throughout this chapter. Similarly, all experiments were carried out using the same converter described in Section 4.1.

4.5.2 Dynamic Performance Under Voltage/Frequency Ramps

To verify the dynamic performance of the developed controllers, acceleration and deceleration frequency ramps were applied to the system. The frequency was ramped from 140 – 200 Hz and vice-versa at voltage/frequency ratio of $V/f = 0.575 \text{ V/Hz}$, which proportionally translated to 80.5 – 115 V and 115 – 80.5 V linear voltage variations. The slope of the frequency ramps was set to $\pm 250 \text{ Hz/s}$ as recommended by the Aircraft Electric Power Characteristics MIL-STD-704 standard [147].



(a)



(b)

Figure 4.23: Experimental validation for the variable-voltage/frequency study case; (a) hardware setup, and (b) oscilloscope screenshot of the MRAC scheme under a V/f ramp increase.

Similar to the variable-frequency study case presented in Section 4.4, the model-based dq -controller was used as the benchmark control scheme. More specifically, the gains ωL_f and $-\omega L_f$, shown in Fig. 2.12, were set for two distinct tuning cases:

- i. Constant $\pm\omega L_f$: the gains ωL_f and $-\omega L_f$ remain fixed throughout any frequency excursion using a frequency of $f = 140$ Hz.
- ii. Variable $\pm\omega L_f$: the gains ωL_f and $-\omega L_f$ are dynamically adjusted using the frequency f estimated by the direct-phase angle detection scheme shown in Fig. 2.9(c).

On the other hand, the indirect adaptive step-ahead and seamless dynamics direct model reference adaptive control schemes were implemented in the hardware with the parameters and settings described in Sections 4.3 and 4.4, respectively.

For each study case, the ac-voltages and currents shown in this section were captured using a Teledyne LeCroy HDO6104A 1 GHz high definition oscilloscope at a rate of 500 kS/s, whereas the reactive power, dc-bus, and controller gains were recorded at a sampling frequency of 20 kHz using the dSpace ControlDesk acquisition toolbox. In both cases, the collected data was re-plotted to enhance its visibility and presentation clarity. In addition, two limit lines are plotted in every v_{dc} plot as the minimum and maximum voltages permitted by the MIL-STD-704 standard [147].

Fig. 4.24 shows the dynamic performance of the model-based dq -controller when the frequency is increased and decreased at ± 250 Hz/s from 140 – 200 Hz and vice-versa at voltage/frequency ratio of $V/f = 0.575$ V/Hz. More specifically, Figs. 4.24(a) and 4.24(b) represent case (i), whereas case (ii) is depicted in Figs. 4.24(c) and 4.24(d). In both cases, the model-based dq -controller exhibits high overshoots for v_{dc} and Q , with v_{dc} exceeding the maximum voltage permitted by the MIL-STD-704 standard. That being said, case (ii) improves the v_{dc} dynamics, while slightly increasing the Q overshoots.

As depicted in Fig. 4.25, the step-ahead control scheme outperforms the model-based dq -controller in terms of the v_{dc} and Q overshoots, maintaining the dc-bus voltage between the aviation standard permissible limits during the voltage/frequency excursions. In addition, the step-ahead controller demonstrated approximately zero overshoot in reactive power. Nonetheless, the previously mentioned control scheme exhibits a reactive power steady-state tracking error due to its model predictive nature, and the lack of a Q error minimization technique, e.g., integrator. This error is present even with the implementation of the adap-

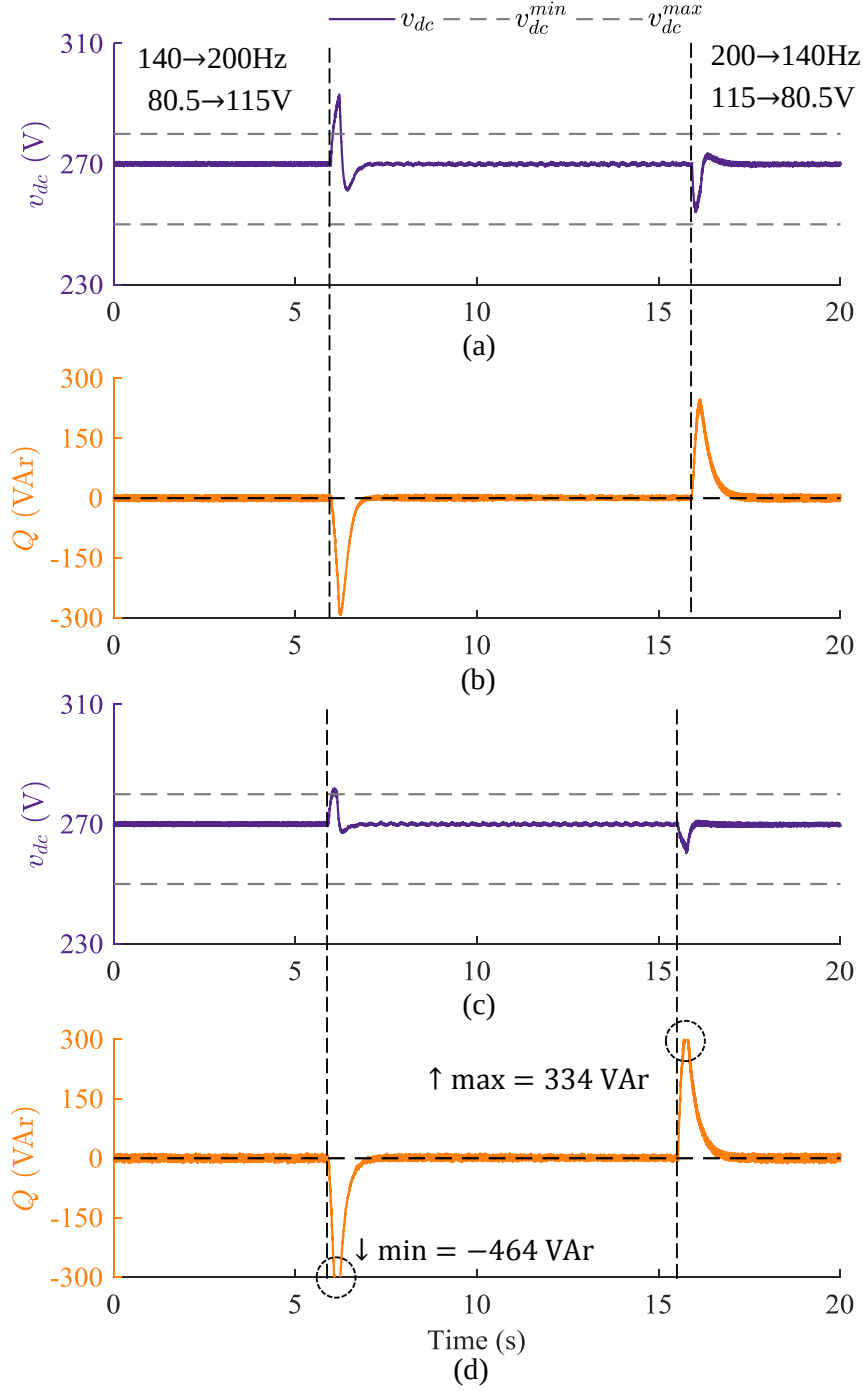


Figure 4.24: Experimentally obtained dynamic performance of the model-based dq -controller when the frequency is increased and decreased at $\pm 250 \text{ Hz/s}$ with $V/f = 0.575 \text{ V/Hz}$; (a) v_{dc} and (b) reactive power Q regulation tuned for 150 Hz, (c) v_{dc} and (d) reactive power Q regulation with the controller with variable frequency tuning.

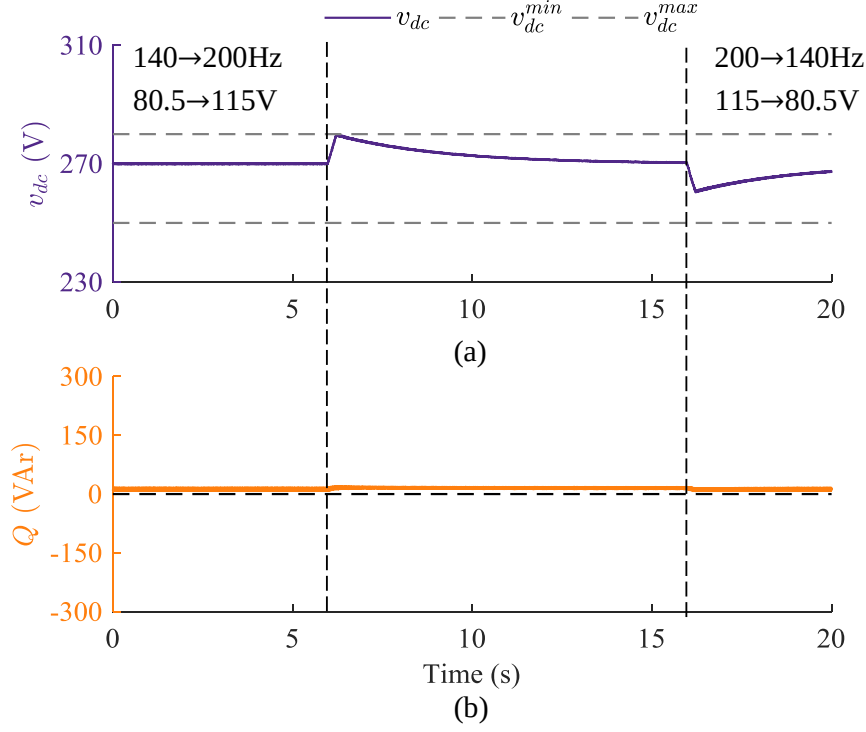


Figure 4.25: Experimentally obtained dynamic performance of the step-ahead controller when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz; (a) v_{dc} and (b) reactive power Q regulation tuned.

tive filter parameter estimation technique described in Section 3.2 due to the additional measurements and computational errors that are inherent in any hardware implementation. Furthermore, due to the PI-loop required to regulate v_{dc} , an increase in the settling time of this control variable can be seen in Fig. 4.25(a).

In contrast with the step-ahead and model-based dq -controllers, the seamless direct MRAC shows minimum overshoots, lower settling times, and zero steady-state errors for v_{dc} and Q . This seamless dynamics behavior is depicted in Fig. 4.26. Here, the dc-bus voltage is kept inside the permissible limits, whereas the reactive power exhibits minimum overshoots. On the other hand, Figs. 4.26(c) and 4.26(d) display the dynamics of two adaptive gains, i.e., $\hat{K}_{12}$, and $\hat{K}_{22}$. These two gains were selected as examples to demonstrate the adaptive gain adjustment as part of the direct MRAC scheme when the voltage/frequency of the input change.

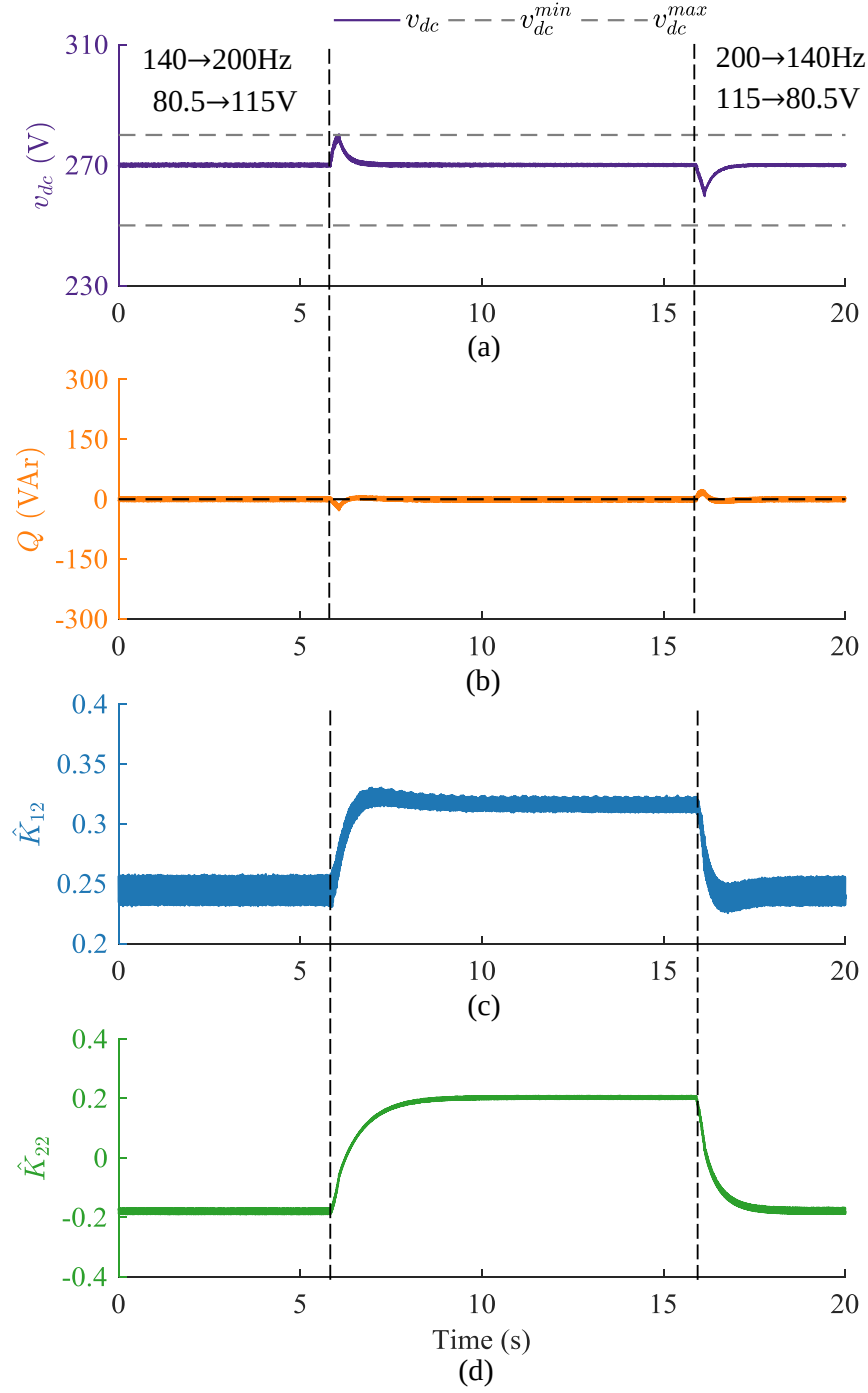


Figure 4.26: Experimentally obtained dynamic performance of the MRAC when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz; (a) v_{dc} , (b) reactive power Q , (c) adaptive gains $\hat{K}_{12}$, and (d) $\hat{K}_{22}$.

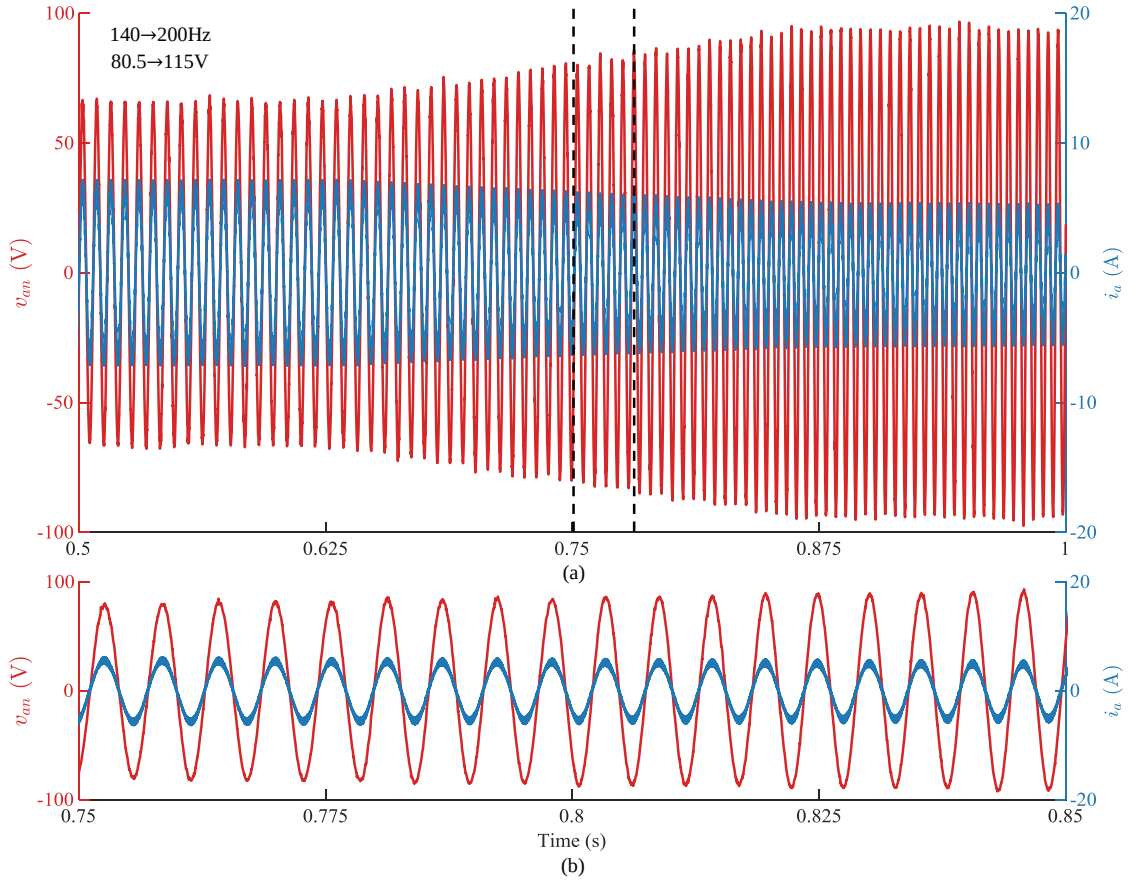


Figure 4.27: Experimentally obtained line-to-neutral voltage and line current for the MRAC when the frequency is increased from 140 – 200 Hz at ± 250 Hz/s with $V/f = 0.575$ V/Hz; (a) full transient view, and (b) zoomed in view as the frequency and voltage increase.

Lastly, Figs. 4.27 and 4.28 show the line-to-neutral voltage v_a and line current i_a for the MRAC when the frequency is increased from 140 – 200 Hz at ± 250 Hz/s with $V/f = 0.575$ V/Hz, and decreased from 200 – 140 Hz, respectively. These figures are a subset of the transients presented in Fig. 4.26. In Fig. 4.27, the proportional increase of the voltage is observed as the frequency increases; whereas, the line current decreases with voltage since the dc-bus voltage remained regulated and with the same resistive load. A similar behavior is observed in Fig. 4.28 when the frequency is decreased. In both cases, the voltage and current stay in phase, which demonstrates the unity power factor regulation achieved by the controller during the frequency/voltage ramping transients, see Figs. 4.27(b) and 4.28(b).

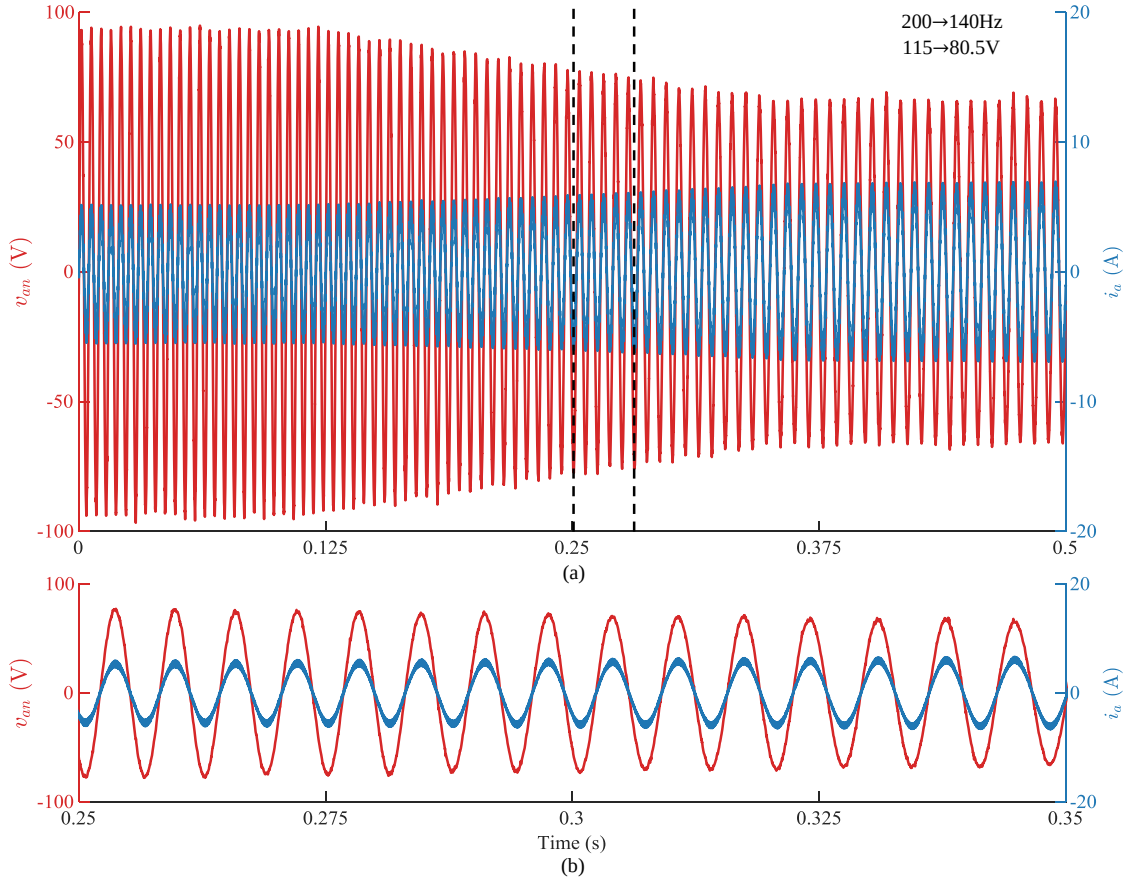


Figure 4.28: Experimentally obtained line-to-neutral voltage and line current for the MRAC when the frequency is decreased from 200 – 140 Hz at ± 250 Hz/s with $V/f = 0.575$ V/Hz; (a) full transient view, and (b) zoomed in view as the frequency and voltage decrease.

4.5.3 Dynamic Performance Under Voltage/Frequency

Step-Changes

To further investigate the dynamic performance of the controllers, a frequency/voltage-amplitude step-change was applied to the system. This step increase/decrease in frequency and voltage represents an artificial test case, intended only to demonstrate the efficacy of the MRAC under extreme frequency/voltage variations. Notice that, in a more electric powertrain, the variation in the frequency and voltage generated by a permanent magnet synchronous generator is dictated by the equivalent inertia of the mechanical system.

Fig. 4.29 depicts the dynamic performance under a step-change in the supply frequency and voltage amplitude with $V/f = 0.575$ V/Hz for the the seamless direct MRAC and model-

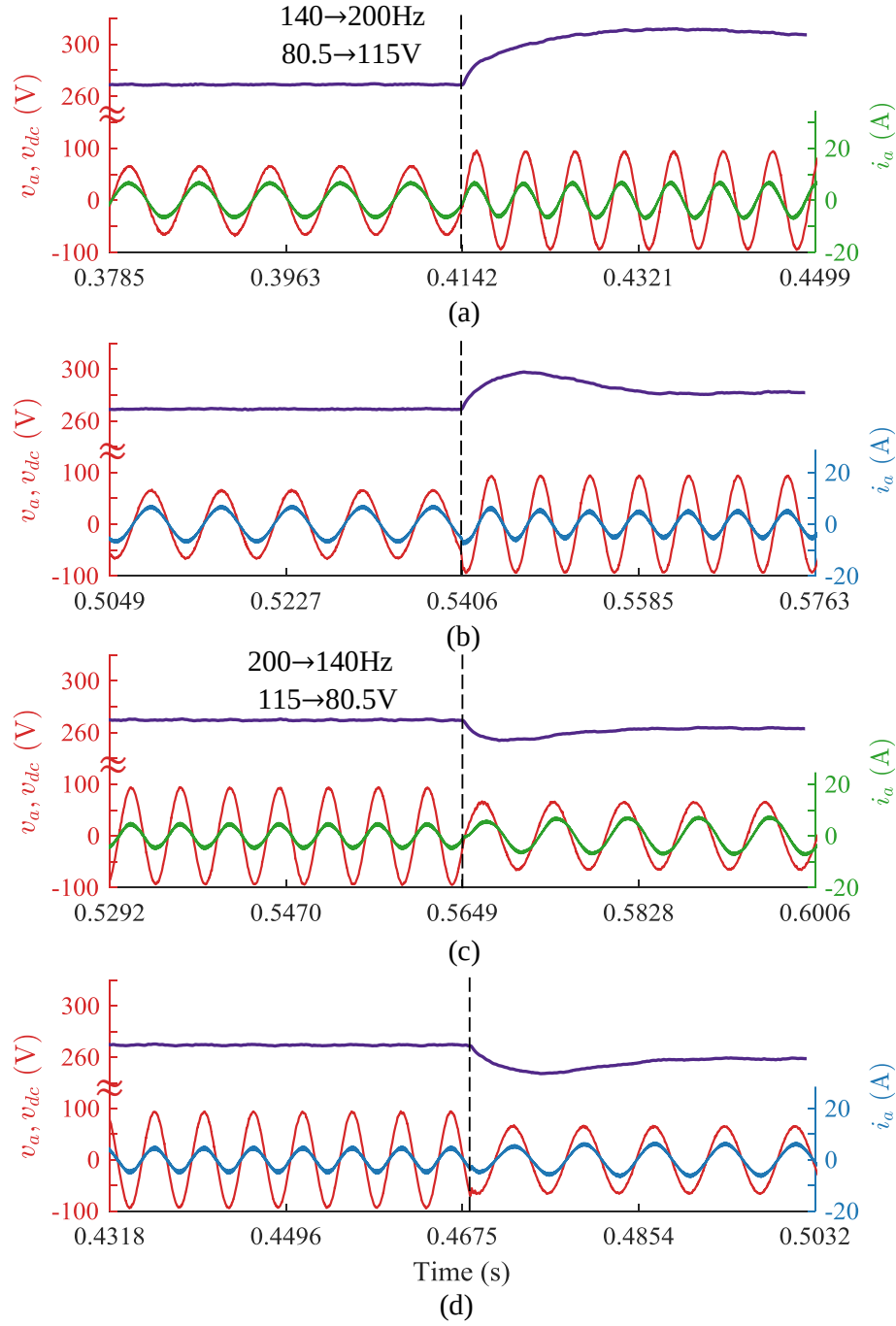


Figure 4.29: Experimentally obtained dynamic performance under a step-change in the supply frequency and voltage amplitude with $V/f = 0.575$ V/Hz; (a) model-based dq -controller, 140 to 200 Hz, (b) model-based dq -controller, 140 to 200 Hz, (c) MRAC, 140 to 200 Hz, (d) MRAC, 200 to 140 Hz.

based dq -controllers. More specifically, Figs. 4.29(a) and 4.29(c) show the step increase and decrease for the model-based dq -controller with variable frequency tuning, i.e., case (ii);

whereas, Figs. 4.29(b) and 4.29(d) display the equivalent data for the MRAC. In all figures, the phase-to-neutral v_a , line current i_a , and dc-bus voltage v_{dc} are plotted to demonstrate the dynamics of each variable. As can be seen in Fig. 4.29, the MRAC outperforms the model-based dq -controller in the dc-bus overshoots, settling time, and the unity power factor regulation. In both, the increase and decrease frequency/voltage step-changes, the MRAC maintains a lower v_{dc} overshoot and settling time. In addition, the voltage and current are in phase at all times, thus guaranteeing a unity power factor operation.

4.5.4 Dynamic Performance Indicators

Tables 4.15 and 4.16 outline the dynamic performance indicators for all the control schemes tested under the variable-amplitude/frequency study case. Table 4.15 displays the dc-bus voltage v_{dc} and reactive power Q minimum and maximum overshoots when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz. Herein, the step-ahead controller showed the lowest overshoots in both v_{dc} and Q , followed by the seamless dynamics direct MRAC; and ending with the variable ω tuning model-based dq -controller, which generated the higher overshoots.

Similarly, Table 4.16 outlines the settling times and steady-state-errors (SSE). As can be seen from the previously mentioned table, all controllers exhibit similar settling times except the strap-ahead controller. In addition, the step-ahead controller maintained an average SSE throughout the voltage/frequency excursion of approximately 14 VAr. Based on Table 4.15, Table 4.17 outlines the percent overshoot for each controller using $v_{dc}^{ref} = 270$ V. Once

Table 4.15: Experimentally obtained minimum and maximum reactive power and dc-bus voltage overshoot values for the MRAC, step-ahead, and model-based dq -controller when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz

Control Scheme	Tuning	Max. Value		Min. Value	
		Q (VAr)	v_{dc} (V)	Q (VAr)	v_{dc} (V)
Model-based dq -controller	Constant ω ($f = 150$ Hz)	247.83	293.10	-293.12	254.05
	Variable ω	334.33	281.97	-464.04	260.32
Step-Ahead	Variable ω	19.22	279.91	7.52	260.38
MRAC	Adaptive	23.23	280.22	-27.84	259.72

Table 4.16: Experimentally obtained settling time and steady-state error (SSE) reactive power and dc-bus voltage overshoot values for the MRAC, step-ahead, and model-based dq -controller when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz

Control Scheme	Tuning	Settling Time	SSE	
		v_{dc}	Q (VAr)	v_{dc} (V)
Model-based dq -controller	Constant ω ($f = 150$ Hz)	0.2960 s	≈ 0	≈ 0
	Variable ω	0.2650 s	≈ 0	≈ 0
Step-Ahead	Variable ω	2.3470 s	14.03	≈ 0
MRAC	Adaptive	0.3310 s	≈ 0	≈ 0

again, the step-ahead controller outperforms the rest of the controller with approximately 3.5% overshoots, followed by the MRAC with 3.8%. Moreover, compared to the constant ω tuning model-based dq -controller, both the step-ahead and MRAC schemes accomplished over 90% reduction in the reactive power overshoots, as demonstrated in Table 4.18.

In conclusion, it is clear from the results exposed in section, that the outcomes presented for variable-frequency can be extended to a variable-amplitude/frequency case. As a result, due to its low overshoots, settling times, and steady-state error, the MRAC scheme proved its effectiveness by outperforming the step-ahead and model-based dq -controllers.

Table 4.17: Experimentally obtained dc-bus percent overshoot for the MRAC, step-ahead, and model-based dq -controller when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz

Control Scheme	Tuning	Max. Value	Min. Value
Model-based dq -controller	Constant ω ($f = 150$ Hz)	8.56%	-5.91%
	Variable ω	4.43%	-3.59%
Step-Ahead	Variable ω	3.67%	-3.56%
MRAC	Adaptive	3.79%	-3.81%

Table 4.18: Experimentally obtained reactive power tracking error percent reduction for the MRAC, step-ahead, and model-based dq -controller when the frequency is increased and decreased at ± 250 Hz/s with $V/f = 0.575$ V/Hz

Control Scheme	Tuning	Max. Value	Min. Value
Model-based dq -controller	Constant ω ($f = 150$ Hz)	—	—
	Variable ω	22.09%	58.31%
Step-Ahead	Variable ω	-92.98%	-102.57%
MRAC	Adaptive	-91.52%	-90.50%

4.6 Summary

In summary, this chapter presented the experimental results for the step-ahead controller, the indirect adaptive step-ahead controller, and the seamless dynamics direct MRAC scheme under the context of a more electric powertrain with variable-frequency/constant amplitude voltage, variable-frequency/amplitude voltage, and system parameter variations. In the next chapter, conclusions and future work are presented.

Chapter 5

Conclusion and Future Work

In this chapter, the summary of the accomplishments and main contributions of this dissertation are reviewed, followed by suggestions regarding possible directions of future research topics.

5.1 Summary and Conclusions

Modern transportation systems are shifting toward more electrification technologies to address fuel efficiency and emission requirements, as well as market demands for lower operational costs. At the center of this technological transition, power electronic converters take the spotlight as the key enabling technology. Led by the advances in power electronics, full-electric railroad systems, and hybrid/all-electric vehicles, e.g., automobiles and aircraft, are now a reality. However, the full integration of all-electric vehicles in modern societies is bounded by the current state-of-the-art electrical energy storage systems. This limitation shines a light in hybrid architectures that can enable fuel and cost savings while providing similar autonomy and transportation capabilities as modern fuel-based vehicles. That being said, the heart of a hybrid vehicle is its powertrain. Enabled by power electronic converters, a more electric powertrain uses a combination of energy sources to drive any vehicle efficiently. However, more electric powertrain power converters are subject to rapid input-frequency

variations, load changes, and extreme ambient conditions fluctuations. In this dissertation, a smart three-phase power converter for more electric powertrains has been presented.

At first, an ultrafast step-ahead control scheme for more electric powertrain ac-dc three-phase converters has been presented. Experimental results demonstrate that the developed control technique outperforms the classic model-based dq -controller in terms of both dc-bus voltage regulation and reactive power control. The performance of the two control schemes, i.e., the step-ahead and the model-based dq -controller, has been tested under a ramp and step-change in the supply frequency. Under these scenarios, the step-ahead controller regulates the dc voltage with minimum overshoot and almost instant settling time while also maintaining a unity power. In addition, it has been shown that at a switching frequency of 20 kHz and a filter inductance of 1 mH, the step-ahead controller generates sinusoidal input currents with a THD $< 1.7\%$ at 1 kW. Altogether, the step-ahead control scheme regulates the dc-bus voltage with sinusoidal input currents at unity power factor under variable-frequency by combining the step-ahead model predictive algorithm with the instantaneous direct phase-angle detection technique. Both the simplicity of the step-ahead algorithm and the low computational requirements and accuracy of the instantaneous direct phase-angle detection make the step-ahead controller attractive for high switching frequencies, which allows the size reduction of the filter components. Moreover, the near-zero delay in the frequency estimation provided by the instantaneous direct phase-angle detection, and the fast execution of the step-ahead algorithm allows the control scheme to achieve an ultrafast dynamic performance even under the extreme case of a step-change of the supply frequency.

Secondly, the dynamic performance under parameter mismatch of the step-ahead control scheme has been analyzed and adequately addressed. In this regard, it has been experimentally proven that the reactive power tracking error increases proportionally with the input frequency, thus affecting the steady-state and dynamic performance of the model predictive step-ahead controller. To address this, a Lyapunov-based adaptive parameter estimation technique has been developed. Subsequently, the effectiveness of the step-ahead control scheme with the adaptive parameter estimation technique has been validated using fre-

quency acceleration and deceleration input ramps, as well as by intentionally increasing the inductance of the filter in the hardware employing an external circuit breaker. In both cases, it has been shown that by implementing the developed adaptive parameter estimation, the performance of the controller improved, which translated into a satisfactory regulation of the dc-bus output voltage and unity power factor. Finally, it has been proven that the developed adaptive parameter estimation technique offers fast and accurate convergence with minimum overshoot and settling time in the parameter estimation. These performance achievements can be extended to higher power rated converters and increased dc-bus voltages with the proper tuning the v_{dc} PI control loop gains, thus paving the way for smarter converters capable of adaptively tracking the variations in their filter parameters.

Despite its sound performance and enhanced capabilities, the step-ahead control scheme with adaptive parameter estimation shows a reactive power steady-state tracking error proportional to the input frequency. Consequently, a seamless dynamics MRAC scheme for more electric powertrain applications has been developed. The developed MRAC has been implemented in a 2L-VSC, whose main objective is to regulate the dc-bus voltage and operate at unity power factor while drawing sinusoidal currents from the PCC. Experimental testing has been carried out to demonstrate the superiority of the MRAC using a classical PI model-based dq -controller as a benchmark for comparison, similar to the benchmarking process performed with the step-ahead controller. The steady-state performance of the model-based dq -control and MRAC schemes has been investigated at different input frequencies, whereas the transient performance has been studied under frequency step-changes. Based on the experimental results obtained in this dissertation, it can be concluded that both controllers offer similar steady-state performance and current THDs. Both control schemes effectively regulate the dc-bus voltage while maintaining a unity power factor with $< 5\%$ ac-side current THD.

In terms of the dynamic performance, the developed MRAC accomplishes seamless dynamics under frequency step-changes, whereas the classic model-based dq -control scheme fails in providing an acceptable dynamic performance regardless of its tuning settings. Furthermore, the seamless dynamics performance is maintained under distorted and unbalances

input voltages. All the previously mentioned advantages of the developed MRAC place this controller as a superior alternative for more electric powertrains based on the control schemes developed and investigated in this work.

Lastly, to further validate the concepts introduced in this dissertation, all the developed controllers have been tested under variable-frequency/variable-amplitude input voltage, which emulates a more electric powertrain variable-speed permanent magnet generator directly connected to its primer mover, e.g., internal combustion engine or jet engine. Herein, it is clear from the experimental results, that the outcomes presented for variable-frequency can be extended to a variable-amplitude/frequency case. As a result, due to its low overshoots, settling times, and steady-state error, the MRAC scheme proved its effectiveness by outperforming the step-ahead and model-based dq -controllers, thus solidifying the developed smart converter concept for more electric powertrains.

5.2 Future Work

This work has advanced the state-of-the-art in variable-frequency/variable-amplitude voltage ac-dc power converters for more electric powertrains. While the efficacy of the control schemes presented in this dissertation has been thoroughly validated, there are opportunities to build upon this work.

One initiative can focus on the hardware development of the smart converter itself using state-of-the-art GaN semiconductors. This path has been initiated in Appendix C and can be continued by focusing on the efficiency, power density, and specific energy of the new generation of GaN-based smart converters, thus making this technology attractive for the automobile and aerospace industry due to its size, weight, and temperature profile. Furthermore, the next generation converter can be designed to work at 540V, which is set to become the new standard dc-bus voltage in future aircraft power systems.

A more comprehensive advance in the research can be achieved by focusing on the combined design of the electric generator and GaN-based power converter as a whole. Once again, the focus can be on efficiency, power density, and specific power. Due to its high-

temperature properties, the GaN power converter can share the same housing or structure as the electric machine. In this scenario, the thermal management system of both apparatuses becomes of pivotal importance, thus becoming another significant research area.

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Appendix A

Arbitrary Reference Frame Theory for Power Converters

The reference-frame theory, initially applied to electric machines, dates back to the 1920s when R.H. Park introduced it for machine modeling analysis in [153]. Since then, the so-called Park's transformation revolutionized electric machine analysis by eliminating the rotor-position dependence inductances from the voltage equations of a synchronous machine. This unique property is accomplished by transforming or referring the stator variables to a frame of reference fixed in the rotor [122].

Park's transformation opened the door for multiple researchers that started particularizing or even generalizing the reference-frame theory to certain electric machines or power system apparatuses. H.C Stanley implemented a transformation applicable to symmetrical three-phase induction machines, in which the rotor variables are transformed to a frame of reference fixed in the stator [154]. Similarly, G. Kron introduced a transformation for the same type of machine that transformed both the stator and rotor variables to a reference frame rotating in synchronism with the fundamental angular velocity of the stator variables [155]. Conversely, D.S Brereton et al. employed a changed of variables for induction machines that transformed the stator variables to a reference frame rotating at the electrical angular velocity of the rotor – essentially Park's transformation for induction machines [156].

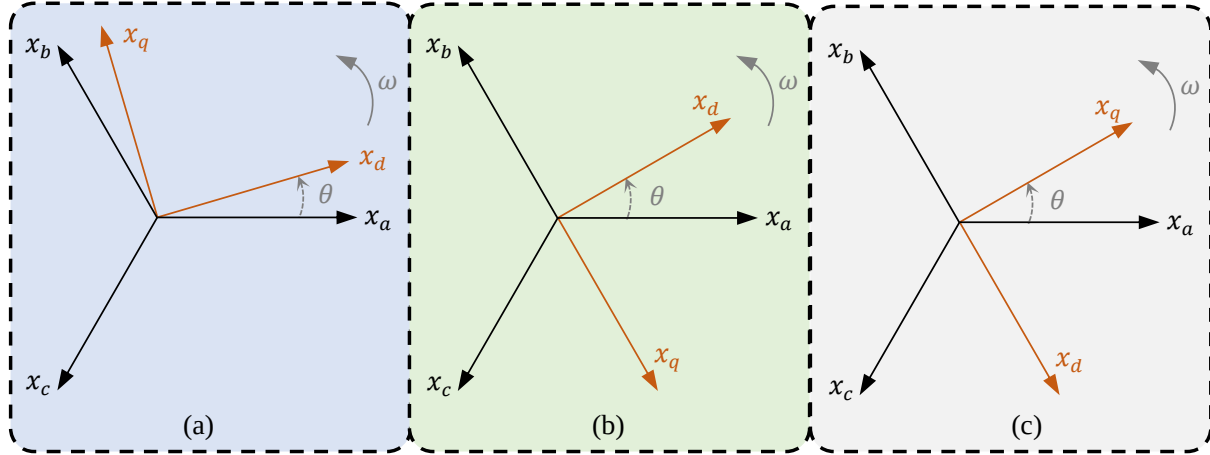


Figure A.1: Vector diagrams for $dq0$ synchronous frame of reference transformations; (a) $dq0$ original transformation, (b) $dq0$ transformation with d -axis leading, and (c) $dq0$ transformation with q -axis leading [1-4].

It was not until 1965, in [157], that all of the separately developed transformations for induction machines were unified by simply assigning the appropriate speed of rotation of the reference frame – i.e., the so-called *arbitrary reference frame*. Later, the same theory was extended to synchronous machines in [158].

A.1 Arbitrary Reference-Frame Theory Applied to Power Electronics

Although developed for electric machines, the simplifications that the arbitrary reference-frame theory brought to three-phase systems soon became attractive for power electronics applications. Naturally, these applications started with motor vector control [159] and then extended to power converters [140]. Since then, many researchers have adopted the also-called $dq0$ transformation for grid-tied and stand-alone power electronic-based applications.

Nowadays, there are mainly three conventional transformations currently being used in the literature, see Fig. A.1:

- The original Park's transformation has the q -axis leading the d -axis, with the angle θ defined between the d -axis and the reference axis a , see Fig. A.1(a).

- The $dq0$ transformation with the d -axis leading the q -axis, and the angle θ between the d -axis and the reference axis a , see Fig. A.1(b).
- The $dq0$ transformation with the q -axis leading the d -axis, and the angle θ between the q -axis and the reference axis a , see Fig. A.1(c).

In general, any control scheme can be developed or adapted to any of the aforementioned transformation variations. Depending on the convention used, the “ d ” or “ q ” components are tied to a specific control variable. For example, when using the $dq0$ transformation with the q -axis leading the d -axis, and the angle θ between the q -axis and the reference axis a , see Fig. A.1(c), represented by

$$\begin{bmatrix} x_q \\ x_d \\ x_0 \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos(\theta) & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta - \frac{4\pi}{3}\right) \\ \sin(\theta) & \sin\left(\theta - \frac{2\pi}{3}\right) & \sin\left(\theta - \frac{4\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix}, \quad (\text{A.1})$$

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos(\theta) & \sin(\theta) & 1 \\ \cos\left(\theta - \frac{2\pi}{3}\right) & \sin\left(\theta - \frac{2\pi}{3}\right) & 1 \\ \cos\left(\theta - \frac{4\pi}{3}\right) & \sin\left(\theta - \frac{4\pi}{3}\right) & 1 \end{bmatrix} \begin{bmatrix} x_q \\ x_d \\ x_0 \end{bmatrix}; \quad (\text{A.2})$$

the active and reactive power of a three-phase system can be expressed as:

$$P = \frac{3}{2} (v_d i_d + v_q i_q); \quad (\text{A.3})$$

$$Q = \frac{3}{2} (v_q i_d - v_d i_q). \quad (\text{A.4})$$

Herein, if $\theta = \omega t$ (synchronous reference-frame), where θ is locked to phase-a of a three-phase balanced system with the implementation of a PLL, then $v_d = 0$; thus resulting in

$$P = \frac{3}{2} v_q i_q; \quad (\text{A.5})$$

$$Q = \frac{3}{2}v_q i_d. \quad (\text{A.6})$$

Consequently, a controller can be designed to regulate the active power P with i_q and the reactive power Q with i_d . That being said, utilizing the transformation illustrated in Fig. A.1(b) will swap the dq axes with respect to P and Q . Notice that while both approaches are valid, the convention depicted in Fig. A.1(c) was selected for the modeling and controllers throughout this work. This transformation is the original one proposed in [122, 140, 159, 160].

Lastly, another common transformation derived from (A.2) is the $dq0$ stationary reference-frame. Here, θ is set to zero, i.e., $\theta = 0$; resulting in

$$\begin{bmatrix} x_q \\ x_d \\ x_0 \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix}. \quad (\text{A.7})$$

In the stationary frame of reference, the $dq0$ variables are not constant, as in the synchronous frame of reference. However, this approach still simplifies the dynamic equations in three-phase power systems and electric machines.

Appendix B

A Model Predictive Control Scheme Formulation for Active Rectifiers with *LCL* Filter

In this chapter, an *LCL* filter formulation for the step-ahead controller developed in Chapter 3 is presented [131]. Based on the state-space representation of the *LCL* filter in the dq –stationary frame of reference, two discrete-time transfer functions are derived and then combined to form a novel single discrete-time equation that can predict the grid-side current by only measuring the grid voltages, the grid-side currents, and the dc-bus voltage.

This chapter contains five sections, which start with a background and state-of-the-art. In Section B.2, the state-space model and grid-side current prediction equation are developed in the $dq0$ –stationary frame of reference. In Section B.3, the step-ahead DPC scheme is derived for *LCL* filter implementation. Simulation results demonstrating the steady-state and dynamic performance of the controller, are provided in Section B.4. Finally, concluding remarks are presented in Section B.5.

B.1 Background

The three most common types of filters for power converters are (i) a first-order L filter, (ii) second-order LC filter, and (iii) third-order LCL filter. Among these three filters, the LCL offers higher levels of harmonic attenuation, and it is, therefore, the preferred choice [161]. As a result, most PI-based controllers use LCL filters in commercial applications. However, due to their third-order topology and resonance problems, LCL filters are rarely implemented with MPC schemes. The main reasons LCL filters are rarely implemented with MPC schemes is that they require additional sensors for the converter-side currents and/or capacitor voltage drop and that in some variable-switching frequency MPC techniques, e.g., Finite-Control-Set MPC (FCS-MPC), the resonance frequency of the filter may be excited at some operating point causing a substantial deterioration of the power quality of the system and jeopardizing its stability. Consequently, for applications where a fast dynamic response is required with improved filtering, a robust MPC scheme that can operate effectively with an LCL filter is highly desirable.

In [162], a model predictive direct current control (MPDCC) strategy is proposed for a medium-voltage (MV) neutral-point-clamped converter with an LCL filter. In the aforementioned work, the authors implement a FCS-MPC technique, which generates a variable-switching-frequency that can excite the resonance frequency of the LCL filter. Consequently, the MPDCC requires active damping by means of a virtual resistor. Similarly, in [163], an online optimized FCS-MPC for a grid-connected inverter with LCL filters based on different state feed-backs is presented to minimize switching losses and achieve acceptable current distortion. On the other hand, in [164], the authors use active damping for a hysteresis MPC grid-tied inverter with an LCL filter. Based on [163], the authors in [165] make use of multi-variable system control in the FCS-MPC cost function to successfully control the current. A common characteristic in many of the aforementioned state-of-the-art controllers is that they operate the converters at variable switching frequency, thus requiring active damping techniques. Recently, the authors in [166] propose a continuous-set MPC (CS-MPC) with active damping and embedded integrator for a three-phase voltage-source inverter. By im-

plementing a CS-MPC scheme, the converter can generate a constant switching frequency. However, even when considering constant switching frequency MPC schemes, e.g., CS-MPC and deadbeat [134, 135, 167], the state-of-the-art model predictive controllers designed for converters with input *LCL* filters require the measurement of the capacitor voltage and/or the converter-side current, thus adding sensors to the system, which increases its complexity and reduces its reliability.

B.2 System Description & Modeling

In this section, the model of the *LCL* filter is developed. First, the state-space model is presented based on the KVL equations of the system in *abc*—coordinates. Subsequently, the grid-side current prediction formula is derived after discretizing the continuous state-space model previously presented.

B.2.1 State-Space Model of an *LCL* Filter

The general structure of a 2L-VSC active rectifier is shown in Fig. B.1. This three-phase ac-dc converter is connected to the grid at the PCC through an *LCL* filter, whereas the output of the converter supplies a dc-load connected in parallel with a filtering dc-bus capacitor. The system is modeled in the *dq0*—stationary frame of reference by means of

$$\vec{x} = \frac{2}{3} (x_a + \alpha x_b + \alpha^2 x_c) = x_q - jx_d, \quad (\text{B.1})$$

where $\alpha = e^{j\frac{2\pi}{3}}$ [122]. Using this modeling approach, the dynamics of the grid-side current, converter-side current, and voltage drop on the filter's capacitance can be described by the following linear time-invariant state-space system:

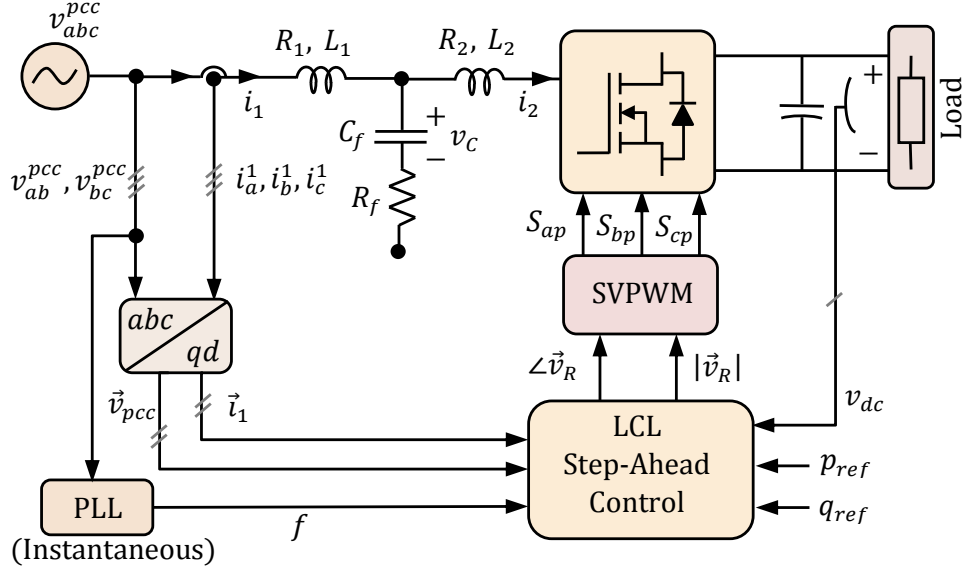


Figure B.1: Detailed block diagram of the step-ahead control scheme for an active rectifier with an *LCL* Filter [131].

$$\begin{aligned} \frac{d}{dt} \begin{bmatrix} \vec{i}_1 \\ \vec{i}_2 \\ \vec{v}_c \end{bmatrix} &= A \begin{bmatrix} \vec{i}_1 \\ \vec{i}_2 \\ \vec{v}_c \end{bmatrix} + B \begin{bmatrix} \vec{v}_{pcc} \\ \vec{v}_R \end{bmatrix} \\ \vec{i}_1 &= C \begin{bmatrix} \vec{i}_1 \\ \vec{i}_2 \\ \vec{v}_c \end{bmatrix} + D \begin{bmatrix} \vec{v}_{pcc} \\ \vec{v}_R \end{bmatrix}; \end{aligned} \quad (\text{B.2})$$

where, matrices A , B , C , and D are defined as follows:

$$A = \begin{bmatrix} -\frac{(R_1+R_f)}{L_1} & \frac{R_f}{L_1} & -\frac{1}{L_1} \\ \frac{R_f}{L_2} & -\frac{(R_2+R_f)}{L_2} & \frac{1}{L_2} \\ \frac{1}{C_f} & -\frac{1}{C_f} & 0 \end{bmatrix}; \quad B = \begin{bmatrix} \frac{1}{L_1} & 0 \\ 0 & -\frac{1}{L_2} \\ 0 & 0 \end{bmatrix}; \quad \begin{aligned} C &= [1 \quad 0 \quad 0] \\ D &= [0 \quad 0] \end{aligned}$$

In (B.2), $\vec{i}_1$ is the grid-side line current space-vector; $\vec{i}_2$ is the converter-side line current space-vector; $\vec{v}_c$ is the space-vector phase-to-neutral voltage drop in the filter's capacitance C_f ; $\vec{v}_R$ is the rectifier phase-to-neutral voltage space-vector; $\vec{v}_{pcc}$ is the PCC phase-to-neutral

Table B.1: $H_{pcc}(s)$ and $H_R(s)$ coefficients [131]

$a_3 = L_1 L_2 C_f$
$a_2 = C_f (R_1 L_2 + R_2 L_1) + R_f C_f (L_1 + L_2)$
$a_1 = R_1 R_2 C_f + L_1 + L_2 + R_f C_f (R_1 + R_2)$
$a_0 = R_1 + R_2$
$b_2 = C_f L_2$
$b_1 = C_f (R_2 + R_f)$
$c_1 = R_f C_f$

voltage space-vector; L_1 is the grid-side inductance of the *LCL* filter, being R_1 its inherent parasitic resistance; L_2 is the converter-side inductance of the *LCL* filter, being R_2 its inherent parasitic resistance; and R_f is the damping resistance of the *LCL* filter. Furthermore, the equations in (B.2) can be written in the classical matrix state-space form:

$$\begin{aligned} \dot{x} &= Ax + Bu \\ y &= Cx + Du \end{aligned}, \quad (\text{B.3})$$

where, the state vector $x = \begin{bmatrix} \vec{i}_1 & \vec{i}_2 & \vec{v}_c \end{bmatrix}^T$, the input vector $u = \begin{bmatrix} \vec{v}_{pcc} & \vec{v}_R \end{bmatrix}^T$, and the output of the system is the grid-side current $y = \vec{i}_1$.

B.2.2 Grid-Side Current Prediction

The step-ahead controller developed in Chapter 3 can be implemented as a current control, direct power control (*PQ*-control) or dc-bus/reactive power control (*QV_{dc}*-control) depending on the application [68]. In all cases, an accurate prediction of the grid-side current is required.

Based on (B.2), the transfer function vector of the system in the Laplace domain can be derived as

$$H(s) = C(sI - A)^{-1}B + D = \begin{bmatrix} H_{pcc}(s) & H_R(s) \end{bmatrix} = \begin{bmatrix} \frac{\vec{i}_1(s)}{\vec{v}_{pcc}(s)} & \frac{\vec{i}_1(s)}{\vec{v}_R(s)} \end{bmatrix}, \quad (\text{B.4})$$

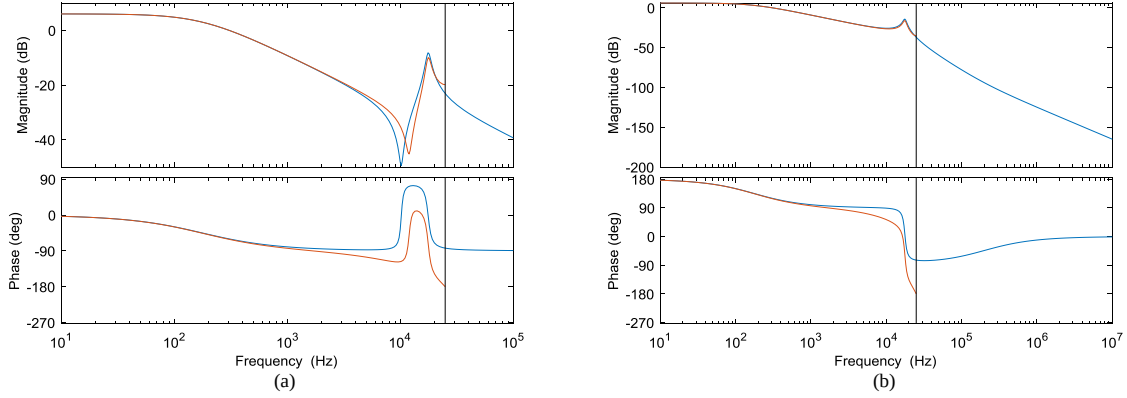


Figure B.2: Bode plots of the continuous- (blue trace) and discrete-time (red trace) transfer functions of the *LCL* filter; (a) $H_{pcc}(z) = \frac{\vec{i}_1(z)}{\vec{v}_{pcc}(z)}$ and (b) $H_R(z) = \frac{\vec{i}_1(z)}{\vec{v}_R(z)}$.

where, $H_{pcc}(s)$ and $H_R(s)$ are a function of the *LCL* filter parameters (i.e., R_1 , L_1 , R_2 , L_2 , R_f , and C_f). More explicitly, $H_{pcc}(s)$ and $H_R(s)$ can be written as:

$$H_{pcc}(s) = \frac{b_2 s^2 + b_1 s + 1}{a_3 s^3 + a_2 s^2 + a_1 s + a_0}, \quad (\text{B.5})$$

$$H_R(s) = \frac{-(c_1 s + 1)}{a_3 s^3 + a_2 s^2 + a_1 s + a_0}. \quad (\text{B.6})$$

Table B.1 summarizes the expressions to compute the coefficients present in (B.5) and (B.6). Once the transfer functions of the *LCL* filter in the s -domain are fully defined, $H_{pcc}(s)$ and $H_R(s)$ must be transformed into the z -domain in order to predict the grid-side current. Accordingly, the zero-order hold transformation with a sampling time T_s can be applied to (B.5) and (B.6); thus resulting in the following transfer function in the z -domain:

$$H_{pcc}(z) = \frac{\beta_0 + \beta_1 z^{-1} + \beta_2 z^{-2}}{z + \alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2}}, \quad (\text{B.7})$$

$$H_R(z) = \frac{\gamma_0 + \gamma_1 z^{-1} + \gamma_2 z^{-2}}{z + \alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2}}. \quad (\text{B.8})$$

Fig. B.2 shows the comparison of the continuous-time transfer functions (B.5) and (B.6)

Table B.2: Coefficients present in (B.10) [131]

$\begin{bmatrix} \alpha_0 & \alpha_1 & \alpha_2 \end{bmatrix} = \begin{bmatrix} 0.0999 & -0.2579 & -0.7788 \end{bmatrix}$
$\begin{bmatrix} \beta_0 & \beta_1 & \beta_2 \end{bmatrix} = \begin{bmatrix} 0.0721 & -0.0089 & 0.0631 \end{bmatrix}$
$\begin{bmatrix} \gamma_1 & \gamma_2 \end{bmatrix} = \begin{bmatrix} -0.0756 & -0.0211 \end{bmatrix}$
$\gamma_0 = -0.0296$

with their discrete-time counterparts (B.7) and (B.8) for an *LCL* filter whose parameters are defined in Table B.3. As can be seen from Fig. B.2, the discrete transfer functions (B.7) and (B.8) mimic the magnitude and phase response of their continuous counterparts accurately up to approximately 2 kHz. The discretization accuracy is reduced as the frequency reaches the Nyquist frequency of the system, i.e. 25 kHz. Nonetheless, from the MPC perspective the transfer functions need to be accurate only around the fundamental frequency and its main harmonics and not up to the Nyquist frequency. This being the case, since (B.4) maintains for the z -domain, the grid-side current space-vector i_1 can be computed using the superposition principle as

$$\vec{i}_1(z) = H_R(z) \vec{v}_{pcc}(z) + H_{pcc}(z) \vec{v}_R(z); \quad (\text{B.9})$$

which, when transformed into the discrete-time domain outputs

$$\begin{aligned} \vec{i}_1[k+1] = & \gamma_0 \vec{v}_R[k] + \begin{bmatrix} \beta_0 & \beta_1 & \beta_2 \end{bmatrix} \begin{bmatrix} \vec{v}_{pcc}[k] \\ \vec{v}_{pcc}[k-1] \\ \vec{v}_{pcc}[k-2] \end{bmatrix} + \begin{bmatrix} \gamma_1 & \gamma_2 \end{bmatrix} \begin{bmatrix} \vec{v}_R[k-1] \\ \vec{v}_R[k-2] \end{bmatrix} \\ & - \begin{bmatrix} \alpha_0 & \alpha_1 & \alpha_2 \end{bmatrix} \begin{bmatrix} \vec{i}_1[k] \\ \vec{i}_1[k-1] \\ \vec{i}_1[k-2] \end{bmatrix}. \end{aligned} \quad (\text{B.10})$$

where, the numeric values of the coefficients present in (B.10) are outlined in Table B.2 using the parameters defined in Table B.3.

Table B.3: System parameters [131]

Grid-side inductance	$L_1 = 150 \mu\text{H}, R_1 0.25 \Omega$
Converter-side inductance	$L_2 = 300 \mu\text{H}, R_1 0.25 \Omega$
Filter Capacitance	$C_f = 810 \text{ nF}$
Damping Resistance	$R_f = 1.1 \Omega$
Filter Resonance Frequency	$f_r = 17.68 \text{ kHz}$
Sampling time	$T_s = 20 \mu\text{s}$

B.3 Step-Ahead Control Scheme Formulation for an Active Rectifier with an *LCL* Filter

The step-ahead control scheme as presented in [99] for an L filter can be considered as a deadbeat model predictive controller (MPC), which operates at constant switching frequency due to the incorporation of a pulse-width-modulator (PWM) in its final stage as depicted in Figs. B.1 and B.3. In essence, the step-ahead controller computes the targeted rectifier voltage $\vec{v}_R[k]$ at each control step k in order to regulate the desired variables based on the prediction of the grid-side current for the next control step $\vec{i}_1[k+1]$. By itself, (B.10) has all the information to predict the grid-side current $\vec{i}_1$ at instant $k+1$ and perform a current control. However, to implement a direct power control (DPC) the active and reactive power instant $k+1$ must be set to the desired p_{ref} and q_{ref} , i.e., $p[k+1] + jq[k+1] = p_{ref} + jq_{ref}$. In addition, $p[k+1]$ and $q[k+1]$ are related to $\vec{i}_1[k+1]$ and $\vec{v}_{pcc}[k+1]$ by

$$p[k+1] + jq[k+1] = (3/2) \vec{v}_{pcc}[k+1] \left(\vec{i}_s[k+1] \right)^*. \quad (\text{B.11})$$

With this equation, the active and reactive power at the PCC can be controlled by isolating $\vec{i}_1[k+1]$ from (B.11), where the PCC voltage space-vector at instant $[k+1]$ is computed as

$$\vec{v}_{pcc}[k+1] = \vec{v}_{pcc}[k] e^{j\omega T_s}. \quad (\text{B.12})$$

Here, $\omega = 2\pi f$ is the angular frequency at the PCC and can be measured using the direct

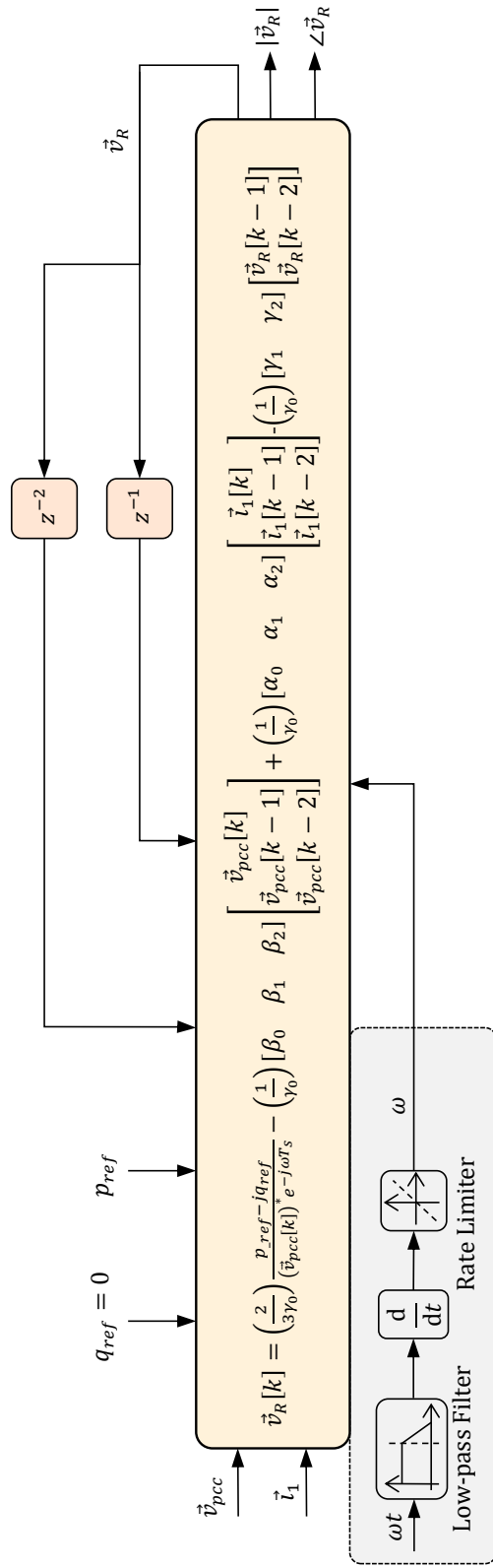


Figure B.3: Block diagram of the 2L-VSC active rectifier with *LCL* filter [131].

phase-angle detection described in Section 2.3 [127].

$$\vec{i}_s[k+1] = \left(\frac{2}{3}\right) \frac{p_{ref} - jq_{ref}}{(\vec{v}_{pcc}[k])^* e^{-j\omega T_s}}. \quad (\text{B.13})$$

Finally, by replacing (B.13) into (B.10), the targeted rectifier voltage $\vec{v}_R[k]$, which is required to regulate the active and reactive power at the PCC, can be calculated at each sampling time k as

$$\begin{aligned} \vec{v}_R[k] = & \left(\frac{2}{3\gamma_0}\right) \frac{p_{ref} - jq_{ref}}{(\vec{v}_{pcc}[k])^* e^{-j\omega T_s}} - \left(\frac{1}{\gamma_0}\right) \begin{bmatrix} \beta_0 & \beta_1 & \beta_2 \end{bmatrix} \begin{bmatrix} \vec{v}_{pcc}[k] \\ \vec{v}_{pcc}[k-1] \\ \vec{v}_{pcc}[k-2] \end{bmatrix} \\ & + \left(\frac{1}{\gamma_0}\right) \begin{bmatrix} \alpha_0 & \alpha_1 & \alpha_2 \end{bmatrix} \begin{bmatrix} \vec{i}_1[k] \\ \vec{i}_1[k-1] \\ \vec{i}_1[k-2] \end{bmatrix} - \left(\frac{1}{\gamma_0}\right) \begin{bmatrix} \gamma_1 & \gamma_2 \end{bmatrix} \begin{bmatrix} \vec{v}_R[k-1] \\ \vec{v}_R[k-2] \end{bmatrix}. \end{aligned} \quad (\text{B.14})$$

Once the targeted space-vector is computed using (B.14), $\vec{v}_R[k]$ is normalized by v_{dc} and used to form SVPWM switching pattern. Notice that to compute $\vec{v}_R[k]$ only the voltages at the PCC and the grid-side current need to be measured, given that the voltage generated by the rectifier can be estimated using the duty cycle signals generated by the SVPWM, i.e., S_{ap} , S_{bp} , and S_{cp} , in conjunction with v_{dc} as [140]

$$\begin{bmatrix} v_a^R \\ v_b^R \\ v_c^R \end{bmatrix} = \left(\frac{v_{dc}}{3\sqrt{3}}\right) \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_{ap} \\ S_{bp} \\ S_{cp} \end{bmatrix}. \quad (\text{B.15})$$

As a result, the step-ahead control scheme for active rectifiers with *LCL* grid interfacing filters can be implemented with the same amount of sensors required for an *L* filter.

B.4 Simulation Results and Discussion

A digital computer simulation using Matlab/Simulink was used to demonstrate the validity and performance of the step-ahead DPC active rectifier with an *LCL* filter. The simulated system and control scheme are depicted in Fig. B.1. Tables B.3 and B.2 contain the parameter of *LCL* filter and the coefficients used in the controller. The PWM switching frequency of the converter was set at 50 kHz. The voltage at the PCC was set to 115 V at 60 Hz, and the load is considered to be constant throughout the complete simulation. Finally, as a benchmark to compare the step-ahead DPC scheme, the developed controller is compared to a classic PI-based *PQ*-control [14] using the same 2L-VSC topology, *LCL* filter, and system parameters. Finally, to establish a fair comparison, both control schemes used the direct phase-angle detection described in Section 2.3, and the PI-based *PQ*-control scheme was tuned adequately for a load set-point of 10 kW.

B.4.1 Steady-State Performance

As a first benchmark for comparison, the steady-state performance of the developed controller is shown in Fig. B.4. Here, the PCC phase-to-neutral voltage of phase *a* (v_{an}) and the grid-side line current of the same phase (i_a) are shown with both controllers set at $p_{ref} = 10$ kW and $q_{ref} = 0$. As can be seen from Fig. B.4, both controllers keep the voltage and current in phase, demonstrating the unity power factor operation set-point given by $q_{ref} = 0$. In addition, the THD for the PI-based *PQ*-control scheme and the developed step-ahead DPC were measured at 0.87% and 1.16%, respectively. Thus, proving that both control schemes are able to generate sinusoidal currents with a low harmonic content by taking advantage of the third-order filtering capabilities of the *LCL* topology.

B.4.2 Dynamic Performance

Fig. B.5 shows the dynamic performance of the developed control scheme for a step-change of 5 kW in the active power reference p_{ref} at $t = 0.15$ s. Subsequently, p_{ref} was

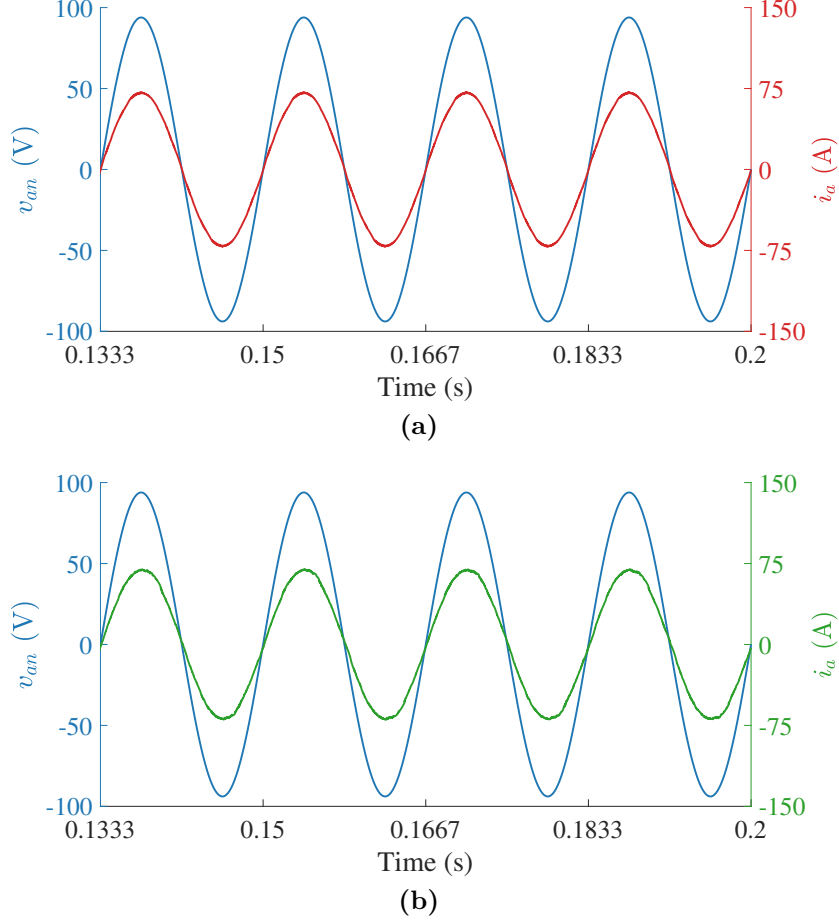


Figure B.4: Simulated steady-state performance comparison at $p_{ref} = 10 \text{ kW}$ and $q_{ref} = 0$; (a) PI-based PQ -control scheme and (b) developed DPC [131].

returned to $p_{ref} = 5 \text{ kW}$ after three cycles ($3/60 \text{ s}$). Figs. B.5(a) and Figs. B.5(b) depict the grid-side line currents, while Figs. B.5(c) and Figs. B.5(d) display the active and reactive power. As can be seen from Figs. B.5(c) and Figs. B.5(d), the developed control scheme outperforms the PI-based PQ -controller in terms of settling time and overshoot of the active power. Initially, the settling time for the 5 kW step increase appears similar for both controllers since the PI-based PQ -control scheme was tuned for load set-point of 10 kW . On the other hand, when p_{ref} was returned to $p_{ref} = 5 \text{ kW}$, the PI-based PQ -control scheme shows a less satisfactory dynamic performance. Conversely, the developed controller responds to the reference change with minimum overshoot and rapid settling time. Such an exceptionally dynamic response is a well-known characteristic of MPC techniques that can now be maintained for converters using an LCL filter.

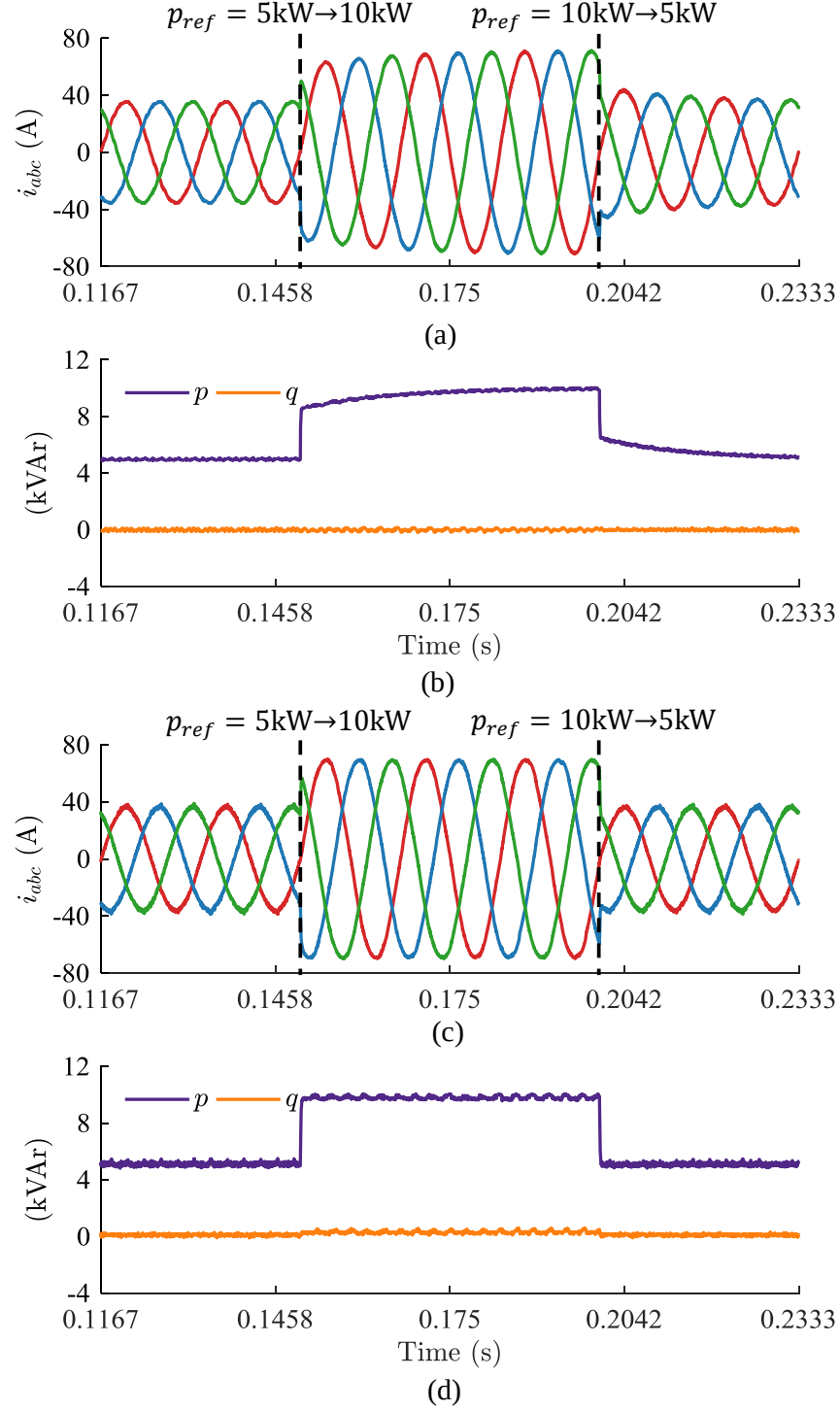


Figure B.5: Simulated dynamic performance comparison for a 5kW step-change in the active power reference p_{ref} ; (a) grid-side line currents for the PI-based PQ -control scheme, (b) grid-side line currents for the developed DPC, (c) active and reactive power for the PI-based PQ -control scheme, and (d) active and reactive power for the developed DPC [131].

B.5 Conclusion

In this chapter, a step-ahead direct power control scheme with reduced sensor count for grid-connected active rectifiers with *LCL* filter has been presented [131]. The developed formulation allows the controller to regulate the active and reactive power at the PCC by predicting the aforementioned variables one-step-ahead. Such a task has been achieved by implementing a novel grid-side current prediction expression that only requires measurements at the PCC, thus maintaining a reduced sensor count. More specifically, the step-ahead control scheme can be implemented with the same amount of sensors for an *LCL* filter than for an *L* filter [131].

The steady-state and dynamic performance of the step-ahead DPC scheme has been evaluated using Matlab/Simulink simulations and taking a PI-based *PQ*–control scheme as a benchmark controller due to its popularity and industrial implementation. As expected, both controllers perform similarly in steady-state conditions in terms of unity power factor operation and low THD for the grid-side line currents due to the implementation of an *LCL*. On the other hand, in terms of the dynamic performance, the step-ahead DPC scheme outperforms the PI-based *PQ*–controller by maintaining minimum overshoot and rapid settling time in the active power for step-changes in p_{ref} . Consequently, with the formulation provided in this chapter, the fast dynamic response and sound steady-state performance of MPC techniques, such as the strap-ahead DPC, can be preserved while taking advantage of the improved filtering capabilities of the *LCL* filter.

Appendix C

Next Generation GaN Power Converter Hardware Development

Power electronic converters take the spotlight as the main technological enabler for more electric powertrains applications. Furthermore, through the utilization of wide-bandgap (WBG) semiconductors, modern power converters can achieve unprecedented power densities and efficiencies by exploiting the inherent benefits of WBG devices. Among these devices, gallium nitride (GaN) semiconductors offer superior high-efficiency at a reduced size when compared to their silicon carbide (SiC) counterparts [53, 168, 169], as shown in Fig. 1.5. However, despite their immense potential and recent commercial availability at higher rated voltages, GaN devices bring to the table a series of design challenges that must be addressed, such as switching behavior [170], gate driver requirement [171], printable circuit board (PCB) layout [172], cross-talk [173], and thermal design [174]. Only through the mitigation of the previously mentioned challenges can the new generation of GaN-based power converters reach the desired power density ($\geq 10 \text{ kW/kg}$) and efficiency ($\geq 98\%$) required by modern more electric powertrain applications as recommended by recent government-funded programs at NASA and the U.S. Department of Energy (DOE) [175].

In this appendix chapter, the development of the second generation three-phase GaN E-HEMT power converter is developed and experimentally tested. First, the general con-

verter structure is described, followed by the first generation GaN converter analysis. This converter was used to tune critical gate-driver parameters such as the turn-off gate resistor, and the gate-source capacitor. With the experience gained from tuning and testing the first generation GaN converter, the following section describes the development of the second generation GaN power board, together with the snubber circuit tuning. Subsequently, Section C.4 demonstrates the performance of the second generation GaN power converter providing experimental results using the converter as a stand-alone inverter and as a variable-frequency active rectifier. Lastly, the challenges and future work regarding the developed GaN converter are discussed in the last section of this chapter.

C.1 General Converter Description

The design consists of two printable circuit boards (PCBs) (i) gate driver and (ii) power stage. The structure of the converter resembles an L-type shape, in which the gate driver PCB is connected at 90° with respect to the power PCB. This perpendicular connection minimizes the inductive coupling between the current loops in the two boards. The following subsections describe each PCB in detail.

C.1.1 Gate Driver Board

The gate driver PCB, as shown in Fig. C.1, receives the gate signals from an external microprocessor and creates the gate-source voltages that are physically applied to each GaN device. Typically, the gate signals are logic-type signals 1/0; where 1 represents 3.3 or 5 V and 0 is 0 V. For the designed converter, the gate signals are provided by a dSpace MircoLabBox 1202. These signals are then converted from 0 and 5 V to the specific gate-source voltage levels that the switching devices require based on their datasheet.

The designed converter implements a 650 V Enhancement Mode GaN transistor (E-HEMT) manufactured by GaN Systems. Table C.1 outlines the main parameters of the GaN E-HEMT used in the first and second generation converters. Here, the gate-to-source

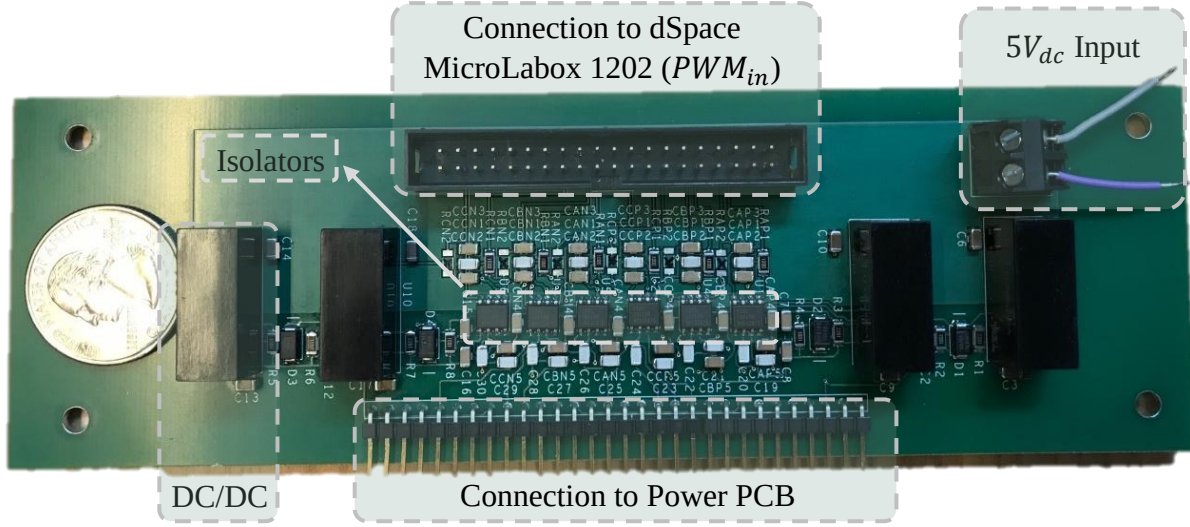


Figure C.1: GaN gate driver PCB.

threshold voltage, $V_{GS(th)}$, is typically 1.7 V, whereas V_{GS} ranges from -10 to $+7$ V. Consequently, $+6$ V and -3 V is selected as the turn-on and turn-off V_{GS} levels, respectively. A SI8271AB-ISR isolator is implemented to generate the previously mentioned voltages using the circuit topology depicted in Fig. C.2.

In terms of its power supply, the gate driver board is fed by $5 V_{dc}$. This voltage is used to power the isolators and to generate the $+6$ V/ -3 V voltages from a 5 V and 9 V dc-dc converter as shown in Fig. C.2(a). In addition, the SI8271AB-ISR isolator includes an “enable” port that is activated using a bit directly from the dSpace MicroLabBox 1202. Each isolator requires its own enable signal, which allows the converter to be used as a single-phase, dc-dc, and three-phase converter. Once enabled, each isolator outputs three

Table C.1: GaN E-HEMT parameters

Parameter	1st Generation	2nd Generation
Part Number	GS66506T	GS66508T
V_{DS}	650 V	650 V
I_{DS}	22.5 A	30 A
$R_{DS(on)}$	67 m Ω	50 m Ω
Q_G	4.5 nC	6.1 nC
V_{GS}	-10 to $+7$ V	-10 to $+7$ V
Gate-to-Source Threshold, $V_{GS(th)}$	1.1 to 2.6 V	1.1 to 2.6 V
Dimension (mm)	5.6 x 4.5 x 0.54	7.0 x 4.5 x 0.54

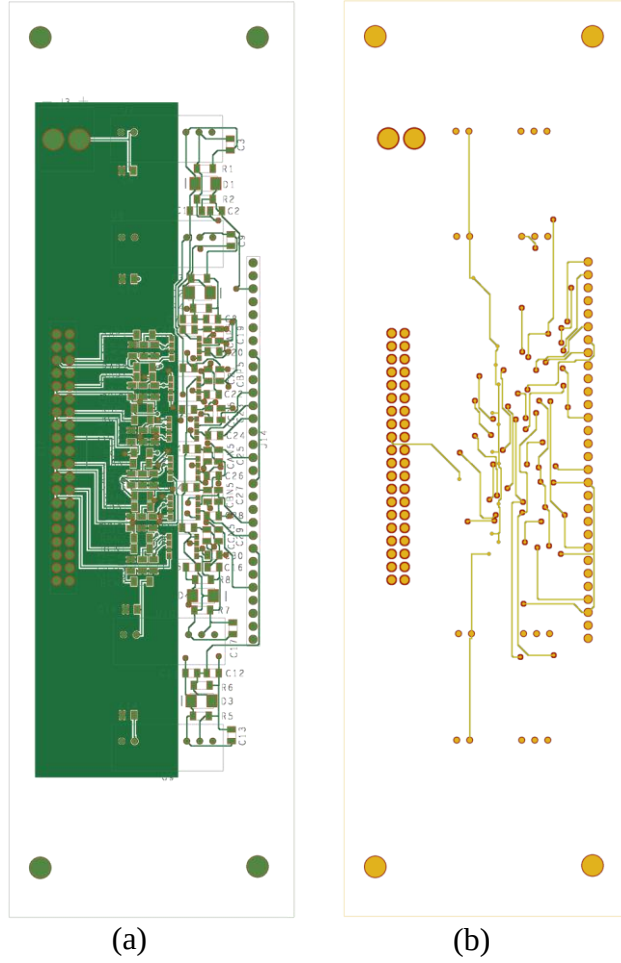


Figure C.3: GaN gate driver PCB layout art files; (a) top layer/solder-mask/silkscreen, and (b) bottom layer/solder-mask/silkscreen.

GaN gate driver PCB is specifically designed to minimize the length of the traces from the output of the isolators to the power board connector.

C.1.2 Power Board

The power PCB contains the power stage of the converter and the gate driver circuit. Fig. C.2(c) depicts the gate driver circuit applied to the six GaN E-HEMTs. The input of each gate drive circuit is the three signals coming from its corresponding isolator. The output of the gate drive circuit is directly connected between the gate and source of each transistor.

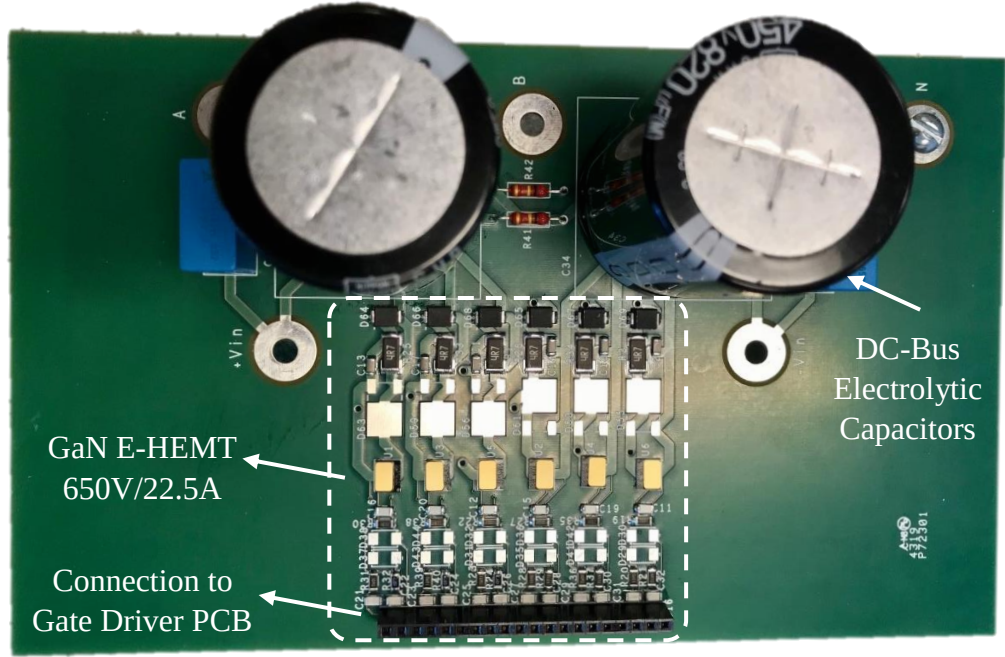
The gate driver circuit contains three resistors, one capacitor, and two Zener diodes. The turn-on $R_{g,on}$ and turn-off $R_{g,off}$ resistors can be used to adjust the rise and fall times of the switching devices. On the other hand, the Zener diodes are designed to clamp the turn-on and turn-off voltages. Lastly, the gate-source C_{gs} capacitor attenuates the high-frequency component of the signals generated by the inductive nature of the traces and the high dv/dt .

In addition to the gate driver circuits, the power PCB includes the dc-bus electrolytic capacitors and film capacitors connected in parallel with the dc-bus to reduce the switching transients due to the dc-bus loop. Additional switch protection is provided with an RCD (resistance-capacitance-diode) snubber circuit per-switch. Notice that the GaN E-HEMT offers a reverse current capability similar to the body-diodes present in conventional MOSFETs. As a result, no anti-parallel diode is required per-switch, reducing the effective footprint area of each GaN E-HEMT.

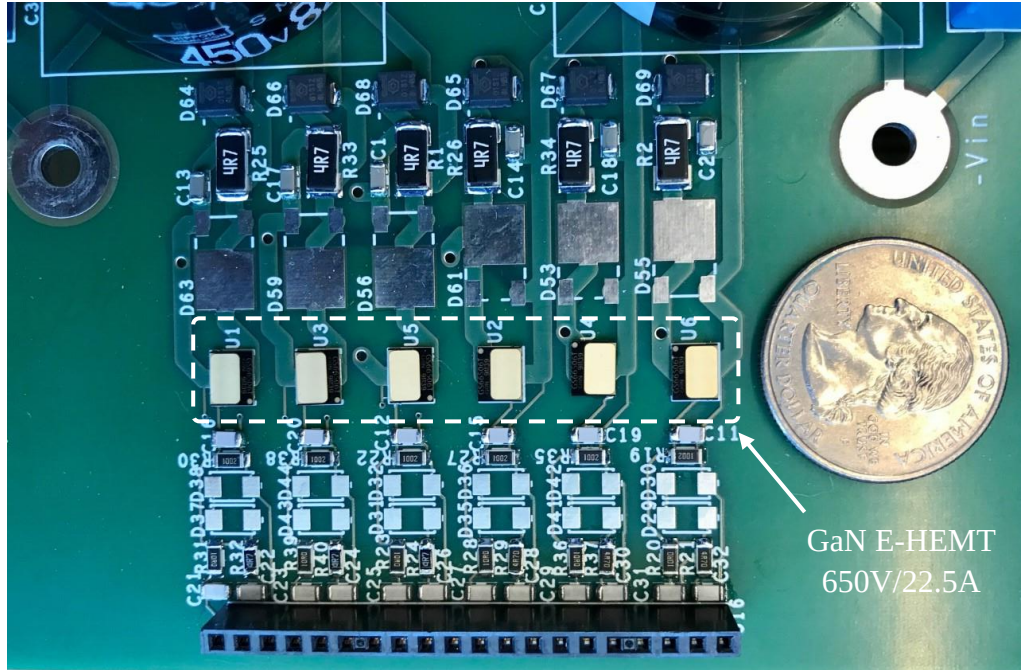
C.2 First Generation GaN Converter

The first generation GaN power converter used the gate driver PCB depicted in Fig. C.1 in conjunction with the power PCB shown in Fig. C.4. The first generation power PCB operation is limited due to its thin traces and inadequate gate driver circuit design. Nonetheless, this board was used to experimentally tune the circuit parameters of the gate driver circuit. Due to the novelty of the GaN E-HEMTs, the gate drive circuit parameters cannot be taken from standard Si- or SiC-based semiconductor devices. Consequently, the main purpose of the first generation board was to serve as an experimental board to adjust the gate resistors and C_{gs} . In addition, this first generation PCB allowed the visualization of the form factor and power density that can be achieved with GaN E-HEMTs.

Fig. C.5 demonstrates the effects of changing the gate turn-off resistance $R_{g,off}$ from 2Ω to 4.7Ω . As expected, no variations are shown in the turn-on transients, whereas the turn-off transient reduced the v_{gs} overshoot in almost 20%. That being said, the trade-off to reduce the v_{gs} overshoot is increasing the fall time by 10%. Notice that increasing the resistance does not have a significant impact on the v_{gs} oscillation frequency.



(a)



(b)

Figure C.4: First generation GaN power board; (a) full top view and (b) close-up top view.

To further reduce the v_{gs} overshoot and effectively decrease the transient oscillation frequency, the gate-source capacitance C_{gs} was augmented from $100\ \mu\text{F}$ to $560\ \mu\text{F}$. Fig. C.6 shows the effects on v_{gs} after increasing C_{gs} . It is clear from the previously mention figure

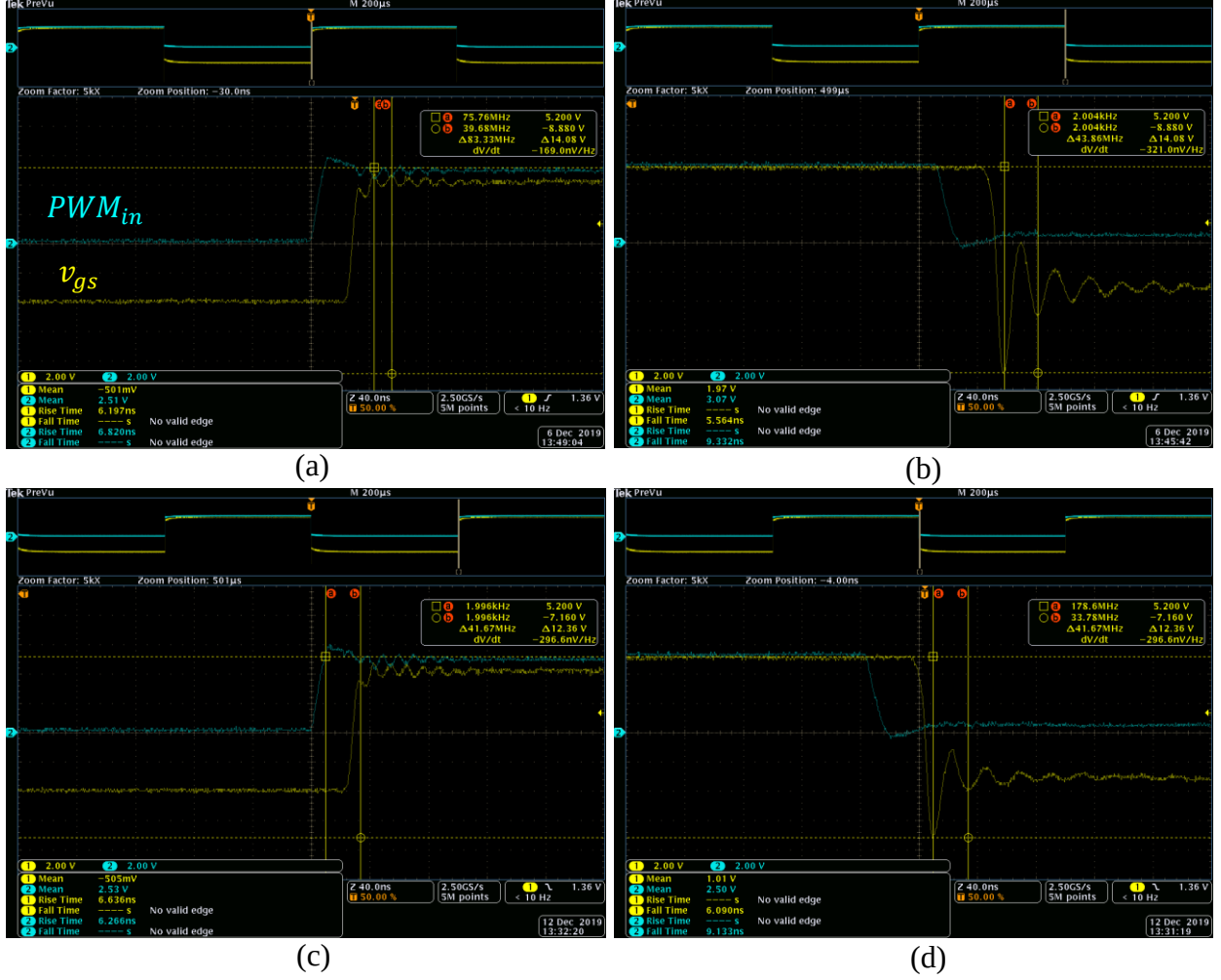


Figure C.5: Experimentally obtained effects of $R_{g,off}$ variation on the gate-source voltage v_{gs} ; (a) turn-on transient for $R_{g,off} = 2\Omega$, (b) turn-off transient for $R_{g,off} = 2\Omega$, (c) turn-on transient for $R_{g,off} = 4.7\Omega$, (d) turn-off transient for $R_{g,off} = 4.7\Omega$.

that the frequency of oscillation reduced in almost half at the expense of almost doubling the fall time of the signal. Furthermore, since C_{gs} affects both the turn-on and turn-off transients, the rise time of v_{gs} almost quadrupled. Notice that increasing C_{gs} , reduced the v_{gs} overshoot by almost 15%. Finally, based on Figs. C.5 and C.6, $R_{g,off}$ was set to 4.7Ω and C_{gs} to $560\mu F$ for the second generation GaN power board.

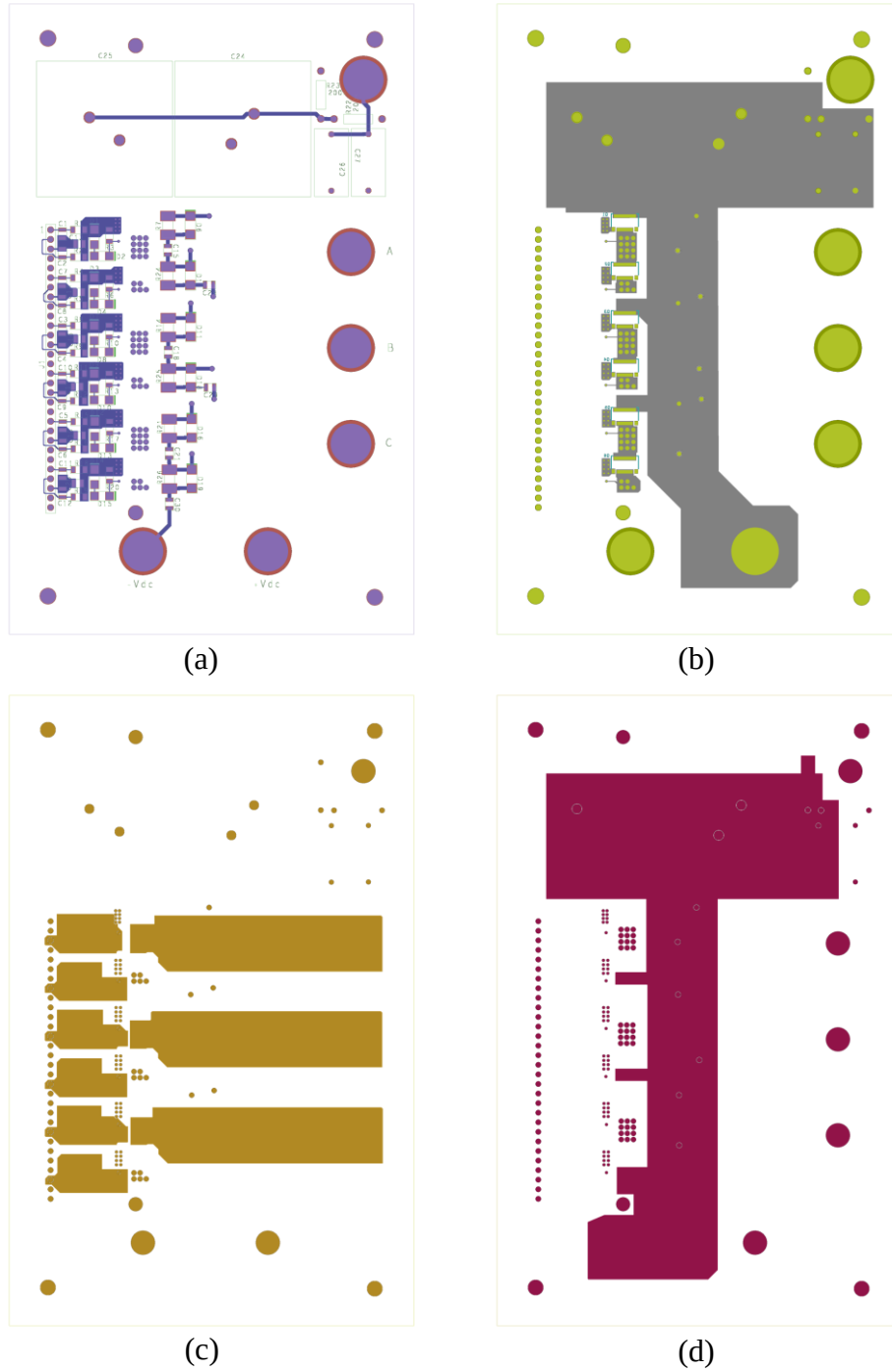


Figure C.7: GaN power PCB layout art files; (a) top layer/solder-mask/silkscreen, (b) bottom layer/solder-mask/silkscreen, (c) second layer, and (d) third layer.

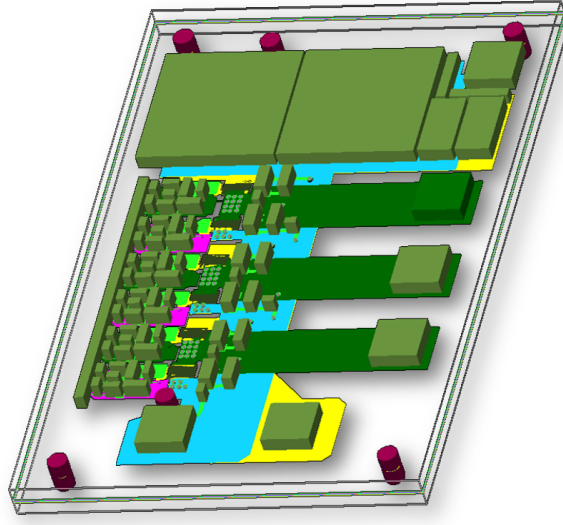


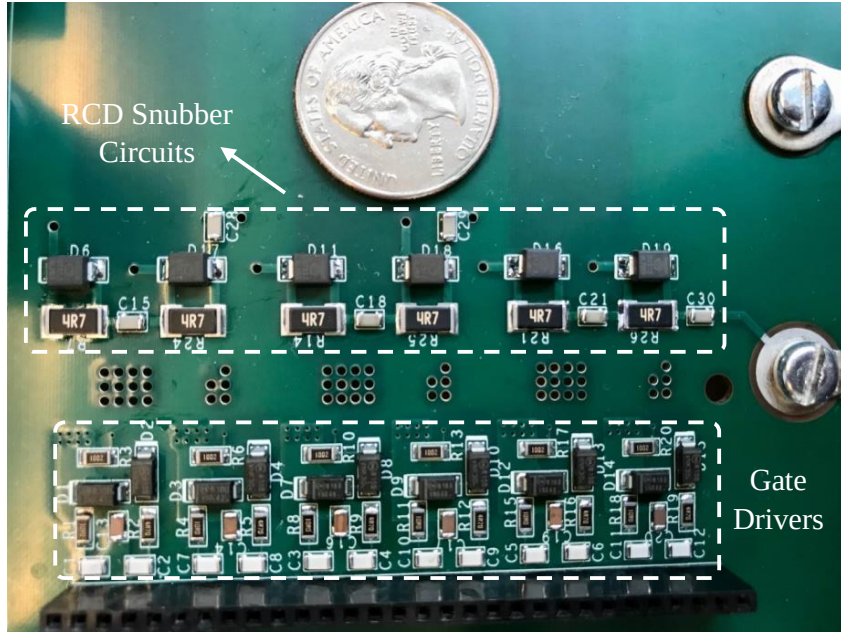
Figure C.8: Second generation GaN converter power PCB 3D view.

C.3.1 Board Layout and Routing

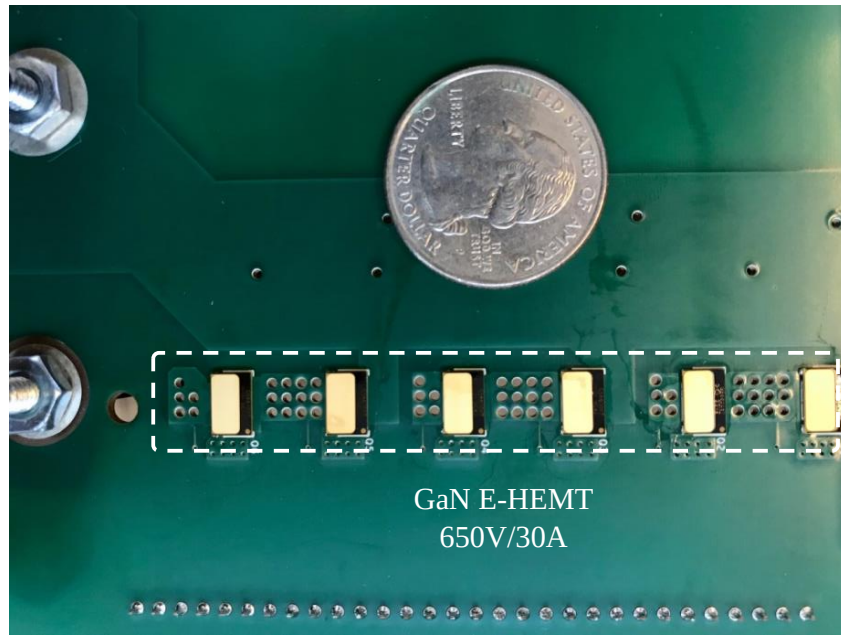
Fig. C.7 shows the PCB art files for the second generation power board. Similar to the first generation PCB, the board is designed with four layers:

1. *Top Layer:* gate driver circuits, snubber circuits, dc-bus electrolytic capacitors, dc-bus film capacitors, capacitor voltage balancing resistors.
2. *2nd Layer:* ac-traces for the three-phase output.
3. *3rd Layer:* positive dc-bus rail.
4. *Bottom Layer:* GaN E-HEMTs, and negative dc-bus rail.

A 3D view of the second generation GaN E-HEMT power PCB is depicted in Fig. C.8. This figure shows the bottom layer location of the top -cooled GaN E-HEMTs, which allows for the attachment of the heatsink to the bottom of the converter and not the top-side as in the first generation board. In addition, several via-arrays can be seen to reduce the equivalent parasitic inductance of the current loops at the desired rated current. Notice, the gate driver circuits are now compacted to reduce the parasitic inductance of the gate driver circuit.



(a)



(b)

Figure C.9: Second generation GaN power board; (a) top view and (b) bottom view.

Finally, Fig. C.9 shows the top and bottom views of the second generation power PCB. In Fig. C.9(a), the gate driver, and RCD snubber circuits can be observed; whereas Fig. C.9(b) depicts the bottom side of the converter, which houses the GaN E-HEMTs and highlights their reduced footprint size.

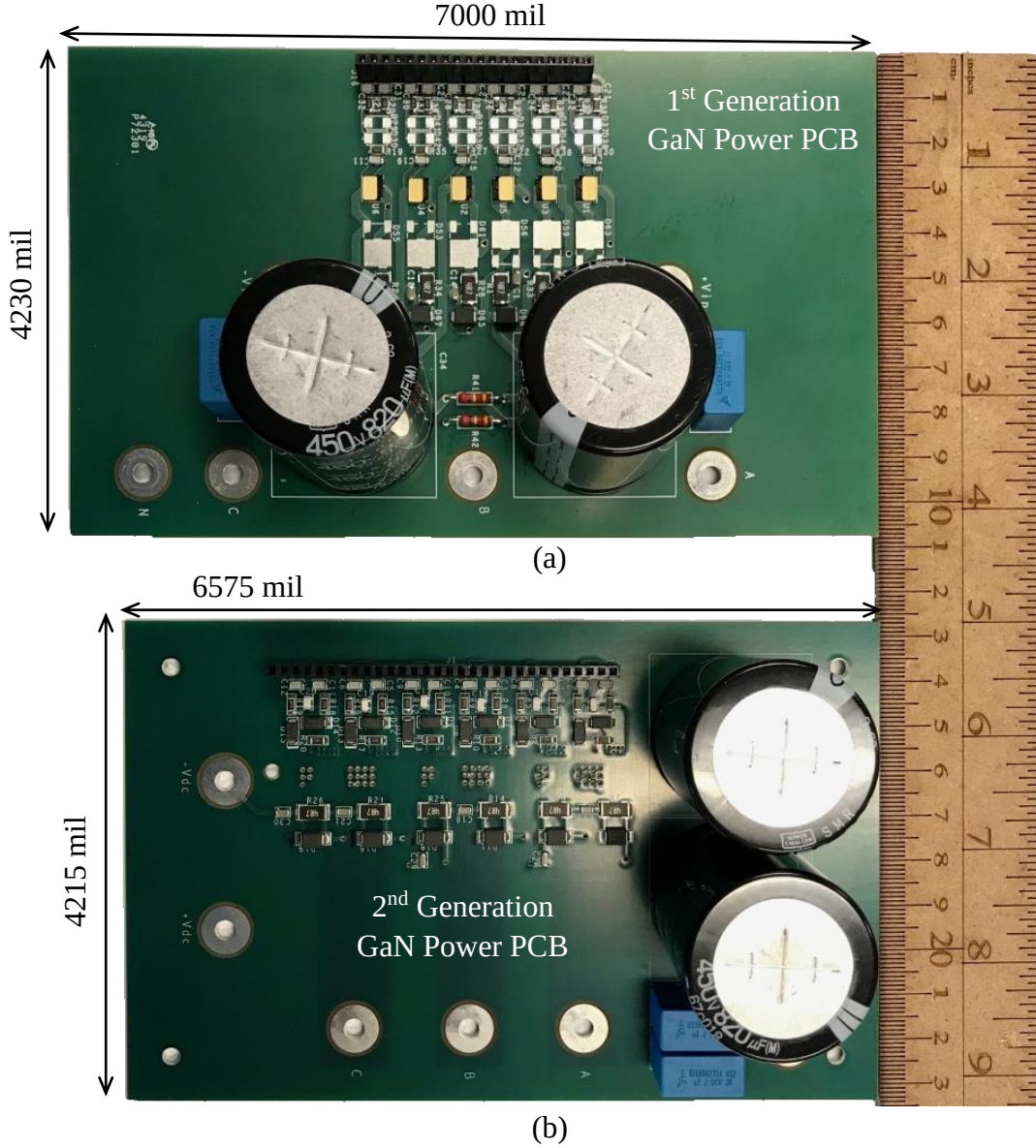


Figure C.10: First and second GaN power board comparison; (a) first generation and (b) second generation.

C.3.2 First and Second Generation GaN E-HEMT

Power Boards Comparison

Aside from lowering the parasitic inductance of the dc-bus and gate driver current loops, the second generation GaN power PCB achieved a 6% surface reduction. This size reduction has been accomplished by utilizing the inherent reverse conduction properties of the GaN E-HEMT; thus eliminating the use of an anti-parallel diode. Moreover, utilizing copper planes

to maximize the PCB area usage aided in the size reduction of the board.

Fig. C.10 demonstrates the size reduction achieved in the second generation GaN power board, while showcasing the more effective PCB area utilization accomplished in the second generation board. On the other hand, Fig. C.11 depicts the turn-on and turn-off transients for the first and second generation GaN power boards. As can be seen from this figure, the second generation power PCB exhibited a reduction in the turn-off gate voltage overshoot and oscillation when using the same circuit parameters in both boards; thus demonstrating that the parasitic inductance of the gate driver circuit current loop has been reduced in the new power PCB design.

C.3.3 Snubber Circuit Tuning

Once the gate driver circuit parameters are tuned, and the gate-source voltage signals exhibit minimum overshoots and oscillations, the snubber circuit parameters are tuned by switching only one leg of the converter using a 50% duty ratio 1 kHz signal. The snubber circuit is designed to damp the ringing in the voltage per-switch. Similar to the gate-source capacitance C_{gs} , the increase of the snubber capacitance C_s , while beneficial in terms of protecting the switch, slows down the transistor by increasing its rise and fall times. In addition to C_s , a discharging resistor with a diode connected in parallel is added to the snubber circuit, as depicted in Fig. C.2(c). This Schottky diode bypasses the resistance R_s during the charging of C_s while allowing its discharge through R_s when needed.

Fig. C.12 demonstrates the snubber capacitor tuning for the second generation GaN power board when $C_s = 470$ pF and $C_s = 680$ pF for branch B with a dc-bus voltage of $v_{dc} = 300$ V. Here, it can be observed the attenuation gained by increasing the capacitance at the expense of increasing the rise time of the switch in approximately 45%. In addition, Fig. C.12 displays the dc-bus current during the GaN E-HEMTs on/off transitions in order to visualize any possible dc-bus short-circuit. Lastly, based on the increase of the rise- and fall-times the snubber capacitance is set to $C_s = 470$ pF.

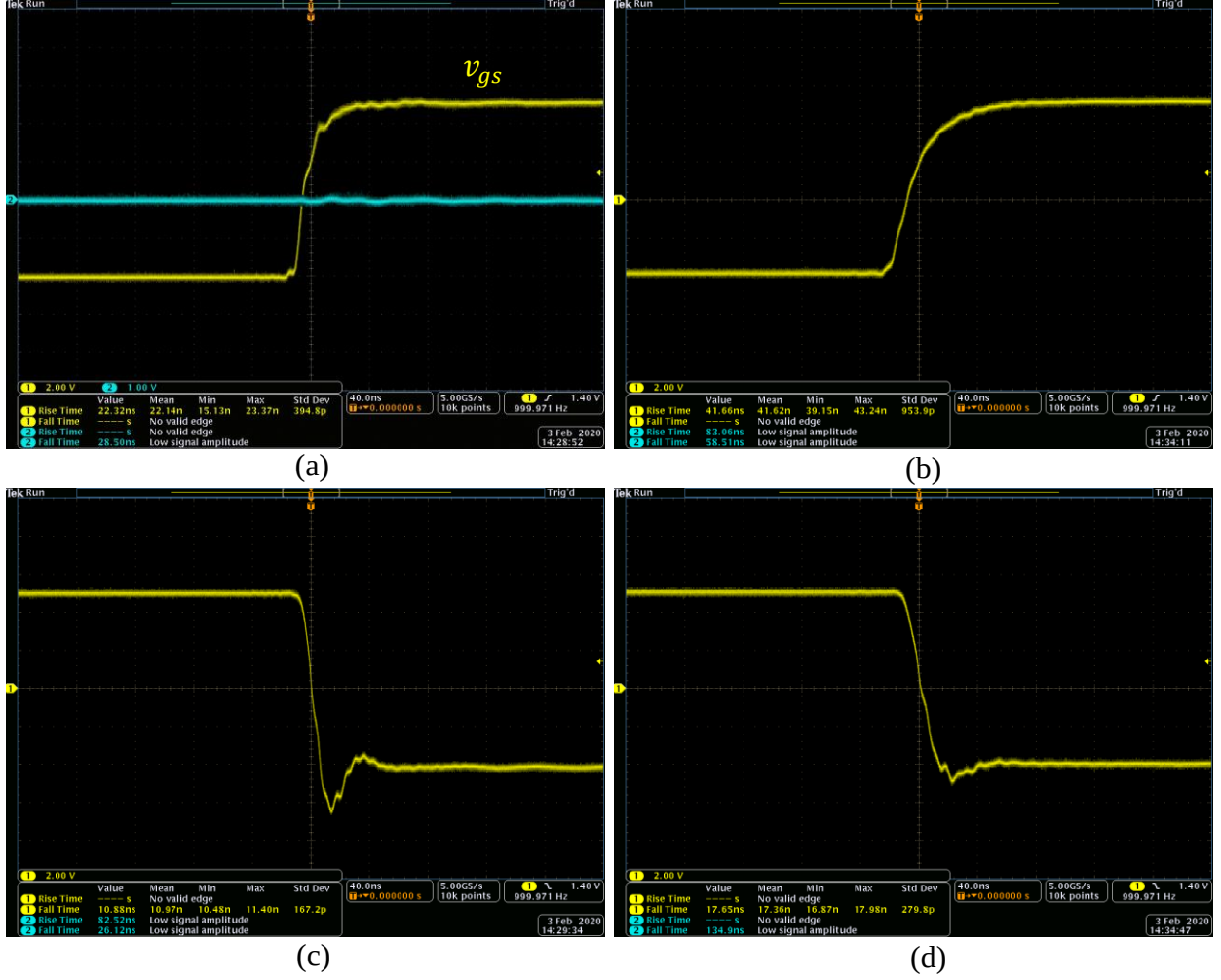


Figure C.11: Experimentally obtained gate-source voltage v_{gs} comparison between the first and second generation GaN power boards; (a) turn-on transient for the first generation board, (b) turn-off transient for the first generation board, (c) turn-on transient for the second generation board, and (d) turn-off transient for the second generation board.

C.4 Experimental Results

In this section, the second generation GaN power board, together with the gate driver PCB, described in Section C.1.1 are experimentally tested as a stand-alone three-phase inverter and a variable-frequency active rectifier. For both modes of operation, the GaN-based three-phase converter received the gate signals, and isolators enable bits from a dSPACE MicroLabBox 1202. These signals were guided to the gate driver PCB, see Fig. C.1, using a ribbon cable; which utilized a total of 12 ports (6 PWM signals + 6 Enable bits)

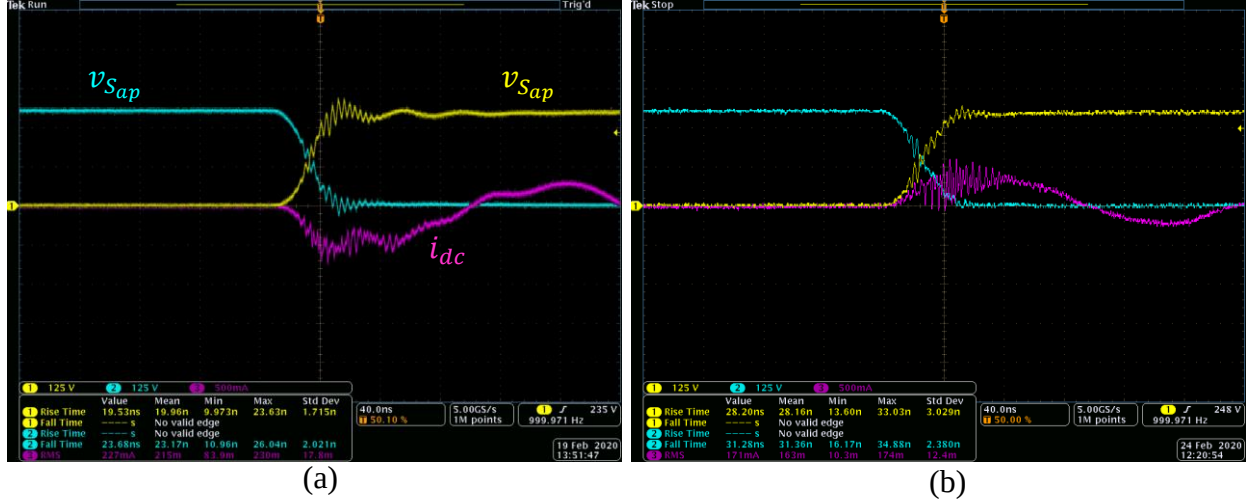


Figure C.12: Experimentally obtained snubber capacitor tuning for the second generation GaN power board for branch B ; (a) $C_s = 470$ pF, and (b) $C_s = 680$ pF.

and a common analog ground. Lastly, each operation mode, i.e., stand-alone inverter and variable-frequency active rectifier, had its particular experimental setup in terms of input, output, switching frequency, and sensors. These parameters are mentioned in the following subsections. Lastly, the ac-voltages and currents shown in this section were captured using a Teledyne LeCroy WaveRunner 64Xi-A 600 MHz oscilloscope at a rate of 500 kS/s for the variable-frequency active rectifier operation and 2.5 MS/s for the inverter operation. The collected data was re-plotted to enhance its visibility and presentation clarity.

C.4.1 Stand-Alone Inverter

For the stand-alone inverter operation, the three-phase GaN power converter is fed by a dc-power supply to provide power to constant impedance resistive-inductive (RL) three-phase load. The input to the converter was set to $v_{dc} = 100$ V, while the three-phase generated voltages were synthesized using a sinusoidal PWM (SPWM) switching scheme with a unitary modulation index, i.e., $m = 1$. Fig. C.13 shows the hardware setup for the stand-alone inverter operation with the converter running at 20 kHz supplying a 1 A RL load at a power factor PF of 0.8. The line-to-line voltage v_{ab} and line currents for the aforementioned test scenario are shown in Fig. C.14. In addition, Fig. C.14(c) depicts the

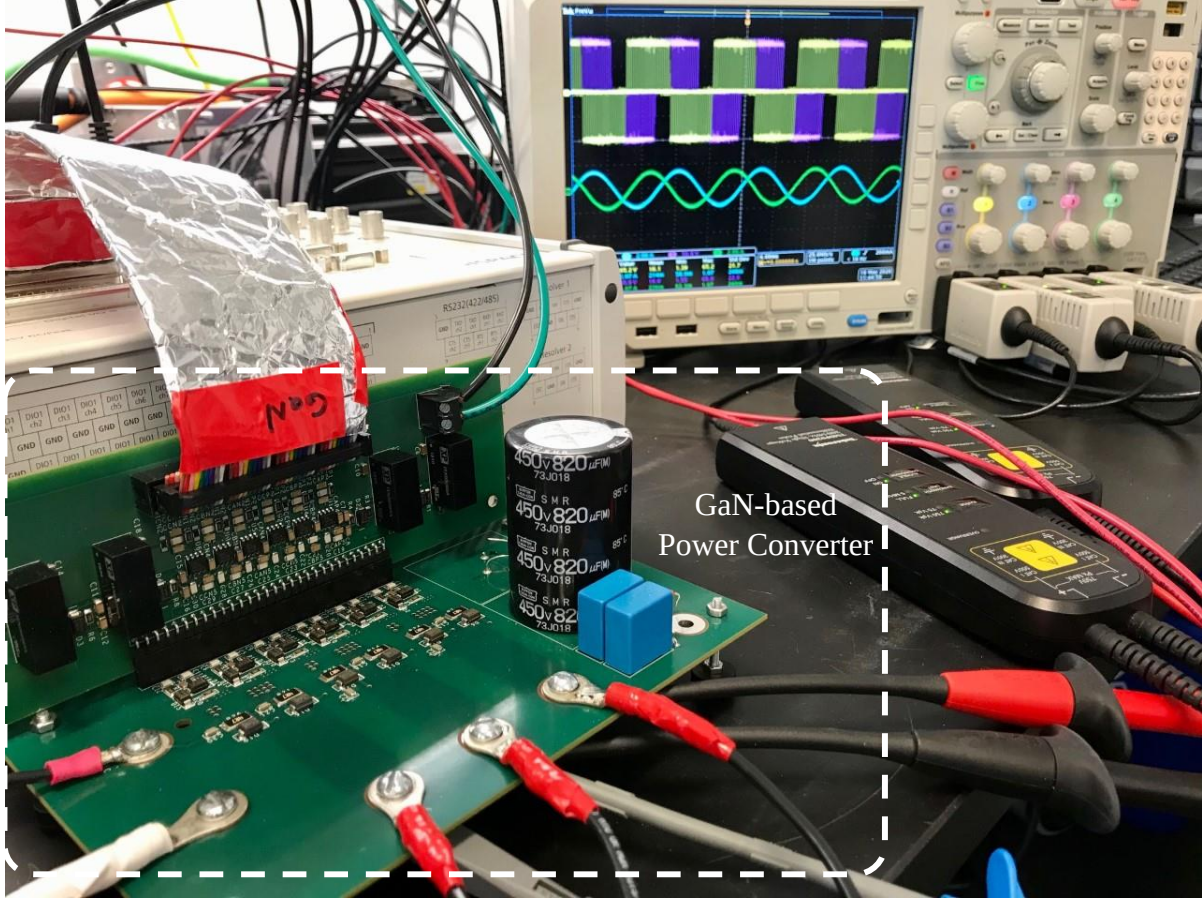


Figure C.13: Experimental laboratory setup with the second generation GaN power converter running as a stand-alone three-phase inverter supplying a constant three-phase RL load.

frequency spectrum graph of v_{ab} demonstrating the 20 kHz switching frequency operation of the inverter.

To further demonstrate the high-switching frequency capabilities of the GaN E-HEMT implemented in the design power converter. The switching-frequency of the inverter was increased to 75 kHz at $v_{dc} = 100$ V. The line currents, line-to-line voltage v_{ab} , and its frequency spectrum are shown in Fig. C.15. Here, the high switching-frequency capabilities of the designed converter are demonstrated while also supplying an RL at a lower current.

C.4.2 Variable-Frequency Active Rectifier

The second set of experimental tests were run using the second generation GaN-based power converter. Here, the converter was operated as an active rectifier with a variable-

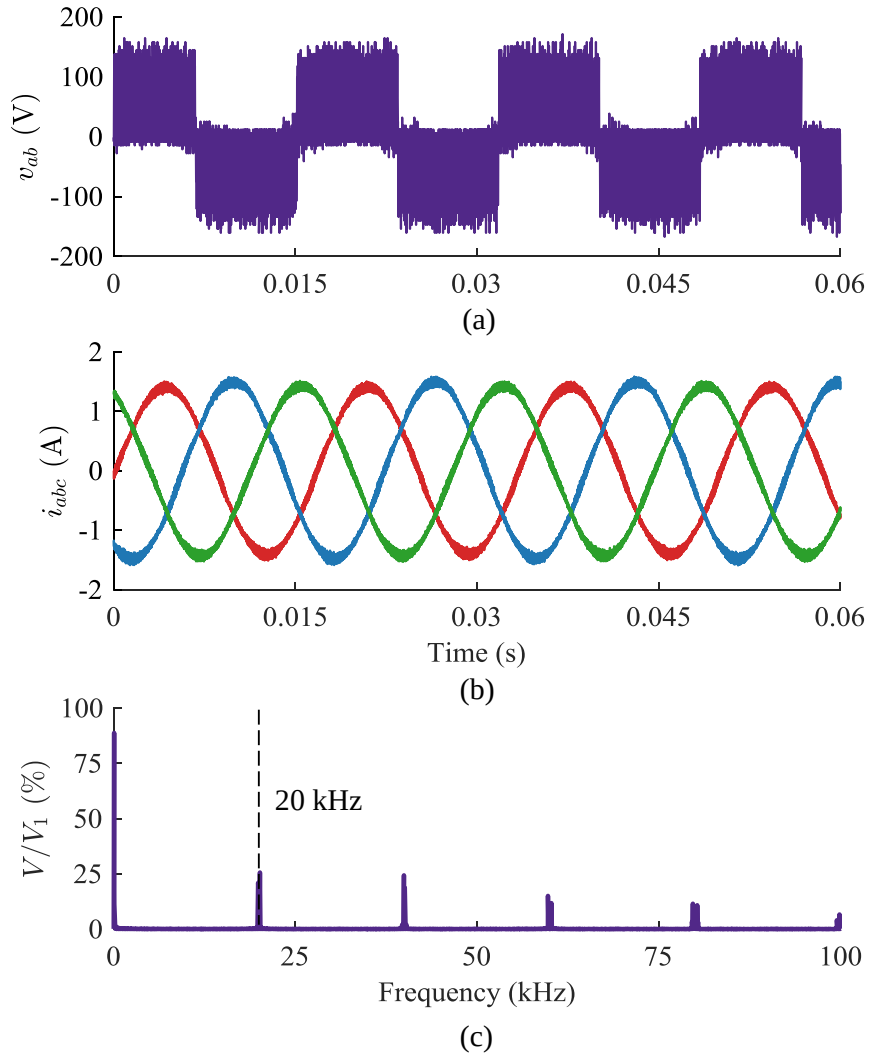


Figure C.14: Second generation GaN power converter running as a stand-alone three-phase inverter supplying a constant three-phase RL load at 20 kHz switching frequency; (a) v_{ab} , (b) line currents, and (c) frequency spectrum of v_{ab} .

frequency/constant amplitude ac-input similar to the study cases presented in Chapter 4 for the SiC-based 2L-VSC.

Among the three control schemes developed in this work and explained in Chapter 3, the step-ahead controller with no adaptive parameter estimation was selected as the controller to implement with the GaN-based power converter. The decision to implement the step-ahead control scheme among the other more advanced controllers is based on its computational burden. Between all the controllers, the step-ahead control scheme has the lowest

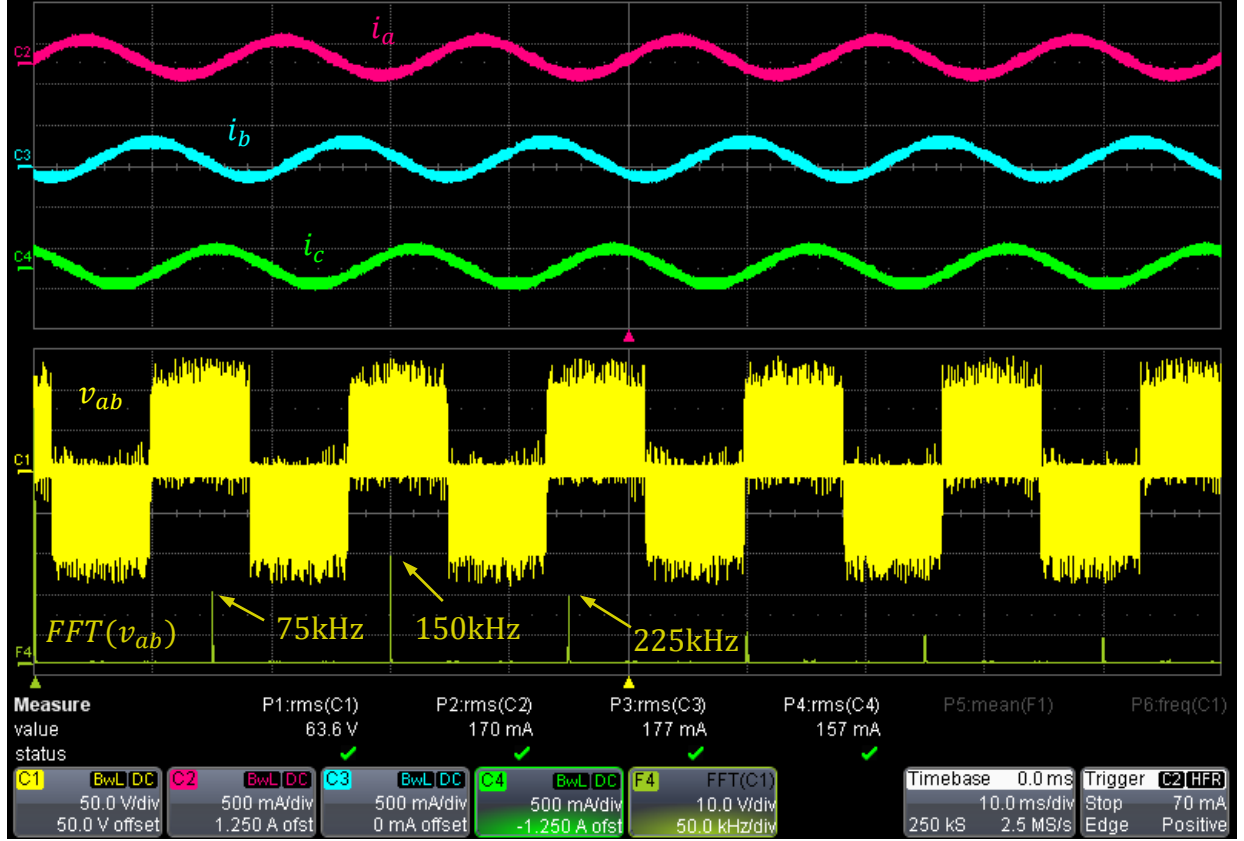


Figure C.15: Second generation GaN power converter running as a stand-alone three-phase inverter supplying a constant three-phase RL load at 75 kHz switching frequency.

turn-around time when implemented on the dSPACE MicroLabBox 1202. As a result, a higher switching-frequency can be used when implementing step-ahead control scheme, thus exploiting the benefits of the GaN E-HEMTs. Notice that this assumption is valid as long as the sampling frequency of the controller is set to the same value as the switching-frequency.

Similar to the hardware setup described in Chapter 4 for the SiC-based 2L-VSC, the GaN-based power converter was connected to a 12 kW Primate Power programmable grid emulator from Abstract Power Electronics with an input line-to-line voltage of 28 V at a variable-frequency between 100 – 400 Hz, being 400 Hz the maximum output frequency at which the hardware can operate. The converter was loaded at low-power using a resistive load with a dc voltage set-point of 100 V, and unity power factor, i.e., $Q_{ref} = 0$. In addition, a grid interfacing inductive filter with $L_f = 5$ mH and $R_f = 0.50 \Omega$ was implemented between the converter and grid emulator; whereas two 820 μ F/450 V dc-bus electrolytic capacitors

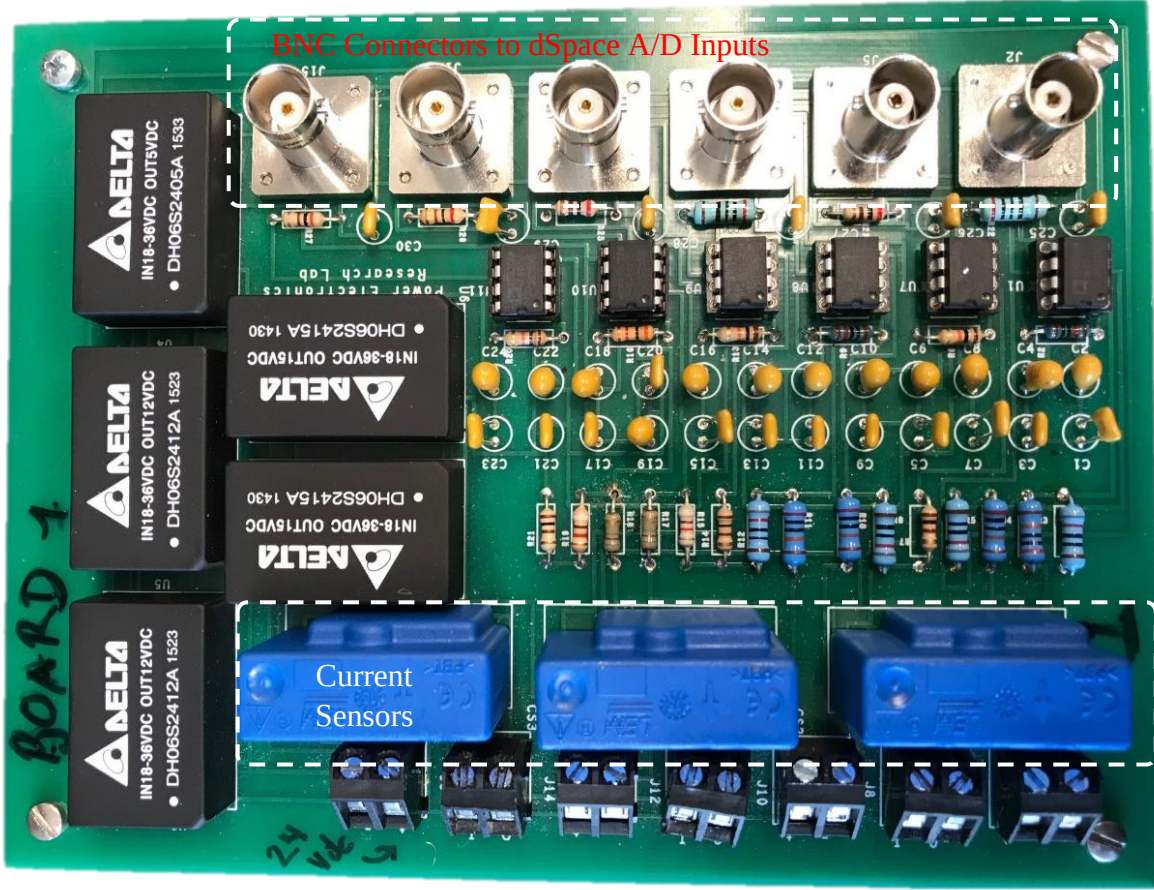
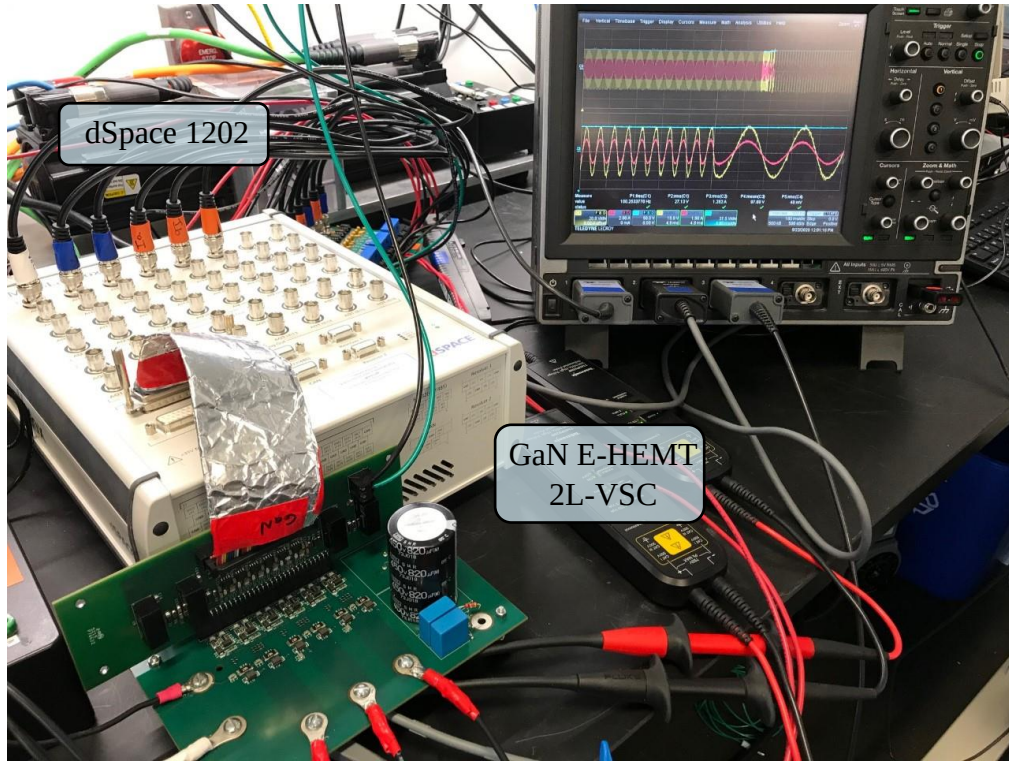


Figure C.16: Measurement board implemented in the hardware testing.

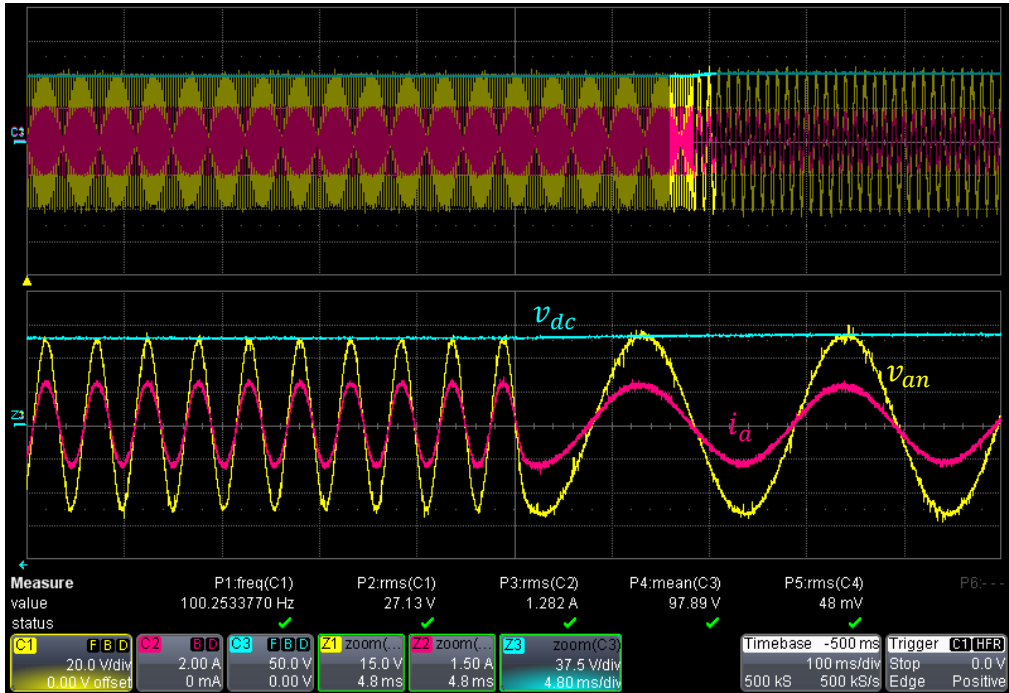
connected in series were connected to the dc-bus.

Once again, the dSPACE MicroLabBox 1202 was utilized to implement the step-ahead controller. In contrast to the SiC-based 2L-VSC, an external sensor board, shown in Fig. C.16 was used to measure v_{ab} , v_{bc} , i_a , i_b , i_c , and v_{dc} . These measurements are sent to the dSPACE MicroLabBox 1202 as part of the closed-loop controller structure. Based on the turn-around time of the step-ahead control scheme, the switching and sampling frequency were set to 40 kHz, i.e., double of the switching-frequency implement with the SiC-based converter. Aside from the switching-frequency, the remaining parameters of the step-ahead controller are the same as the ones outlined in Section 4.2.

Fig. C.17 shows the hardware setup for the GaN power converter running as a variable-frequency active rectifier. The start-up of the controller is depicted in Fig. C.18, where the transition between the distorted passive rectifier currents are changed to sinusoidal wave-



(a)



(b)

Figure C.17: Hardware setup for the variable-frequency active rectifier testing; (a) hardware setup and (b) frequency step-change oscilloscope screenshot later plotted in Fig. C.19(b).

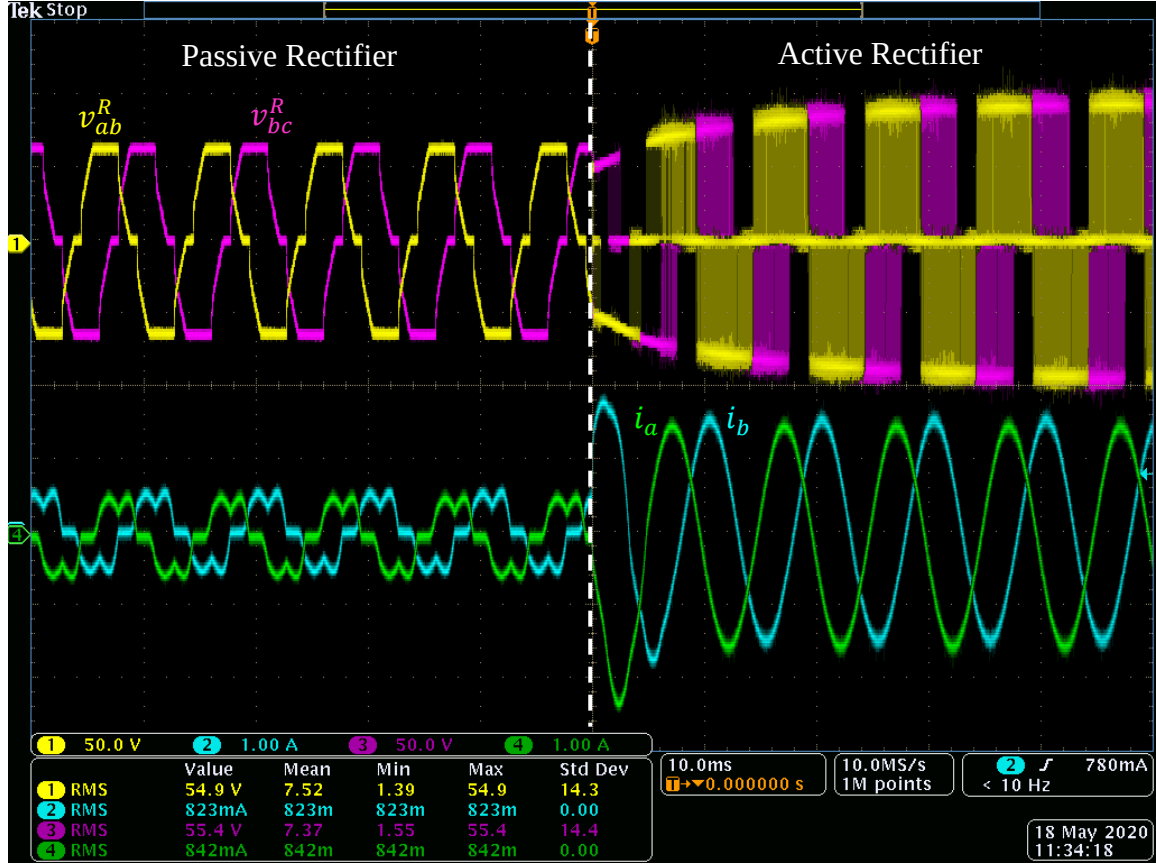


Figure C.18: Second generation GaN power converter step-ahead control scheme startup at 40 kHz switching frequency.

forms. Additionally, a 300 Hz frequency step-change was applied using the grid emulator, and the results are depicted in Fig. C.19. As expected, and based on the result obtained for the equivalent test using the SiC-based converter, with the step-ahead controller the GaN power converter is able to regulate the dc-bus voltage with minimum overshoots while maintaining a unity power factor during the two input frequency step-changes; hence demonstrating the correct hardware performance of the new generation GaN E-HEMTs power converter.

C.5 Conclusion and Future Work

Starting with the first generation of the GaN power converter, this chapter showed the development of the second generation GaN E-HEMT power converter and its effective performance at high switching-frequency running as an inverter and active rectifier. The developed

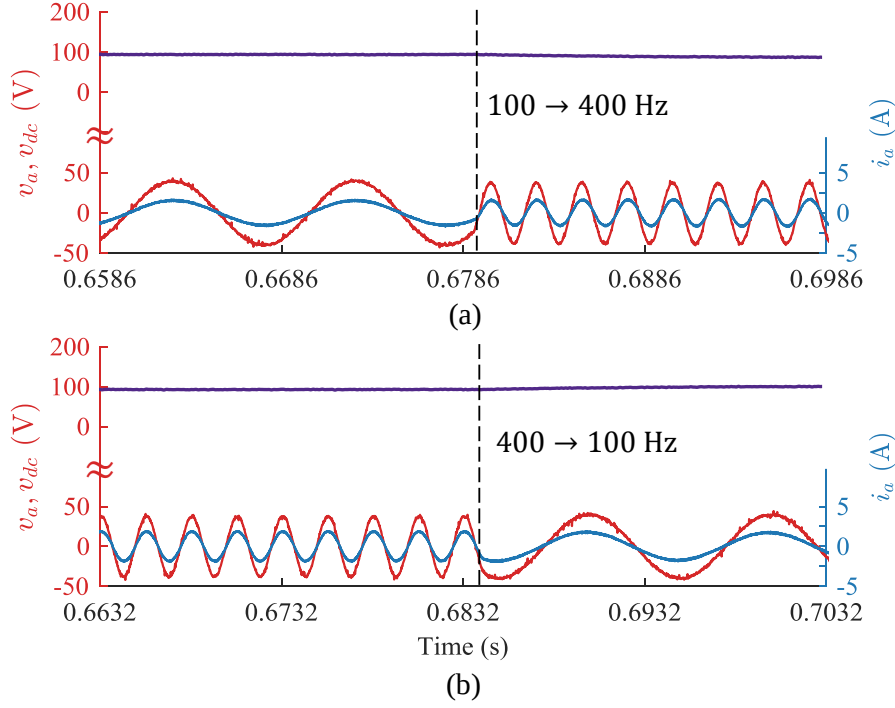


Figure C.19: Second generation GaN power converter running as a variable-frequency active rectifier equipped with the step-ahead control scheme at 40 kHz switching frequency; (a) experimentally obtained input frequency step increase from 100 to 400 Hz, and (b) experimentally obtained input frequency step decrease from 400 to 100 Hz.

design showed an improvement in the gate-source voltage signal, a higher current capability, and a 6% area reduction in the power PCB. These achievements represent some of the first steps toward the implementation of GaN-based converters in more electric powertrains. Furthermore, the experimental results presented in this chapter, while being performed at low-power/voltage provide a proof-of-concept of the operational capabilities of a GaN-based three-phase converter at high-switching frequencies.

The next steps in the development of the next generation GaN converter can now be oriented towards operating the converter at the rated current of the GaN E-HEMTs while reducing the form factor of the converter, thus dramatically increasing the power density and specific power of the new design. Additional boards can be added to the converter providing a complete modular plug-and-play structure that can include an onboard sensor board, microprocessor control unit (MCU) and digital signal processing (DSP) cards, passive

filters, and power supplies. It is expected that higher switching-frequencies (> 75 kHz) can be achieved by implementing faster processors, which will directly translate into smaller magnetic passive components that will allow the converter to achieve even higher power density and specific power.