

Technische Universität München
TUM School of Engineering and Design

Robust and Reliable Control for Three-level NPC Power Converters in Microgrids

Lei Liu

Vollständiger Abdruck der von der TUM School of Engineering and Design der
Technischen Universität München zur Erlangung einer

Doktorin der Ingenieurwissenschaften (Dr.-Ing.)

genehmigten Dissertation.

Vorsitz: Prof. Dr. Marcelo Lobo Heldwein

Prüfende der Dissertation:

1. Prof. Dr.-Ing. Dr. h.c. Ralph Kennel
2. Prof. Dr.-Ing. Zhenbin Zhang

Die Dissertation wurde am 03.07.2024 bei der Technischen Universität München
eingereicht und durch die TUM School of Engineering and Design am 16.12.2024
angenommen.

Acknowledgment

This dissertation is based on my doctoral project at the *High Power Converter Systems (HLU)* (formerly the *Electrical Drive Systems and Power Electronics (EAL)*) at the *School of Engineering and Design* of the **Technical University of Munich (TUM)**. This academic journey has been entire of challenges and growth, and many people have helped me reach this important milestone. I would like to express my sincere gratitude to all those who have given me help and support.

First and foremost, I would like to express my sincere gratitude to my supervisor, Prof. Dr.-Ing. Ralph Kennel. His guidance, support, and encouragement throughout my Ph.D. journey were invaluable. Under his meticulous guidance, I acquired profound academic knowledge, and more importantly, expanded my scholarly horizons. The spirit of free exploration he instilled in me has fostered my ability to work independently and collaborate effectively. His help and teachings to me will always be remembered in my heart. The learning experience at HLU will be valuable for my future career.

I extend my deepest gratitude to my mentor, Prof. Dr.-Ing. Zhenbin Zhang (Shandong University, China). His profound professional knowledge and rigorous academic attitude have greatly inspired me. Thanks to his meticulous mentorship, I have developed into a goal-oriented individual who welcomes challenges, continuously improves work processes, and possesses unwavering determination and perseverance. His wisdom and teachings will profoundly impact my future academic and professional life.

I am genuinely grateful to Prof. D. Eng. Yuxin Zhao (Harbin Engineering University, China), Prof. D. Eng. Ligang Wu (Harbin Institute of Technology, China), and Prof. D. Eng. Hongyi Li (Southwest University, China). Their mentorship, support, and encouragement have been vital to my growth over my Ph.D. journey. Their guidance has shaped my academic journey and taught me to approach problems with a broader perspective.

I am grateful for the valuable guidance Prof. D. Eng. Sergio Vazquez (University of Seville, Spain) provided throughout our research collaboration. He is an expert, and his insightful suggestions have been instrumental in shaping my study's direction and enhancing my work's quality. In addition, I would like to express my sincerest gratitude to Prof. Dr.-Ing. Zhen Li (Shandong University, China), for her help and care during my Ph.D. journey.

Working with my colleagues at HLU has been a pleasant experience. The group meeting exchanges between the MPC and sensorless teams led by Mr. Eyke Aufderheide and Mr.

Darshan Manoharan inspired me greatly. I would like to thank Dr. Julien Cordier, Mr. Wei Tian, and Mr. Alexander Florian for helping me familiarize myself with the institute's work. With Dr.-ing. Florian Bauer, Dr.-ing. Xiaonan Gao, D. Eng. Xicai Liu, Mr. Hannes Börngen, Mr. Stefan Klaub, Mr. Mostafa Ahmed, Dr.-ing. Haotian Xie, Dr.-ing. Qing Chen, Dr.-ing. Ying Tang, Mr. Qi Li, Mr. Qifan Yang, Mr. Dehao Kong, Mr. Yuangxiang Sun, and Mr. Yongdu Wang, the joint working experience with other colleagues, as well as the excellent academic atmosphere, teamwork, and cultural exchanges of the institute will become unforgettable memories for me. Especially, many thanks to D. Eng. Liu and Dr.-ing. Chen for their help in revising the abstract. Thanks to IT administrator Mr. Wolfgang Ebert, whose prompt assistance with IT issues has been invaluable throughout my research. Mrs. Wilma Al Begami, D. Eng. Yongxu He, and D. Eng. Dequan Yang are good friends of mine, with whom I shared many memorable moments in Munich. Many thanks to my friends D. Eng. Shuo Yang, D. Eng. Ge Song, and Mrs. Mengqi Dong for providing some help when I was preparing materials for registration for my doctoral. The financial support of the *China Scholarship Council (CSC)* is gratefully acknowledged.

Deep gratitude is extended to the More Power Electronics Energy Systems team (Shandong University, China), with particular acknowledgment given to D. Eng. Yu Li, Mr. Xiaodong Liu, Mr. Xiaozhe Liu, Mr. Mingyuan Zhang, Mr. Haoyu Chen, Mr. Guangze Chen for their invaluable assistance during the experimental phase. I am deeply appreciative of the cooperation and friendship provided by Mr. Minghao Zhang, Mr. Yafei Yin, Mr. Qianli Xing, Mr. Haitao Li, Mr. Jiateng Gu, Ms. Xinqing Zhang, Mr. Jin Zhang, Mr. Shichang Zhou, Mr. Yilin Wang, and Mr. Yimin Zhang to all friends for the care and companionship.

Last but also the most sincere gratitude goes to my beloved family: my dear parents and brother, who have always been selfless in love and enduring support. I am grateful for my in-laws and extended family's additional warmth and support. To my husband, Prof. D. Eng. Yunfei Yin (from boyfriend to husband), for his love, companionship, and patience over the past 15 years.

I could only finish this work with their support. Once again, I would like to express my deep gratitude.

Munich, in December 2023
Lei Liu

Abstract

This dissertation aims to design robust and reliable control schemes for three-level neutral-point-clamped power converter systems in microgrids. A quasi-sliding mode control is proposed for the system, enhancing power/voltage regulation performance. To customize designs and optimizations of converter control systems, an adaptive sliding mode control is proposed; moreover, it improves the capacitor voltage balancing. Furthermore, a super-twisting algorithm current-voltage control is designed to address the poor sinusoidal current tracking. To alleviate model-based control issues, a data-driven model-free adaptive current-voltage control is presented.

Kurzfassung

Die vorliegende Dissertation verfolgt das Ziel, robuste und zuverlässige Regelungsverfahren für Dreipunkt-Wechselrichtersysteme in Microgrids zu entwerfen. Zu diesem Zweck wird ein Quasi-Sliding-Mode-Regler vorgeschlagen, der das Leistungsvermögen der Spannungs- und Leistungsregelung erhöht. Zur Optimierung des Systems, wird zudem eine adaptive Sliding-Mode-Regelungsstruktur untersucht, die zugleich den Ausgleich der Spannungen an den Zwischenkreiskondensatoren verbessert. Des Weiteren wird unter Verwendung des Super-Twisting-Algorithmus eine Strom-Spannungs-Regelung entwickelt, die ein besseres Folgeverhalten bei sinusförmigen Referenzströmen gewährleisten soll. Zur Lösung modellbasierter Regelungsprobleme wird ein datenbasiertes, modellunabhängiges adaptives Strom-Spannungs-Regelverfahren vorgestellt.

Contents

1	Introduction	1
1.1	Background	1
1.2	Research Motivations	3
1.2.1	3L-NPC Power Converter Topology Benefits	4
1.2.2	Key Issues of 3L-NPC Power Converter Control	4
1.3	State-of-the-Art for 3L-NPC Power Converter Control	5
1.3.1	3L-NPC Power Converter Control Objectives	5
1.3.2	Typical Control Structure of 3L-NPC Power Converter	6
1.3.3	Linear Controller	7
1.3.3.1	Proportional Integral Controller	8
1.3.3.2	Proportional Resonance Controller	8
1.3.3.3	Linear Control in Power Converter Application	8
1.3.4	Nonlinear Control	9
1.3.4.1	Sliding Mode Control	10
1.3.4.2	Sliding Mode Control in Power Converter Application	13
1.3.4.3	Adaptive Control	14
1.3.4.4	Adaptive Control in Power Converter Application	16
1.3.4.5	Other Nonlinear Control in Power Converter Application	18
1.3.5	Observer-based Control	18
1.3.5.1	State Observer	19
1.3.5.2	Disturbance Observer	22
1.3.5.3	Observer-based Control in Power Converter Application	22
1.4	Outline	24
2	3L-NPC Power Converter Principle Analysis and Modeling	27
2.1	3L-NPC Power Converter Circuit Analysis	27
2.2	Switching State Analysis	28
2.3	Commutation Process Analysis	29
2.4	3L-NPC Power Converter Modeling	32
2.4.1	Stationary Coordinate Model (<i>abc</i> -axis)	32
2.4.1.1	Inductance Current Dynamics	33
2.4.1.2	Capacitance Voltage Dynamics	34

2.4.2	Stationary Coordinate Model ($\alpha\beta\gamma$ -axis)	36
2.4.3	Rotating Coordinate Model ($dq0$ -axis)	38
2.5	PWM Technology of 3L-NPC Power Converter	38
2.6	Summary	39
3	Quasi-sliding Mode Power-Voltage Control for 3L-NPC Power Converters	41
3.1	Introduction	41
3.2	System Description and Modeling	43
3.3	Control Strategy Design	46
3.3.1	Control Objectives	46
3.3.2	Power Tracking Loop	46
3.3.2.1	Discrete-time State Observer(DSTO) Design	47
3.3.2.2	Quasi-integral Sliding Mode (QISM) Surface Design	50
3.3.2.3	Quasi-integral Sliding Mode Control (QISM) Law Design	50
3.3.2.4	QISM Stability Analysis	51
3.3.3	Voltage Regulation Loop	51
3.3.3.1	DSTO Design	52
3.3.3.2	QISM Surface Design	52
3.3.3.3	QISM Law Design	52
3.3.4	Voltage Balancing Loop	52
3.3.4.1	Discrete-time Proportional-Integral (DPI) Design	53
3.4	Discrete-time Super-Twisting Observer-based Quasi-Integral Sliding Mode Control (DSTO-QISM) Tuning for the Power Tracking Loop	53
3.4.1	DSTO Parameters Tuning Process	53
3.4.2	QISM Surface Parameters Tuning Process	53
3.4.3	QISM Parameter Tuning Process	54
3.5	DSTO-QISM Tuning for the Voltage Regulation Loop	55
3.6	Robustness and Chattering Suppression for the Power Tracking Loop	55
3.7	Experimental Results	56
3.7.1	An Overall Evaluation Comparison	56
3.7.2	Comparative Voltage Regulation Performance	59
3.7.3	Comparative Power Tracking Performance	61
3.7.4	Evaluating Control Performance via Trajectory Motion	61
3.7.4.1	Voltage/Power Smooth Trajectory on the QISM Surface	63
3.7.4.2	Voltage/Power Travel Distance on the QISM Surface	64
3.7.5	Robustness Evaluation with External Disturbance	65
3.7.6	Performance Evaluation under Parameter Mismatch	65
3.7.6.1	Transient-state Performance with Parameter Mismatch	65
3.7.6.2	Steady-state Performance with Parameter Mismatch	65
3.8	Summary	67
4	Adaptive Sliding Mode Power-Voltage Control for 3L-NPC Power Converters	69
4.1	Introduction	69
4.2	System Description	71
4.3	Control Strategy Design	74

4.3.1	Control Objectives	74
4.3.2	Voltage Regulation Loop	75
4.3.2.1	Adaptive observer (AO) Design	75
4.3.2.2	AO-based Proportional (AO-P) Voltage Controller Design . .	77
4.3.3	Power Tracking Loop	77
4.3.3.1	Disturbance Observer (DO) Design	78
4.3.3.2	DO-based Integral Sliding Mode Control (DO-ISM) Power Controller Design	80
4.3.4	Voltage Balancing Loop	81
4.3.4.1	Sliding Mode Observer (SMO) Design	82
4.3.4.2	SMO-based Proportional (SMO-P) Voltage Balancing Con- troller Design	84
4.4	AO-P Control Tuning for Voltage Regulation	85
4.5	DO-ISM Control Tuning for Voltage Regulation	85
4.6	SMO-P Control Tuning for Balancing Voltage	86
4.7	Experimental Results	86
4.7.1	An Overall Evaluation Comparison	88
4.7.2	Comparison of Voltage Regulation Performance	89
4.7.3	Comparison of Power Tracking Performance	90
4.7.4	Comparison of Voltage Balancing Performance	91
4.7.5	Robustness Evaluation	94
4.7.5.1	External Disturbance DC-load Variations	94
4.7.5.2	Performance Comparison of Inductance Mismatches	95
4.8	Summary	95
5	Super-Twisting Current-Voltage Control for 3L-NPC Power Converters	97
5.1	Introduction	97
5.2	System Description	99
5.2.1	Converter Description and Modeling	99
5.3	Control Strategy Design	101
5.3.1	Control Objectives	102
5.3.2	Current Tracking Loop	102
5.3.2.1	Luenberger Observer (LO) Design	103
5.3.2.2	LO-based Super-Twisting Algorithm (LO-STA) Control Design	103
5.3.3	Voltage Regulation Loop	106
5.3.3.1	Extended Sliding Mode Disturbance Observer (ESMDO) Design	107
5.3.3.2	ESMDO-based Proportional (ESMDO-P) Control Design . .	108
5.3.4	Voltage Balancing Loop	108
5.3.4.1	Proportional-integral (PI) Control Design	108
5.4	Experimental Results	109
5.4.1	An Overall Evaluation Comparison	109
5.4.2	Performance Comparisons of Voltage Regulation	110
5.4.3	Performance Comparison of Current Tracking	110
5.4.4	Performance Evaluation with Parametric Robustness	114
5.4.4.1	Capacitance Parametric Robustness Comparison	114

5.4.4.2	Performance Comparison of Inductance Deviation	114
5.4.5	Performance Comparison of Grid Frequency Variation	117
5.5	Summary	117
6	Data-driven Model-Free Adaptive Control for 3L-NPC Power Converters	119
6.1	Introduction	119
6.2	Converter Description and Modeling	121
6.3	Control Scheme Design	123
6.3.1	Control Objectives	123
6.3.2	Current Tracking Loop	124
6.3.2.1	Model-free Adaptive Control (MFAC) Design	124
6.3.3	Voltage Regulation Loop	127
6.3.3.1	MFAC Design	129
6.3.4	Voltage Balancing Loop	129
6.3.4.1	PI Design	129
6.3.5	Brief Review on Particle swarm optimization (PSO)-based-MFAC Opti- mization Process	129
6.4	Experimental Results	132
6.4.1	An Overall Evaluation Comparison	133
6.4.2	Comparative Voltage Regulation Performance	133
6.4.3	Comparative Current Tracking Performance	136
6.4.4	Robustness Evaluation with External Disturbance	137
6.4.5	Robustness Evaluation under Parameter Mismatches	137
6.4.6	Versatile Multi-scenario Performance Assessment	139
6.5	Summary	142
7	Conclusions and Outlooks	143
7.1	Conclusions	143
7.2	Outlooks	145
A	List of Publications	147
A.1	Books	147
A.2	Journal Papers	147
A.3	Conference Papers	148
B	Nomenclature	149
B.1	Abbreviations	150
C	Testbench	153
C.1	Flow Diagram of the Experiment Infrastructure	153
C.2	Grid Simulator (Chroma 61830)	153
C.2.1	Feature	154
C.2.2	Specification	154
C.3	Current Sensors ($\pm 50\text{A}$ Din Rail-mountable Current Sensors)	155
C.3.1	Key Features and Specifications of Current Sensors	155
C.4	Voltage Sensors ($\pm 800\text{V}$ Din Rail-mountable Voltage Sensors)	155

C.4.1	Key Features and Specifications of Voltage Sensors	156
C.5	Imperix Passive Filters Box	156
C.5.1	Internal Components	156
C.5.2	Main Electrical Schematic	157
C.5.3	Main Electrical Specifications	157
C.6	3L-NPC Power Converter PEN8018	157
C.6.1	Typical Applications	157
C.6.2	Key Features and Specifications & Main Components	157
C.7	Control Platform B-Box RCP ^{3.0}	157
C.7.1	Hardware Components	159
C.7.2	Software Platform	159
C.7.3	Data Storage and Management	159
C.7.4	Power Supply	159
C.7.5	Mechanical Structure and Casing	159
C.7.6	Specification	159
C.8	Hardware-in-the-Loop Platform-RT BOX	160
C.8.1	Processor	160
C.8.2	Simulation Time Steps	162
C.8.3	Scalability	162
C.8.4	Technical Specifications	162
C.9	In summary	162
D	Engineering Guidelines	163
D.1	Engineering Issues in 3L-NPC Converter Systems	163
D.2	Multi-loop Control Architecture	163
D.3	Control Method Classification	163
D.4	Experimental Validation Objectives	164
D.5	Power-Voltage Control – Addressing Key Engineering Issues	165
D.5.1	Solutions	165
D.5.2	Core Solution Idea of Quasi-sliding Mode Control	165
D.5.3	Core Solution Idea of Adaptive Sliding Mode Control (ASMC)	166
D.6	Current and Voltage Control-Addressing Key Engineering Issues	167
D.6.1	Core Solution Idea of Super-twisting Control	167
D.6.2	Core Solution Idea of Data-driven MFAC	169
D.7	Problems Addressed and Application Scenarios of Control Schemes	170
D.8	Experimental Devices and Circuit Assembly	170
D.9	Experimental Procedure	171
D.10	Real-Time Monitoring and Adjustment	172
D.11	Data Collection and Analysis	172
D.12	Experiment Conclusion	172
D.13	Safety Measures	172
D.14	Summary	172

E	Discrete-time Extended State Observer-based Quasi-sliding Mode Control Design	173
E.1	Discrete-time Extended State Observer-based Quasi-sliding Mode Control (DESO-QSMC) Design for Power Tracking	173
E.1.1	Discrete-Time Extended State Observer (DESO) Design	173
E.1.2	Quasi-integral Sliding Mode Control (QSMC) Design	174
E.2	DESO-QSMC Design for Voltage Regulation	175
E.2.1	DESO Design	175
E.2.2	QSMC Design	175
	List of Figures	177
	List of Tables	181
	Bibliography	183

CHAPTER 1

Introduction

This thesis focuses on robust and reliable control for three-level neutral-point-clamped (3L-NPC) power converters in microgrids. This chapter outlines the background of the research, identifying the research motivations and the control objectives. Then, it discusses the state-of-the-art control techniques for 3L-NPC in microgrids. The chapter concludes with an overview of the thesis organization.

1.1 Background

Microgrids represent the future of power grids as they integrate distributed energy resources like solar panels, wind turbines, and energy storage devices, defining a new power system characterized by intelligent flexibility, high efficiency, and environmental protection [1]. Microgrids are local grids comprising distributed generation, distribution lines, and consumer loads that operate independently or link to the main utility grid. They are small, smart electric grids with features that make them particularly effective at managing energy and ensuring reliable delivery. Microgrids have benefits: 1) Emission reduction: by utilizing renewable energy sources, carbon emissions are diminished; 2) Efficiency: it directly employs energy, thereby enhancing efficiency; 3) Grid compatibility: it augments the capacity to integrate with emerging energy sources; 4) Reliability: it operates autonomously during emergencies, guaranteeing a stable power supply; 5) Specialized solutions: it delivers power to remote or hard-to-access areas. The microgrid market is estimated to be USD 15.92 billion in 2024. It is projected to grow to USD 38.12 billion by 2029, with a compound annual growth rate (CAGR) of 19.08% during the forecast period (2024-2029), as shown in Figure 1.1. The European microgrid market is anticipated to record a revenue of 11026.01 million by 2030, with an estimated CAGR of 10.65% during the forecast period from 2022 to 2030. Germany is among the key consumers of microgrids in Europe.

Microgrid control optimizes energy configuration and ensures power supply stability and efficiency within microgrids. The microgrid control market is estimated to reach USD 4.27

billion in 2024. It is expected to grow to USD 7.79 billion by 2029, with a CAGR of 12.77% during the forecast period (2024-2029) ¹ as depicted in Figure 1.1. Therefore, microgrid control will exhibit an essential impact on future power grids [1–3].

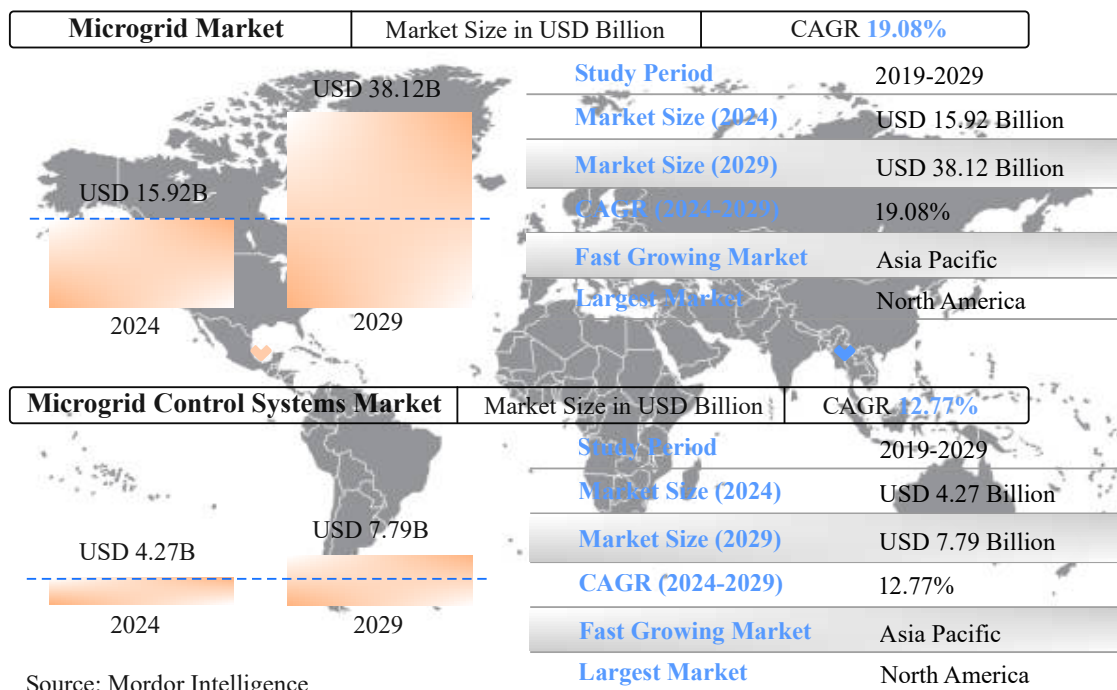


Figure 1.1: The market size of microgrids and microgrid control systems.

Microgrid control technologies offer innovative tools and frameworks for modeling the dynamic behavior of microgrids, regardless of their types and structures [4–6]. Various control technologies are implemented to enhance microgrids’ efficiency, reliability, and sustainability. Microgrid control systems, such as hierarchical structures, operate autonomously to control energy dispatch and enhance power system reliability [7, 8]. Fig. 1.2 illustrates the hierarchical structure of an ac microgrid control system, which operates across three control levels: primary, secondary, and tertiary [9]. The primary control level regulates voltage, current, frequency, power, and virtual impedance, manages energy storage, and ensures power quality within seconds [10, 11]. Secondary control corrects voltage and frequency deviations, optimal power flow, and black start management with minutes [12, 13]. Tertiary control is tasked with market integration, inter-microgrid coordination, and optimal power dispatch [14, 15], ranging from minutes to hours. Among these, the primary control is a crucial control level, requiring rapid responses to changes in grid status to maintain the microgrid’s stability and reliability.

Energy conversion equipment serves as the microgrid’s core for connecting diverse distributed power sources with varying characteristics in a user-friendly and flexible manner as the physical embodiment of primary control for microgrids.

¹The report details the growth of the microgrid control systems market, categorizing it by type (grid-connected, off-grid, and hybrid), application (utilities, cities and municipalities, defense, industrial, and others), and geography (North America, Europe, Asia-Pacific, Latin America, Middle East, and Africa). Market sizes and forecasts are provided in USD for all segments.

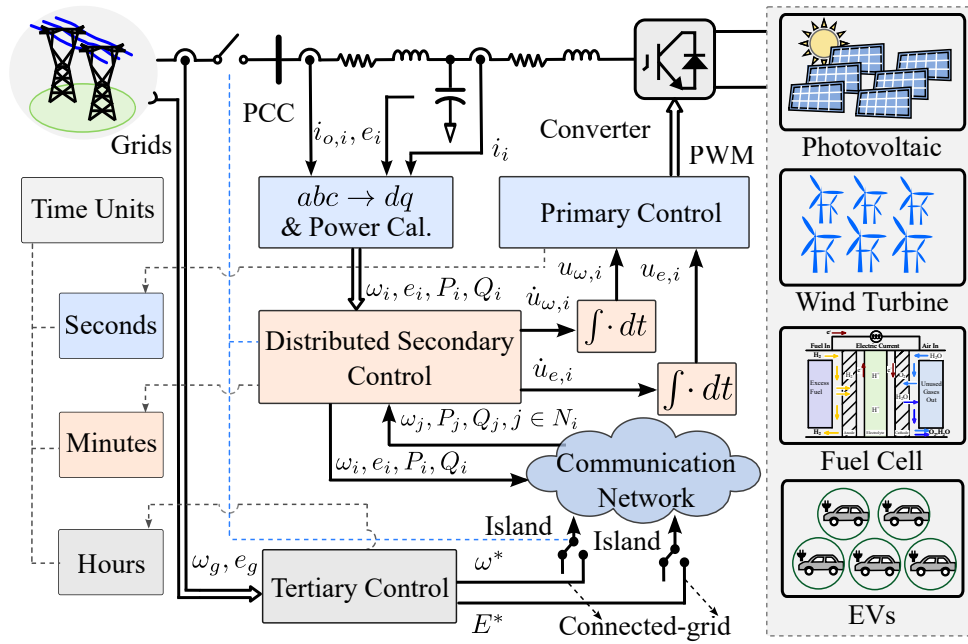


Figure 1.2: A typical hierarchical control structure of an ac microgrid.

1.2 Research Motivations

Power converters ensure efficiency and reliability as a necessary interface for the operation of microgrids [16–18]. Figure 1.3 illustrates an overview of the power converters in microgrids.

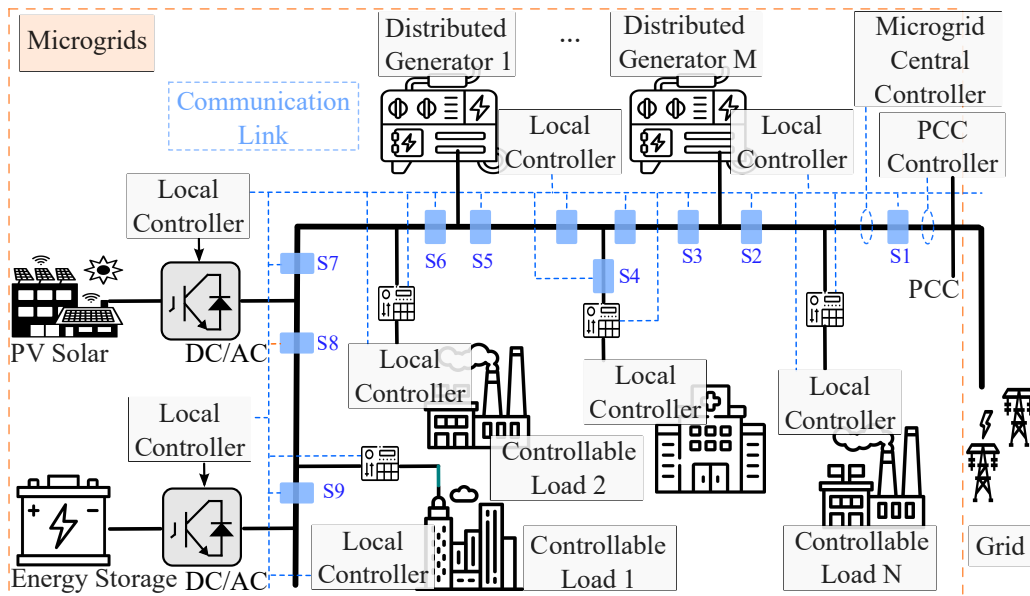


Figure 1.3: An overview of the power converters in microgrids.

1.2.1 3L-NPC Power Converter Topology Benefits

The 3L-NPC power converter is an attractive topology, enabling the integration of renewable energy sources and the management of power quality as shown in Figure 1.4, due to its structure:

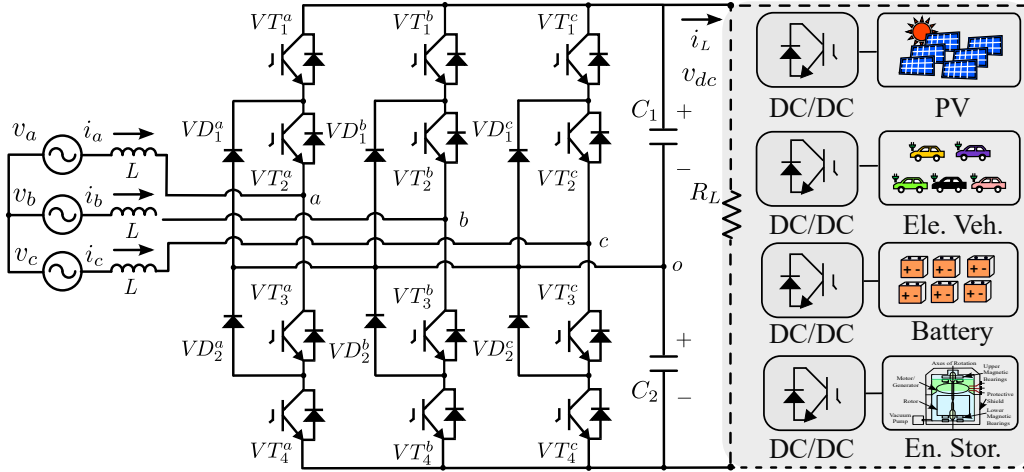


Figure 1.4: A 3L-NPC power converter topology.

1. Power devices withstand a lower shutdown voltage under the same output compared to two-level converters. Moreover, since power devices do not operate simultaneously, the voltage stress and dv/dt values are lower, reducing the harmonic amplitude near the switching frequency.
2. Despite operating at the same switching frequency, the harmonics of the ac-side current are significantly better for 3L-NPC converters than for two-level converters. Due to switching losses and device performance, high-power converters often operate at lower switching frequencies, causing severe harmonics in the output. Therefore, a three-level structure is more suitable than a two-level structure for high-power applications.
3. The 3L-NPC power converters are used widely in high-power fields compared to other multi-level technology converters due to their simpler circuit topologies and control logic.

1.2.2 Key Issues of 3L-NPC Power Converter Control

The 3L-NPC power converters or power generation facilities impose unprecedented control challenges to efficient energy conversion and stable operation, especially existing in multi-disturbances, including internal factors (parameter mismatch, measurement noise, unmodeled dynamics, and bias caused by current path changes and switch mode noise) and external factors (unknown loads, sinusoidal waveform disturbances and frequency variations). Therefore, the primary motivation for further study is to clarify these issues as follows.

1. Current/Power quality

- (a) Switching frequency and harmonics: The switching actions of the power electronic devices in the 3L-NPC converter can introduce high-frequency harmonics into the output current. Poor design or inadequate filtering can exacerbate this, leading to significant harmonic distortion.
- (b) Weak grid conditions: In weak grid scenarios, where the grid impedance is high, the converter's control system might struggle to maintain stable and high-quality power output, leading to voltage sags, swells, or harmonic distortion.

2. Capacitor voltage imbalance

- (a) Midpoint voltage imbalance: The midpoint voltage, ideally half of the total dc-link voltage, can deviate due to various factors such as unequal current drawn from the capacitors and imperfect control strategies.
- (b) Low-frequency ac voltage oscillations: These oscillations can impact the dc-link and cause voltage imbalances. Such oscillations are often tied to the dynamics of the power converter, the loads it serves, and interactions with the grid.

3. DC-voltage regulation

- (a) Nonlinear loads: When the converter supplies power to nonlinear loads (such as rectifiers, variable speed drives, or other power electronic devices), these loads can draw current in a non-sinusoidal manner, leading to harmonic distortion.
- (b) Sudden load changes: Rapid changes in load can cause transient responses in the converter, resulting in temporary distortions in the current and voltage waveforms.

Focusing on these key issues listed in Figure 1.5, a robust and reliable control scheme for 3L-NPC power converters is urgently needed, thus enhancing the efficiency, reliability, and performance of the converter control system in microgrids.

1.3 State-of-the-Art for 3L-NPC Power Converter Control

An effective 3L-NPC power converter control scheme is critical to operational stability and achieving high power conversion efficiency. The underlying control objectives of the 3L-NPC power converter are (at least): dc-link capacitor voltage regulation, power/current tracking, and capacitor voltage balancing [6], [7].

1.3.1 3L-NPC Power Converter Control Objectives

Three control objectives need to be satisfied for the 3L-NPC power converter system [19–22]: power tracking, voltage regulation, and capacitor voltage balancing as shown in Figure 1.6.

1. In the voltage regulation loop, the dc-link voltage v_{dc} is regulated to follow its reference v_{dc}^* . The active power reference p^* , serving as the output of the loop, is considered as the input for the power tracking loop;

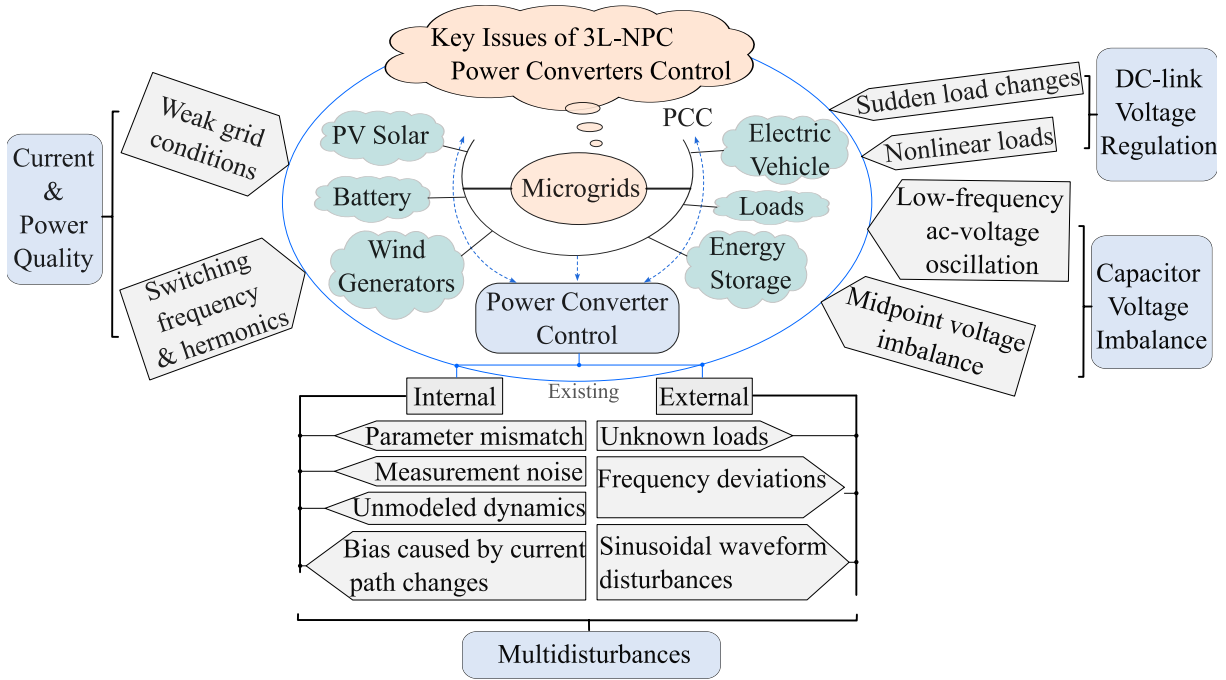


Figure 1.5: The key issues of 3L-NPC power converter control.

2. In the power (or current) tracking loop, the active and reactive powers, denoted as p and q (or $i_{dq}/i_{\alpha\beta}$), are forced to track their references p^* and q^* (or $i_{dq} \rightarrow i_{dq}^*/i_{\alpha\beta} \rightarrow i_{\alpha\beta}^*$). Here, p^* is derived from the voltage regulation loop, while q^* is given to adjust the system power factor;
3. In the voltage balancing loop, the goal is to keep the capacitor voltage difference $v_{dc}^o \rightarrow 0$. The outputs from both this loop and the power tracking loop determine the final average duty cycle, controlling the power converter.

1.3.2 Typical Control Structure of 3L-NPC Power Converter

The power converter control schemes can be categorized into direct power control (DPC) and direct current control (DCC) to meet specific application requirements and achieve the desired performance.

- DPC directly controls active and reactive power, eliminating the need for an intermediate current control loop. This leads to faster dynamic response.
- DCC focuses on directly regulating the output currents of converters, ensuring balanced current sharing among the phases.

Applications requiring fast dynamic response and efficient power conversion benefit most from DPC, while DCC is better suited for applications where precise current regulation and low harmonic distortion are critical.

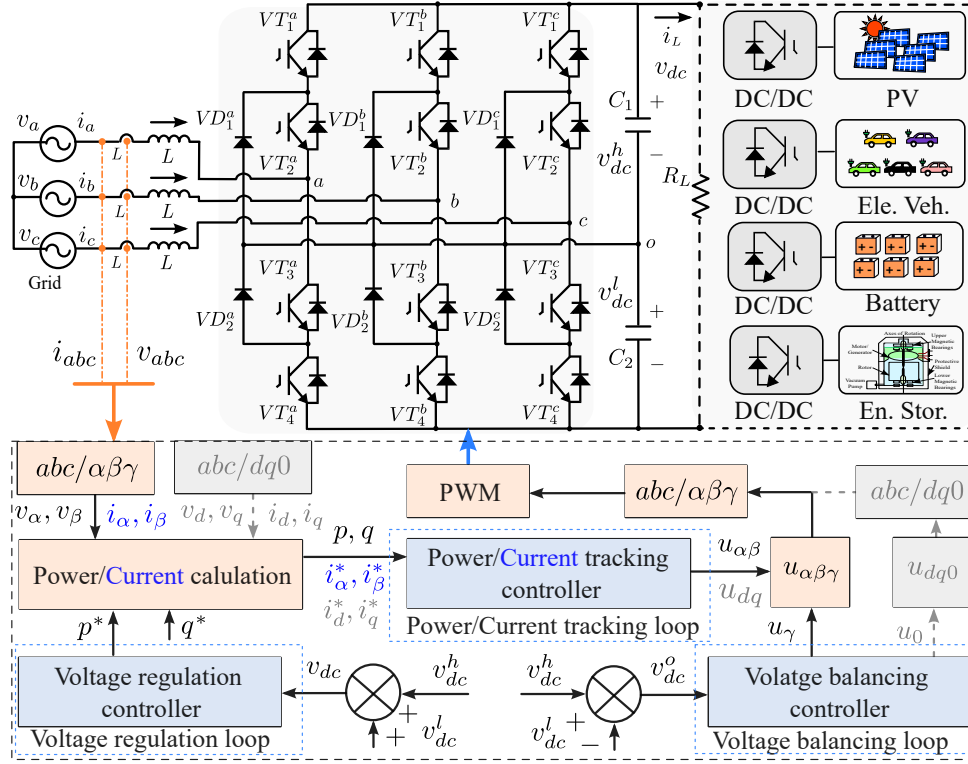


Figure 1.6: Control structure of a 3L-NPC Power Converter

The diverse controller solutions are developed in these control structures to address the key challenges (1.2.2) and achieve control objectives (1.3.1). These control schemes can be broadly categorized into linear and nonlinear methods [23].

1.3.3 Linear Controller

The linear proportional-integral-derivative (PID) control strategy originated in the early 20th century. The first theoretical analysis of PID control can be traced back to the work in the 1920s, which applied it to automatic ship steering systems. With the advent of analog and digital electronics, PID controllers became practical for power electronics during the 1950s-1980s. Today, PID controllers are widely available for use in power electronics.

The PID controller is the most classic and widely used control scheme in industrial applications, known for its simplicity, established theory, and intuitive physical interpretation [24, 25]. Depending on control requirements, various configurations are available, such as proportional control, proportional-integral (PI) control, proportional-derivative (PD) control, and proportional-integral-derivative (PID) control [26, 27]. It is important to note that these control algorithms are applied in the dq -axis. When controlling within the $\alpha\beta$ -axis, alternative controllers such as the proportional resonance (PR) controller are often utilized [28].

1.3.3.1 Proportional Integral Controller

The three-phase ac variables (voltage/current) are transformed into two-phase dc (constant variables) in the dq -axis. The PI controller is commonly employed in dq -control because it can accurately follow a constant signal. Figure 1.7 illustrates the algorithmic structure of the PI controller, where the error signal is e_{dq} , and the control signal is u_{dq} . The transfer function of PI control can be reformulated as

$$G_{PI}^{dq} = k_p + \frac{k_i}{s} \quad (1.1)$$

where k_p and k_i are control gains.

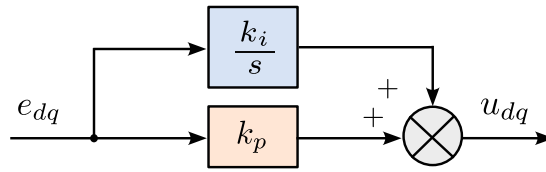


Figure 1.7: PI control structure.

1.3.3.2 Proportional Resonance Controller

In the $\alpha\beta$ -axis, three-phase ac variables are transformed into two-phase ac variables. Thus, a PR controller is applied to track the ac signal. Figure 1.8 depicts the algorithmic structure of the PR controller, in which the error signal is $e_{\alpha\beta}$, and the control signal is $u_{\alpha\beta}$. The transfer function of PR control is obtained as,

$$G_{PR}^{\alpha\beta} = k_p + \frac{k_i s}{s^2 + \omega^2} \quad (1.2)$$

where the k_p and k_i are PR control gains, ω is the grid voltage angular frequency.

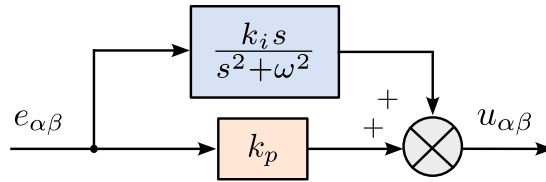


Figure 1.8: PR control structure.

1.3.3.3 Linear Control in Power Converter Application

PID control technology, a well-established method, is crucial in converter control areas [29,30]. A digital multi-loop PID controller, based on loop-shaping of frequency-domain transfer functions, is proposed for dc-dc converters of variable frequency operation, in [31], yielding a variable frequency operation by the on-time and the off-time modulated by the controller. A power controller utilizing feedback from a PID system is commonly presented to withstand a persistent

power supply for ac-dc converters in [32] to compute error in undue power segments between two cycles and regulate the voltage at the load device feedback signal during the remaining periods. Combining a PID controller with modern control technology in renewable energy applications significantly enhances control performance and improves overall system efficiency. In situations requiring specific frequency harmonics processing, the proportional resonance (PR) controller adjusts its resonant frequency based on the fundamental frequency detected by the grid synchronization system [33, 34]. This adjustment enables the PR controller to respond more effectively and compensate for grid harmonics, thus playing a crucial role in maintaining power quality.

As illustrated in the figures 1.9 and 1.10, the PI and PR controllers are employed in the 3L-NPC converter to control the current in microgrids. It is crucial to acknowledge that the efficacy of a linear PID controller depends on an accurate linear model. Nevertheless, its performance may markedly diminish in the presence of parameter mismatches, system nonlinearities, and load disturbances².

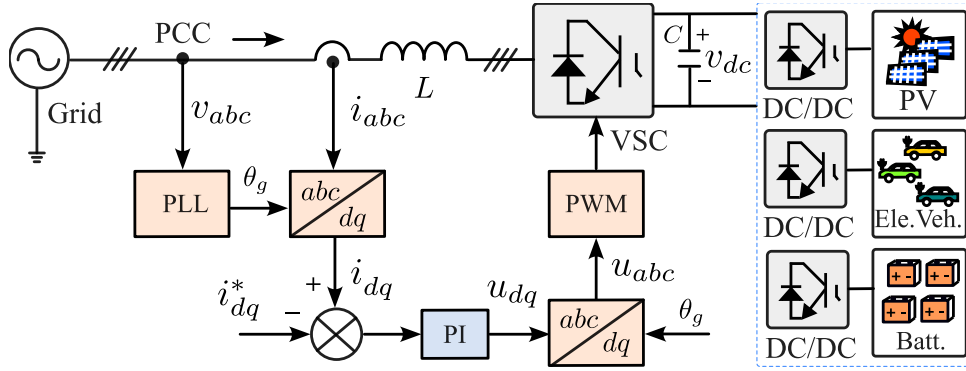


Figure 1.9: PI controller applied in 3L-NPC power converter to control the current in microgrids.

1.3.4 Nonlinear Control

To address the limitations of linear controllers, nonlinear control strategies have been implemented for power converters, encompassing sliding mode control [35–37], adaptive control [38, 39], fuzzy logic control [40, 41], model predictive control [42, 43], event-triggered optimal control [44, 45] and neural network control [46, 47], etc.

Among these nonlinear control strategies, sliding mode control is particularly effective for managing complex nonlinearities and system uncertainties, attributed to its robustness against external disturbances and rapid convergence. Adaptive control automatically adjusts parameters, enhancing its adaptability to system uncertainties and demonstrating strong engineering applicability. Consequently, SMC and AC are critical focuses in power converters.

²These undesirable factors are inherent to power converters, including: 1) The internal parameters of the converter fluctuate within bounds, encompassing filter inductance, phase parasitic resistance, power device switching frequencies, and dc link capacitance. 2) The internal circuit configuration evolves continuously throughout each switching cycle, accompanied by inductor saturation and voltage clamping, which endow the power converter with inherent nonlinearity. 3) The converter consistently encounters unpredictable load fluctuations on both ac and dc sides, potentially causing system disturbances.

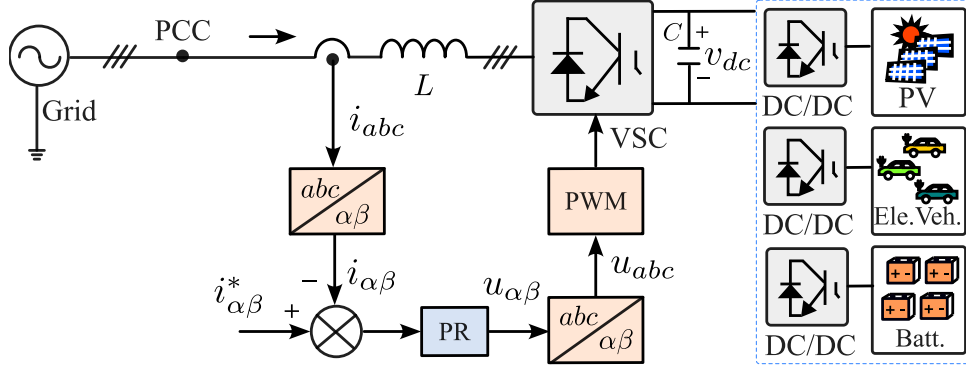


Figure 1.10: PR controller applied in 3L-NPC power converter to control the current in micro-grids.

1.3.4.1 Sliding Mode Control

The concept of sliding mode control (SMC) was initially developed by Soviet scientists, with significant contributions from Emelyanov and V. I. Utkin, who played pivotal roles in the theoretical advancement of SMC in the 1950s. SMC is a nonlinear control method that alters the dynamics of a nonlinear system by applying a discontinuous control signal (or more rigorously, a set-valued control signal) that forces the system to ‘slide’ along a cross-section of the system’s normal behavior in [48].

A dynamical system in sliding mode is defined by:

$$\dot{x} = f(x) = \begin{cases} f^+(x), & s(x) > 0 \\ f^-(x), & s(x) < 0 \end{cases}, x \in \mathbb{R}^n, \quad (1.3)$$

to move on the sliding surface $S = \{x/s(x) = 0\}$, satisfying: if the initial state x_0 is on S , the motion remains on S , i.e., $x_0 \in S$. S is referred to as the sliding mode surface.

Two steps to design an SMC:

1. Design the sliding mode surface S , such that the system state converges to and remains on the sliding manifold.
2. Design the sliding mode controller to ensure that the system state reaches and stays on the sliding manifold S .

Figure 1.11 depicts the sliding surface and two phases of SMC.

(1) An SMC Design: The SMC signal is:

$$u = -k \text{sign}(s(x)) = \begin{cases} -k, & s(x) > 0 \\ k, & s(x) < 0 \end{cases}. \quad (1.4)$$

Thus, under control signal u , the system state can reach the sliding mode surface $s(x) = 0$.

When $s(x)$ is designed to be independent of system perturbations and external disturbances, so the sliding mode of the system exhibits robustness to perturbations and external disturbances,

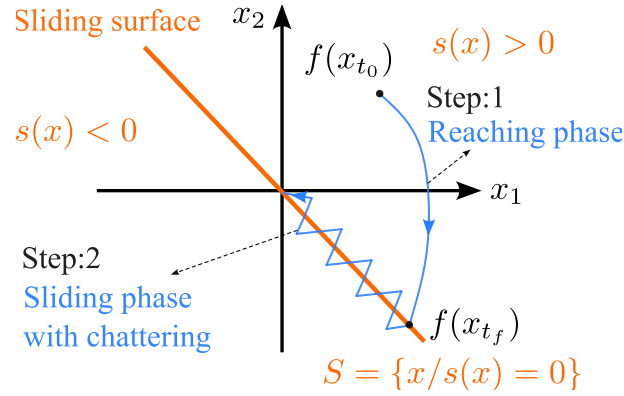


Figure 1.11: SMC depicting sliding surface and two phases.

the SMC suffers from a phenomenon known as chattering—the chattering results in a time delay in the switching signal. The control signal cannot be switched immediately at $s(x) = 0$; it only switches once $s(x) > |\Delta|$ after a period of time delay. Then, the SMC signal is described as follows:

$$u = -k \text{sign}(s(x)) = \begin{cases} -k, & s(x) > \Delta \\ k, & s(x) < -\Delta \end{cases} \quad (1.5)$$

The system trajectory does not fully converge to the sliding mode surface $s(x) = 0$ due to the delay³, instead of forming a limit cycle between the two hypersurfaces $s(x) = \Delta$ and $s(x) = -\Delta$. In the time domain, the limit cycle is characterized by high-frequency oscillations around the origin, as depicted in Figure 1.12.

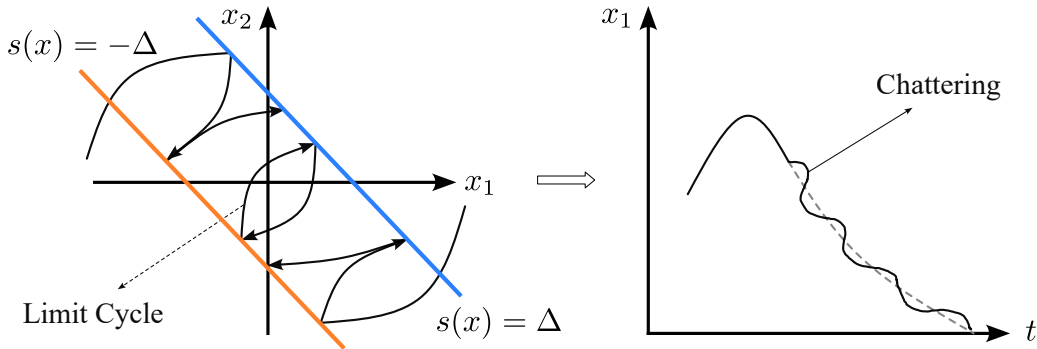


Figure 1.12: Chattering of system trajectory under delayed SMC.

To reduce or eliminate the chattering, various methods are proposed, including integral sliding mode control (ISMC) [49], quasi-sliding mode control (QSMC) [50], filtering-/ observer-based sliding mode control [51, 52], and high-order sliding mode control (HOSMC) methods [53]⁴. At least for now, HOSMC is an effective way to mitigate chattering by integrating and turning

³The main reason causing lag: (1) inertial delay from component response times; (2) sampling delay due to discrete system sampling; (3) dynamic lag from unmodeled system dynamics. Consequently, the delay is inevitable, leading to unavoidable chattering.

⁴HOSMC requires that the switching function $s(x) = 0$, as well as all its derivatives to be zero.

discontinuous control inputs into continuous ones. Admittedly, higher order SMC (generally equal or greater than three orders) is inevitably affected by using higher order time derivatives of sliding variables. Thus, a better alternative is the second-order sliding mode method.

(2) A Second-order SMC Design A second-order sliding mode control (SOSMC) method is considered [54] by the system:

$$\dot{x} = f(t, x, u) \quad (1.6)$$

$$u = U(t, x) \quad (1.7)$$

$$\dot{u} = \Phi(t, x) \quad (1.8)$$

where U is a continuous function, and Φ is a bounded and discontinuous function.

For the sliding function $s(t, x)$, the second-order derivative of $s(t, x)$ can be obtained:

$$\ddot{s}(t, x) = \frac{\partial \dot{s}}{\partial t} + \frac{\partial \dot{s}}{\partial x} f(t, x, u) + \frac{\partial \dot{s}}{\partial u} \Phi(t, x) \quad (1.9)$$

Let $h(t, x) = \frac{\partial \dot{s}}{\partial t} + \frac{\partial \dot{s}}{\partial x} f(t, x, u)$, $g(t, x) = \frac{\partial \dot{s}}{\partial u}$, $u_s = \Phi(t, x)$.

An SOSMC algorithm based on (1.8) can be designed as follows:

$$\Phi(t, x) = \begin{cases} -k_m \text{sign}(s(t, x)), & s\dot{s} \leq 0, \\ -k_M \text{sign}(s(t, x)), & s\dot{s} > 0, \end{cases} \quad (1.10)$$

where k_m and k_M are controller parameters determined by the system parameters and initial conditions, with $k_m > 0 > k_M > 0$. Under the SOSMC method, the system trajectory converges to the super-twisting surface $s = \dot{s} = 0$, as shown in Figure 1.13.

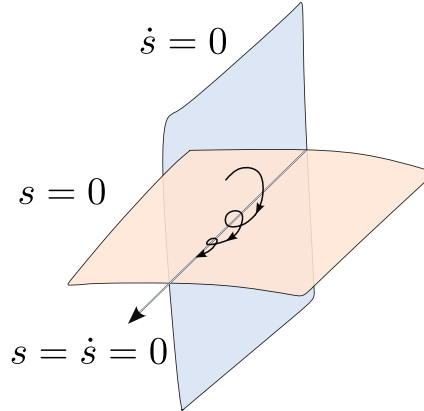


Figure 1.13: System trajectory under SOSMC.

(3) A Super-twisting Algorithm Control Design: A Super-twisting Algorithm (STA) control is one of the SOSMC methods introduced to suppress the chattering effects [55], and the structure of the algorithm is shown in Figure 1.14. The STA-based control is expressed as:

$$u = -\lambda |s|^{\frac{1}{2}} \text{sign}(s) - \int \varphi \text{sign}(s). \quad (1.11)$$

It can be observed that the STA design requires merely the switching function information without its derivative information.

Assuming $0 < G_m < g(t, x) < G_M$, $-H < h(t, x) < H$, the parameters of the STA should satisfy:

$$\varphi > \frac{H}{G_m} \quad (1.12)$$

$$\lambda^2 > \frac{4H}{G_m^2} \frac{G_M \varphi + H}{\varphi - H}, \quad (1.13)$$

based (1.8), then it can be obtained that

$$\Phi(t, x) = -\frac{\lambda}{2} |s|^{\frac{1}{2}} \dot{s} - \varphi \text{sign}(s), \quad (1.14)$$

where $\Phi(t, x)$ is a discontinuous function, which ensures that the system trajectory reaches the hyperplane of $s = \dot{s} = 0$.

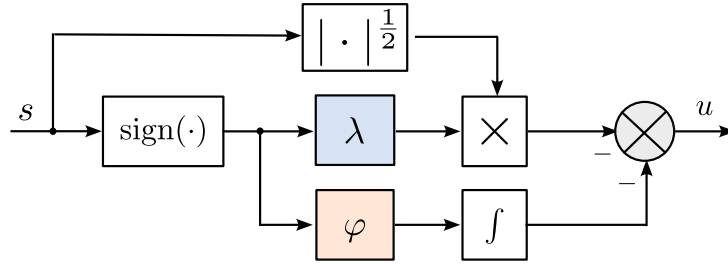


Figure 1.14: STA control structure.

1.3.4.2 Sliding Mode Control in Power Converter Application

SMC methods have many applications [56, 57]. SMC is more effective in achieving the desired dynamic response and robustness for power converters due to its fast dynamic response, disturbance rejection, and simple implementation [58–60]. The power converters controlled by SMC methods include dc-dc converters [61], dc-ac inverters [62], ac-dc rectifiers [63], ac-ac converters [64]. In [65], an SMC method is introduced for designing a current compensator for a parallel-connected converter system. However, the SMC method suffers from chattering in the control reference input, which is not preferred in practice.

To improve the chattering problem, an SMC with a boundary layer technique is utilized to enhance the robustness of the speed control scheme of a permanent magnet synchronous motor in [66]. The boundary layer approach is also a good solution for suppressing chattering. However, the boundary layer approach creates a trade-off between control performance and chattering elimination. [67] proposed a terminal SMC method for the voltage controller of the buck converter, which improved the output voltage response speed by reducing chattering and enhancing the output voltage response speed. Nevertheless, when nonlinear terms are introduced into this control, the singularity of the control input may occur under certain conditions⁵. [68]

⁵For example, a terminal SMC introduces nonlinear terms such as $x^{q/p}$, where x is a state variable, p and q are positive odd numbers, and $p > q$. As $x \rightarrow 0$, if q/p is a proper fraction, then $x^{q/p} \rightarrow \infty$, which may cause the control input to become infinite.

proposes a fuzzy-fractional-order nonsingular terminal sliding-mode controller for the grid current control of LCL–grid-connected converter systems. Using this method, not only is the singularity problem of the terminal sliding mode control resolved, but the tracking performance is also significantly improved through applying the fuzzy controller, which obtains the optimal fractional-order value.

In [69], a modular parallel-connected single-phase inverter is designed using a fixed-switching-frequency QSMC based on zero average dynamics. One of the QSMC benefits is a sufficient and necessary condition for proving discrete-time system stability. Thus, a new quasi-integral sliding mode control (QISMC) has been developed, which achieves high-precision control by adding the integral term to the quasi-sliding mode (QSM) surface in [70]. However, significant disturbances may result in system instability or require higher control gains that lead to the chattering.

At least for now, a HOSMC is an effective way to mitigate the chattering by integrating and turning discontinuous control inputs into continuous ones. Higher-order sliding-mode control (generally equal or greater than three orders) is inevitably affected using higher-order sliding variables' time derivatives. Thus, a better alternative is the second-order sliding mode method. Like, [71] developed a SOSMC to regulate the output voltage of a dc-dc buck converter, estimating the system uncertainties and external disturbances.

An STA is one of the SOSMC methods introduced to suppress the chattering effects, handle systems with uncertainties and disturbances, and offer robustness and high performance. [72] proposes and implements an STA direct power control strategy for a brushless doubly-fed induction generator. In [73], an STA control method for regulating the output voltage of a dc-dc buck converter is proposed, which enables the fastest convergence, the shortest settling time under loaded conditions, and the lowest steady-state error during load changes of all contrasted control methods. Unlike the conventional first-order SMC, STA has a finite-time convergence with continuous control input and strong robustness under high uncertainties. However, a large disturbance may result in system instability or require higher STA controller gains, leading to system chattering. Hence, it is necessary to design an observer to estimate disturbance as feedforward compensation in control input to alleviate the system chattering.

1.3.4.3 Adaptive Control

Adaptive control (AC) studies a class of nonlinear control methods that deal with systems with uncertainty (Dubowsky and Deforges 1979; Goodwin and Sib 1984; Craig 1987) [38, 74]. AC system can be broadly described as any control system that can adjust control design parameters, such as control gains online based on inputs received by the plant to accommodate system uncertainty, as illustrated in Figure 1.15.

AC aims to achieve or maintain optimal performance for systems with unknown or time-varying model parameters and external disturbances by adjusting the control parameters in real-time. AC systems are mainly divided into two categories: (1) Direct AC (typical structure: model reference adaptive control (MRAC)); and (2) Indirect AC (typical structure: self-tuning control systems). AC structures that combine both types are frequently used and are referred to as composite, combined, or hybrid direct–indirect adaptive control [75, 76].

A typical direct AC may be expressed as

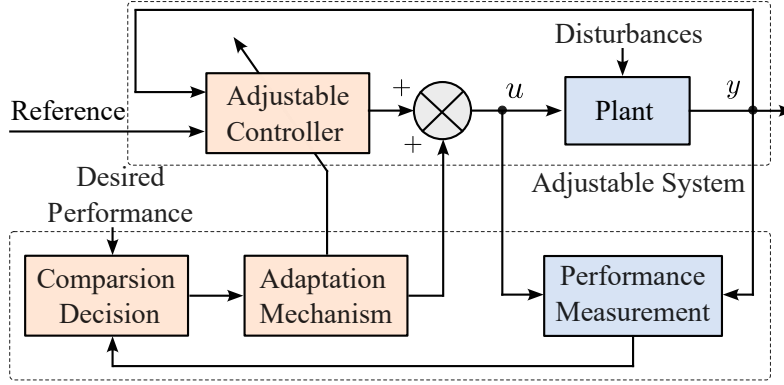


Figure 1.15: AC structure.

$$u = k_x(t)x + k_r(t)r, \quad (1.15)$$

where $k_x(t)$ and $k_r(t)$ are adjustable control gains, x is the system state, r is the reference input, and u is control signal.

Direct AC adjusts a feedback control directly to cancel out unwanted system uncertainties, allowing the control system to maintain performance despite significant uncertainties. In contrast, an indirect AC achieves the same objective by adjusting the control gains indirectly, which can be expressed as

$$u = k_x(p(t))x + k_r(p(t))r \quad (1.16)$$

where $p(t)$ are system parameters estimated online to update the control gains.

(1) MRAC Systems A typical MRAC system block diagram is shown in Figure 1.16. This system employs a reference model as an auxiliary system, designed to achieve desired performance indicators through its output or state. The input signal is simultaneously fed into the reference model and the adaptive control system. The difference between the outputs of the two systems generates the error signal e . The error signal e is fed into the adaptive control system through an adaptive mechanism to adjust the controller parameters or generate an auxiliary input signal, minimizing the integral performance index of the error. In this way, the output of the AC system can asymptotically approach the ideal model output, achieving the control objective.

(2) Self-Tuning Control Systems Figure 1.17 illustrates a typical Self-Tuning Control system structure. This system performs online estimation of uncertain parameters of the controlled object, including unknown or time-varying model characteristics and external disturbances. The estimated values of the uncertain parameters $\hat{\theta}$ are then input into the controller. The controller adjusts the control parameters based on these estimates and predefined performance indicators, with the resulting control signal acting on the controlled object. The system achieves the desired performance indicators through multiple iterations of parameter estimation and control parameter adjustment.

In addition, according to how closely are related to a physical model, the AC theory can be categorized into:

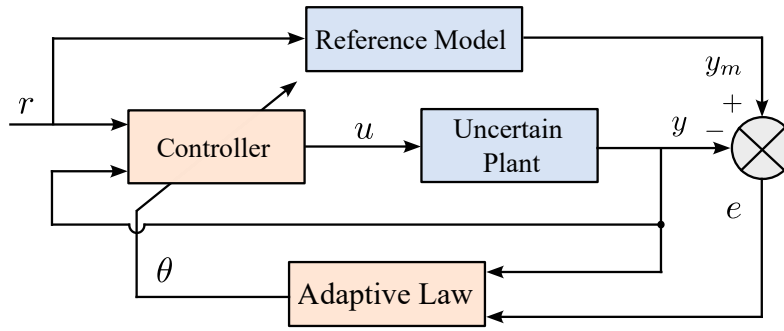


Figure 1.16: An MFAC system.

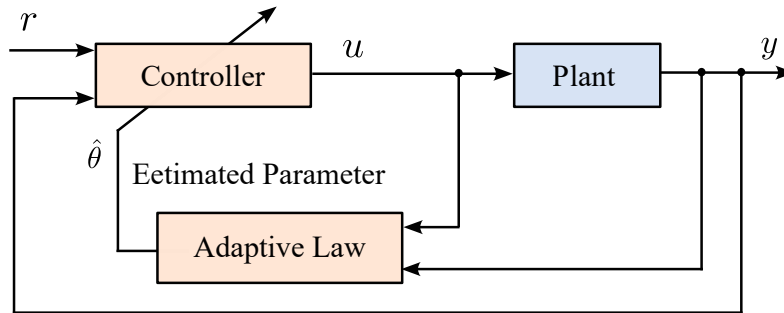


Figure 1.17: A STC system.

1. Model-based adaptive control (MBAC) utilizes physics-based models of the controlled system.
2. Model-free adaptive control (MFAC) relies merely on the interaction of controller with the system.
3. Learning-based adaptive control integrates model-based and model-free approaches to create adaptive controllers.

A physical model is used to categorize adaptive control in Figure 1.18, where an MBAC block ensures stability and an MFAC block optimizes performance through real-time parameter adjustment. MFAC is increasingly in demand. It offers the features: (i) precise control of system dynamics and stability performance without requiring knowledge of the plant model; (ii) ensuring consistent operation even in the face of model variations.

1.3.4.4 Adaptive Control in Power Converter Application

AC is widely used in power converters due to its robustness and flexibility [39, 77–79]. A proposed model-based adaptive direct power control for 3L-NPC converters ensures voltage regulation and balance across the dc-link capacitors while enabling active and reactive power tracking for both rectifier and inverter modes in [19]. Even though this study overcomes problems associated with system parameter uncertainties, the system could still display instability in the presence of disturbances. [80] introduces an adaptive PI control strategy for a wind energy conversion system that utilizes a permanent magnet synchronous generator and a pulse-width

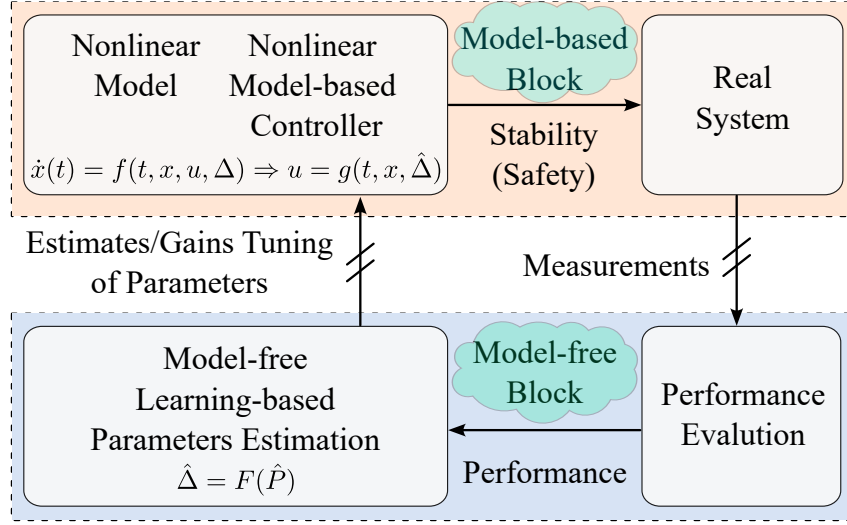


Figure 1.18: The categorization of AC based on physical model.

modulated current source converter, enhancing efficiency and reliability. It is a self-tuning control based on a linear approximation of the power system and a desired closed-loop response. However, the nature is still the PI control and PI control can cause system instability when dealing with large uncertainties. [81] adopts adaptive PR control, automatically adjusting the resonant frequency control parameters based on the current error. A potential limitation of this control strategy is that it may over-rely on the current error signal to adjust the resonant frequency control parameters automatically. According to literature [82], sliding mode adaptive backstepping control is applied to PWM ac-dc converters with three-phase voltage sources to cope with uncertain parameters in the system, but the complex control structure does not lend itself to engineering applications. Adaptive neural network control is used to enhance photovoltaic conversion efficiency by utilizing a dc-dc buck converter in [83]. However, since the method relies heavily on data, it is necessary to carefully tune the learning rate of the neural network in order to minimize the possibility of overfitting.

It should be noted that MBAC usually relies on assumptions and conditions, which may not produce satisfactory results when the model accuracy is insufficient, especially in complex nonlinear systems. Therefore, MBAC has been proposed and applied in power converters. An MFAC only requires input/output data, and its pseudo partial derivative property ensures that it is not affected by system variations, thus proving its robustness. [84] combines data-driven with dual droop control for ac/dc interlink converter to achieve effective power sharing. Despite reducing reliance on accurate models, this control method may need fixing due to input coupling and operational constraints in complex systems. The robust predictive control employs data-driven neural predictors to improve the issue, as presented in [85]. This method leverages real-time and historical data to estimate unknown system dynamics, thereby boosting control robustness and flexibility. It may still be necessary to carefully tune the architecture and learning parameters of the neural network.

1.3.4.5 Other Nonlinear Control in Power Converter Application

Nonlinear control shows great potential and advantages in power converter applications. [86] proposes a finite-set model-predictive control (FCS-MPC) strategy for 3L-NPC inverters, producing fixed switching frequency without pulse width modulation (PWM) as well as maintaining the benefits of predictive control such as fast dynamic response, nonlinearities and restrictions, and multivariable control. To reduce calculation efforts of FCS-MPC without affecting control performance, [87] proposes a simplifying FCS-MPC, utilizing matrix converter-fed permanent magnet synchronous motors. However, both of these methods are highly dependent on parameters.

[88] proposes an adaptive fuzzy droop control for multi-microgrid systems with constant and resistive power loads to increase the accuracy of power allocation and the system stability margin. To enhance the efficiency of hybrid energy storage systems in solar electric vehicle charging stations, [89] introduces an interval type-2 fuzzy logic controller-based power-sharing strategy.

With the rise of artificial intelligence algorithms, a novel neural network-based control strategy is presented in [90] that enhances three-phase grid-tied converter performance and stability and provides superior rapid response and robust disturbance rejection to conventional vector control techniques in renewable energy and power systems. [91] uses an adaptive dynamic programming-trained recurrent neural network to approximate optimal control for single-phase inverters with LCL filters, demonstrating superior performance over conventional methods by allowing lower sampling and switching rates without compromising control effectiveness. There are, however, some challenges associated with these two types of control, such as the large requirement for computing resources and the long durations.

It is still a challenge to thoroughly validate the robustness of these methods to large disturbances and uncertainties. Figure 1.19 illustrates a nonlinear power controller applied in a 3L-NPC power converter in microgrids.

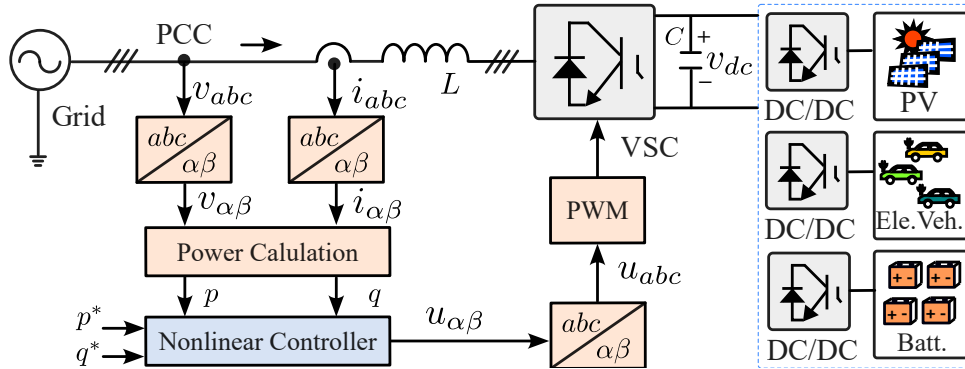


Figure 1.19: A nonlinear power controller applied in a 3L-NPC power converter in microgrids.

1.3.5 Observer-based Control

Feedback control (FBC) systems may not rapidly respond to severe disturbances. Consequently, feedforward compensation control (FFC) is proposed to directly reject disturbances measurements or estimations. Nonetheless, the application of FFC is constrained by the issues of unmeasurable or difficult-measurable disturbances. To utilize the benefits of FFC while addressing the limitations of FBC, observer-based control (OBC) has garnered considerable interest

and application. This motivation forms the theoretical basis of the research conducted in this study. A block diagram of OBC is shown in Figure 1.20. Observers can be categorized into state observer (SO) and disturbance observer (DO), as shown in Figure 1.21. Among them, SO includes extended state observer (ESO), Luenberger observer (LO), reduced-order observer (ROO), sliding mode observer (SMO), adaptive observer (AO), super-twisting observer (STO), etc. The comparison of SO technologies for power converters is listed in Table 1.1. In addition to perturbation observer (PO), unknown input observers (UIOs), improved equivalent input disturbance (I-EID) based-estimator and extended sliding mode disturbance observer (ESMDO). The comparison of DO technologies for power converters is listed in Table 1.2.

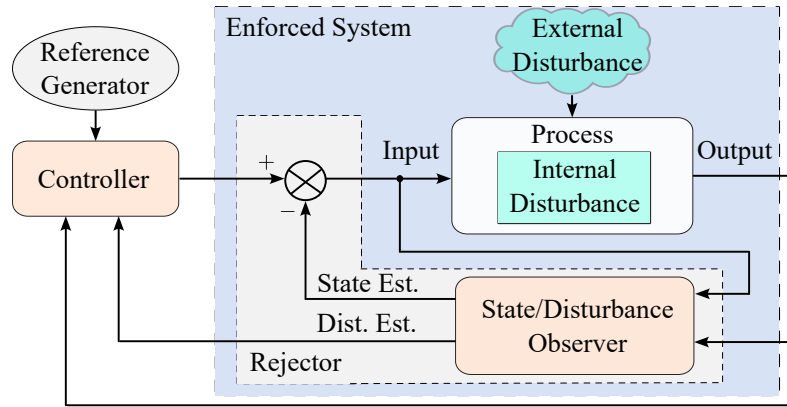


Figure 1.20: A block diagram of OBC structure.

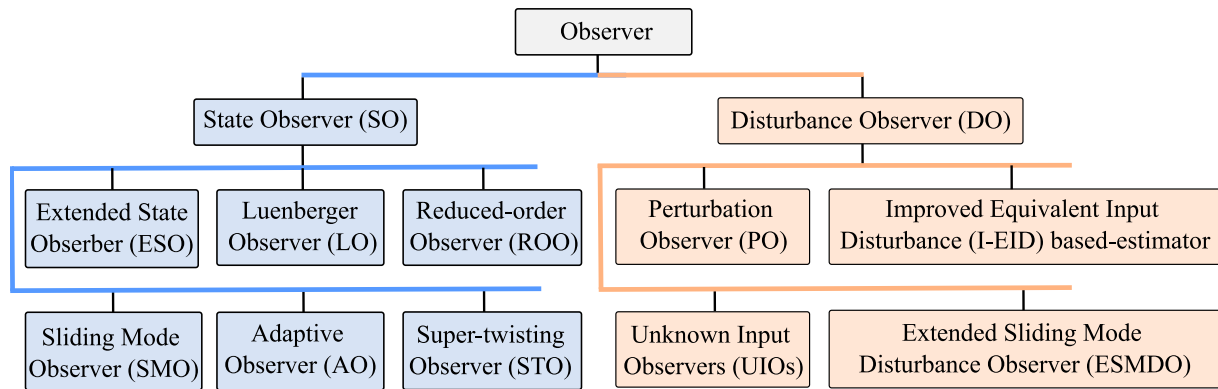


Figure 1.21: Typical classification of observers.

1.3.5.1 State Observer

A state observer is a system that estimates the internal state of a given real system from measurements of the input and output of the real system.

Consider a continuous-time linear system

$$\dot{x} = Ax + Bu, \quad (1.17)$$

$$y = Cx \quad (1.18)$$

Table 1.1: Comparison of SO Technologies for Power Converters

	Observer Type	Benefits	Drawbacks	Applicability
SO	ESO Linear	Easily analyzed and designed	1. Ultimately bounded; 3. Disturbance dynamics assumption.	[92]: Speed tracking in PMSM; [93]: Voltage regulation in AC/DC converters.
	LO Linear	Simple implement and debug	1. Parameter dependence 2. Full-order SO.	[94]: Imbalance of capacitor voltages.
	ROO Linear	1. Low computing 2. More efficient; 3. Nature of LO.	1. Linear systems; 2. Parameter dependence.	[95]: Imbalance of capacitor voltages.
	SMO Nonlinear	2. Robust to uncertainty; 3. Estimated online; 4. Asymptotically estimate disturbances; 5. Reduced one order, when entering the sliding mode; 6. Finite-time convergence	1. Chattering risk with improper setup; 2. Design complex.	Imbalance of capacitor voltages in this work.
	AO Nonlinear	1. Robust to uncertainty; 2. Estimation in real-time; 3. Asymptotic convergence.	Need Assumptions.	Unknow dc-loads estimated in this work.
	STO Nonlinear	1. Benefits of SMO; 2. Reducing chattering	Sensitive to noise.	DC-voltage regulation and power tracking in this work.

where x is the system state, u is the system inputs, and y is the system outputs. A , B , and C are system matrices. If the system is observable then the output of the plant, y can be used to steer the state of the state observer.

Assume that the observer is as follows,

$$\dot{\hat{x}} = A\hat{x} + Bu + L(y - \hat{y}), \quad (1.19)$$

$$\hat{y} = C\hat{x}, \quad (1.20)$$

where $\hat{x}$ represents the estimate for x , and $\hat{y}$ represents the estimate for y . L stands for the observer gain.

The estimated goal of the observer is $\hat{x} \rightarrow x$, i.e.,

$$\dot{\hat{x}} - \dot{x} = (A - LC)(\hat{x} - x), \quad (1.21)$$

The stability of the state observer can be guaranteed that $(A - LC)$ is the Hurwitz matrix, i.e., every eigenvalue has a strictly negative real part. Therefore, it can be concluded that, i.e., $\hat{x} \rightarrow x$, i.e., $\hat{x}$ can be estimated as x . The full-order state observer is described as follows, and its structural block is shown in Figure 1.22

Table 1.2: Comparison of DO Technologies for Power Converters

	Observer Type	Benefits	Drawbacks	Applicability
DO	PO Linear	Multi-disturbances	1. Assume certain dynamic properties; 2. Asymptotically bounded estimation.	[96]: Power decoupling control in DFIG wind turbine.
	UIOs Linear	Applied to decoupled estimation	1. Assumptions for system excitations; 2. Sensitivity to parameters.	[97] : Inter-connected power systems
	(I-EID) based -estimator Linear	Improving robustness	1. Computational complexity; 3. Increase response time; 4. Not suitable for all scenarios.	[98]: Position tracking in drive motor
	DO Linear /Nolinear	1. lumped disturbances; 2. Asymptotic estimate or asymptotically bounded estimation; 3. Online estimation.	1. Disturbance dynamics assumption; 2. Rather than ‘worst-case’ design.	Rejecting disturbances of power converter in this work.
	ESMDO Nolinear	1. Robustness; 2. Fast estimation; 3. Online estimation 4. Lumped disturbances.	Chattering risk.	Rejecting lumped disturbances for dc-voltage of converter in this work.

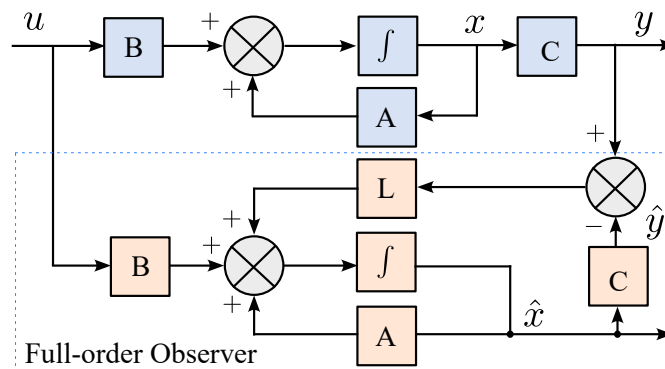


Figure 1.22: A full-order state observer.

1.3.5.2 Disturbance Observer

Consider a linear system with disturbances,

$$\begin{cases} \dot{x} = Ax + B_u u + B_d d, \\ y = Cx \end{cases} \quad (1.22)$$

where x is the system state, u is control input, d represents disturbance, and y is system output. A , B_u , B_d , and C are system matrices.

The DO can be designed to estimate the disturbances of (1.22) by [99],

$$\begin{cases} \dot{z} = -LB_d(z + Lx) - L(Ax + B_u u) \\ \hat{d} = z + Lx \end{cases} \quad (1.23)$$

where $\hat{d}$ is the disturbance estimation, z is the internal variable of the DO, and L is the DO gain.

The estimation error of DO is defined as

$$e_d = \hat{d} - d \quad (1.24)$$

Taking the derivative of (1.24), one get

$$\dot{e}_d = -LB_d e_d - \dot{d}. \quad (1.25)$$

When the observer gain matrix $-LB_d$ is Hurwitz, then disturbance estimation error system (1.25) is bounded-input bounded-output stable. In addition, if the disturbances tend to be constants, (1.25) is asymptotically stable with appropriately chosen parameter L such that $-LB_d$ is Hurwitz.

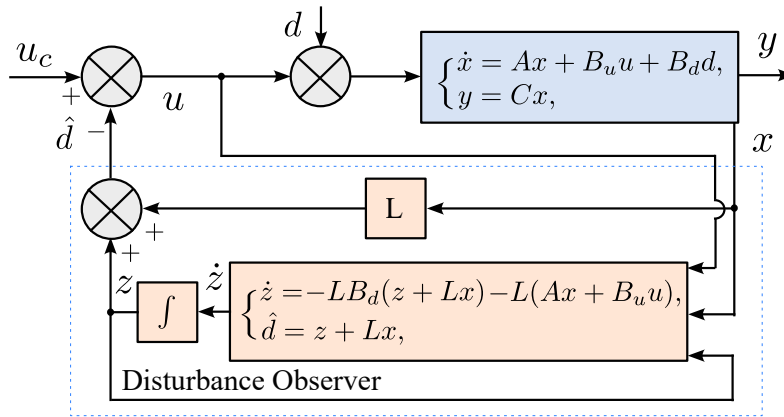


Figure 1.23: A disturbance observer.

1.3.5.3 Observer-based Control in Power Converter Application

In power converter applications, many effective techniques for estimating uncertainties and disturbances exist. As described in [92], an ESO-based SMC is proposed to effectively track input currents and regulate output voltage while rejecting external disturbances. Similarly, a

robust controller based on ESO is proposed for a dc-dc interleaved boost converter for fuel cell application. The uncertainties and external disturbances of the converter are treated as the total disturbance, which is estimated by ESO and then eliminated in the control law. Similarly, [93] proposes an ESO-based robust controller for a dc-dc interleaved boost converter of a fuel cell that addresses uncertainties and external disturbances by integrating them as a total disturbance, estimated by the ESO and then mitigated within the control law, thereby enhancing converter performance and stability. Although ESO has a good anti-disturbance ability against time-invariant disturbance, the disturbance can not be achieved through asymptotic estimating. The proposed maximum power point tracking technique for a photovoltaic system efficiently tracks maximum power without a current sensor by employing a LO in [94], reducing costs, communication burdens, and system complexity. However, the LO is a full-order state observer. A ROO- H_∞ controller is designed to balance the capacitor voltage in [95]. Since the nature of the reduced-order observer is still an LO, the high gain can result in a peaking phenomenon where the initial estimate error is considerably large. A SMO is available, which brings one estimated error to zero in finite time, even in the presence of measurement noise, and the other states have the error that behaves similarly to the error in a LO after peaking has subsided [100]. Furthermore, the SMO boasts rapid convergence, exceptional disturbance rejection, and straightforward implementation. On the other hand, an AO serves as an estimation mechanism where both the parameters and state variables of the system are simultaneously estimated.

Disturbance observer-based control (DOBC) has received a great deal of attention. A PO-based SMC method offers robust control for high voltage direct current systems based on voltage source converters by estimating and compensating for uncertainties and disturbances in [96]. [97] paper introduces a fully decentralized load frequency control based on dynamic state estimation, utilizing a UIO for each power system area to track the dynamic states in a real-time operation to enhance robustness. However, both methods face challenges in achieving asymptotic tracking and require more dynamic assumptions. To reduce constraint issues, a new filter is devised to improve an EID-based estimator in [98]. It is evident that this approach will increase the complexity of the design. [101] presents a strategy for direct current control of a three-phase, two-level power converter using the DO to enhance disturbance rejection. In addition, an ESMDO has the robustness of SMO and has a good disturbance estimation ability but also has an inbuilt low-pass filter without introducing a phased delay in [102]. Figure 1.24 shows the observer-based controller used in a 3L-NPC converter in microgrids.

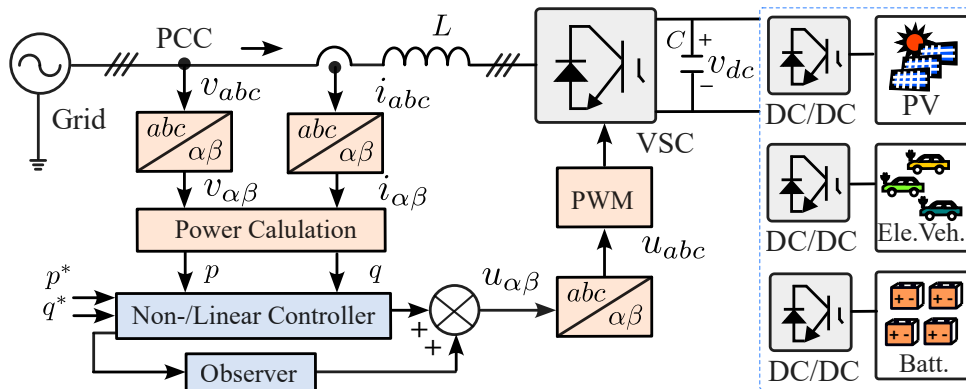


Figure 1.24: OBC applied in a 3L-NPC power converter in microgrids.

1.4 Outline

Following this introduction, Ch. 2 provides a detailed analysis of the topology and operating principles of the 3L-NPC power converter. Based on these, the mathematical model of the 3L-NPC converter is built. Moreover, the typical pulse width modulation techniques for the 3L-NPC power converter are presented.

High penetration of power converters may lead to power ripple, voltage swings, and weak anti-disturbances in microgrids. Thus, reliable control schemes are critical to ensuring converter operation in microgrids. Ch. 3 proposes a robust control scheme, discrete-time super-twisting observer (DSTO)-embedded quasi-integral sliding mode control (QISMC), for a three-level NPC power converter system, dramatically enhancing power/voltage regulation performance and anti-disturbance capability. Attractive features: A fast convergence DSTO is deployed to offset multi-disturbances caused by parameter mismatches, unknown loads, current path changes, switch mode noise, and self-compensating power/voltage tracking biases in QISMC. To further mitigate power/voltage steady-state error and boost system robustness, a new quasi-integral sliding mode surface is built, inherently improving power/voltage tracking performance. Experimental data confirm the proposed control outperforms the discrete-time extended state observer-QISMC, DSTO-quasi-sliding mode control, and discrete-time PI control in power/voltage, grid current harmonics, and robustness.

To customize designs and optimizations of converter control systems outlined in Ch. 3. Additionally, Ch. 4 explicitly improves the capacitor voltage balancing issue, further enhancing system stability and reliability. In detail, Ch. 4 proposes a robust adaptive sliding mode control for the 3L-NPC power converter. An adaptive observer (AO)-based proportional (P) voltage controller is proposed to accurately and quickly regulate dc-voltage in real-time identifying unknown equivalent dc-loads, compensating for the active power reference. To track the reference, a disturbance observer (DO)-based integral Sliding mode control (ISM) is adopted to dramatically enhance the power tracking performance in case of parameter mismatches and bias caused by current path changes and switch mode noise. Additionally, an sliding mode observer (SMO) coupled with P control is established to balance dc-link (*two*) capacitors, rejecting harmonic injection and power ripple behaviors. Experimental data confirm that the proposed control scheme outperforms super-twisting observer (STO)-based ISM, super-twisting algorithm (STA), and proportional-integral (PI) control schemes in the transient/steady state operations in terms of dynamic response, grid current harmonic distortions, and robustness.

Designing a suitable high-quality current control scheme featuring superior dynamics, strong robustness, and simple structure is a promising solution to guarantee the more efficient operation of the converter. Ch. 5 proposes a robust high-quality current control strategy for the 3L-NPC power converter in the stationary $\alpha\beta$ frame. An STA coupled with a Luenberger observer (LO) current controller is proposed to deal with the poor sinusoidal current tracking issue due to the existing inductance/grid frequency deviations and the disturbance of the sinusoidal dynamic nature. The LO is designed to offset the disturbance, alleviating the system chattering. The reason is that the STA switching gain is simply selected to be bigger than the bound of the sinusoidal disturbance estimation error and parameter deviations rather than that of the lumped disturbances. Additionally, an extended sliding mode disturbance observer (ESMDO)-based P control is built to dramatically enhance voltage regulation performance, in the case of capacitance deviations and unknown dc-loads. The ESMDO is designed to introduce the sliding mode reaching law to

control the voltage convergence rate, reject uncertainties, and identify loads. Experimental data confirm the effectiveness of the proposed solution outperforms the conventional proportional resonance (PR) and PI control in terms of accurate tracking current/voltage, anti-disturbance, and grid current total harmonic distortion.

Although model-based control (MBC) is widely used in converters due to its ease, it poses challenges when dealing with inaccurate models, critical loads, and multidisturbances. To deal with these issues, Ch. 6 proposes a data-driven model-free adaptive control (MFAC) for a 3L-NPC power converter system to realize robust high-quality current and voltage control without any system parameters. Specifically, an MFAC is a potential approach to estimating the system dynamics by designing adaptive laws derived intuitively from Lyapunov theory to regulate current and voltage, guaranteeing self-adaptability to unmodeled dynamics, model variations, parameter mismatches, and disturbances. Striving for easier access to adaptive law gains, a particle swarm optimization (PSO) algorithm is embedded in MFAC to automatically determine them utilizing the fitness function defined by the voltage and current. Experimental data confirm that the proposal outperforms artificial neural networks-based PI (ANN-PI) and model-based STA control schemes under multi-scenarios in terms of transient-/ steady-state, grid current harmonic distortions, and robustness.

Ch. 7 summarizes and outlooks the work of the 3L-NPC power converter control in microgrid. Additionally, a block diagram of the entire workflow is shown in Figure 1.25.

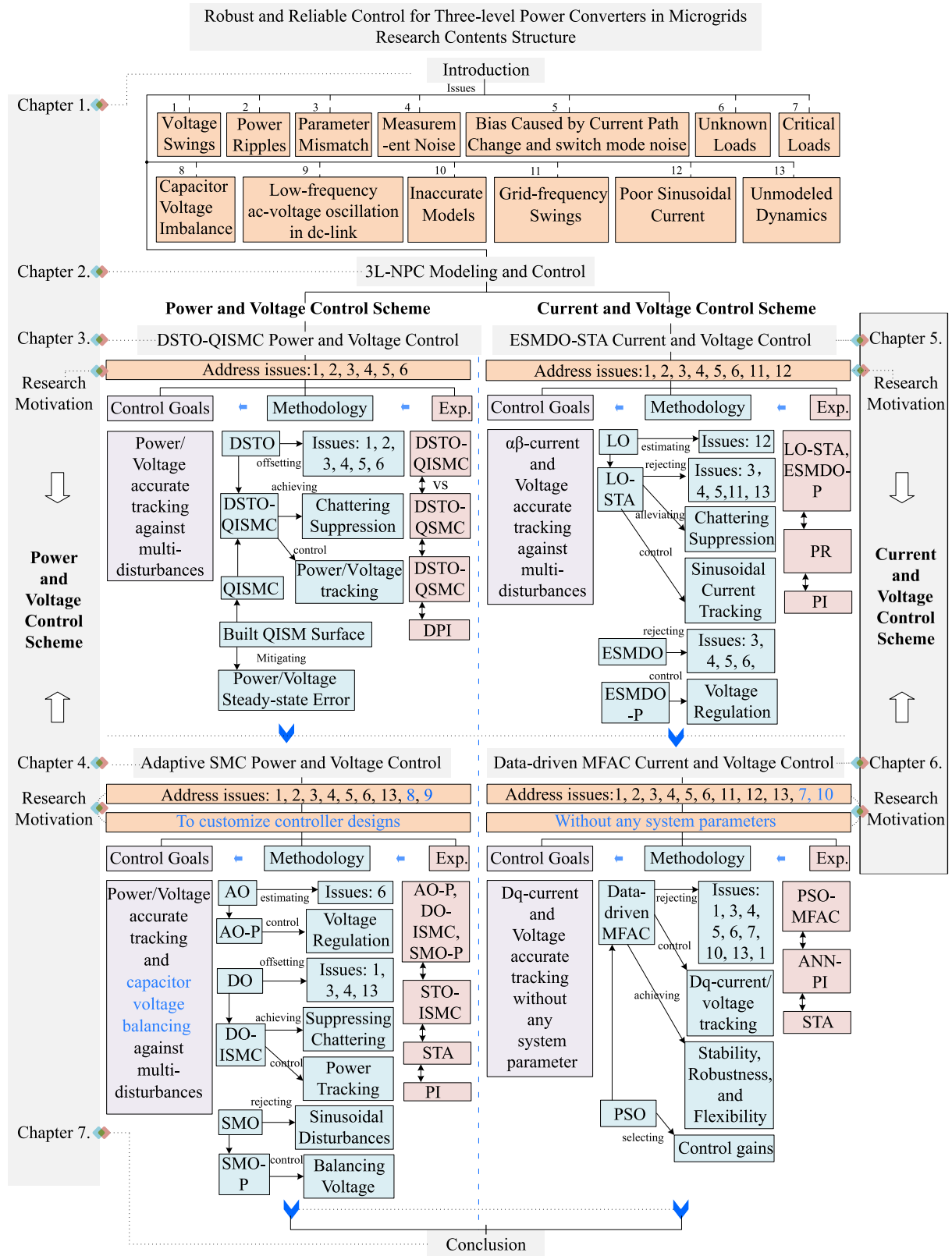


Figure 1.25: A block diagram of the entire workflow.

CHAPTER 2

3L-NPC Power Converter Principle Analysis and Modeling

This chapter provides a detailed analysis of operating principles for the 3L-NPC power converter and then establishes the mathematical model in three-phase stationary coordinates. Utilizing the Park-Clarke transformation, the mathematical model of the converter is further developed in two-phase stationary synchronous and two-phase rotating reference frames. Furthermore, this chapter outlines several PWM strategies.

2.1 3L-NPC Power Converter Circuit Analysis

The topology of 3L-NPC power converter is depicted in Figure 2.1. Assumptions for simplified

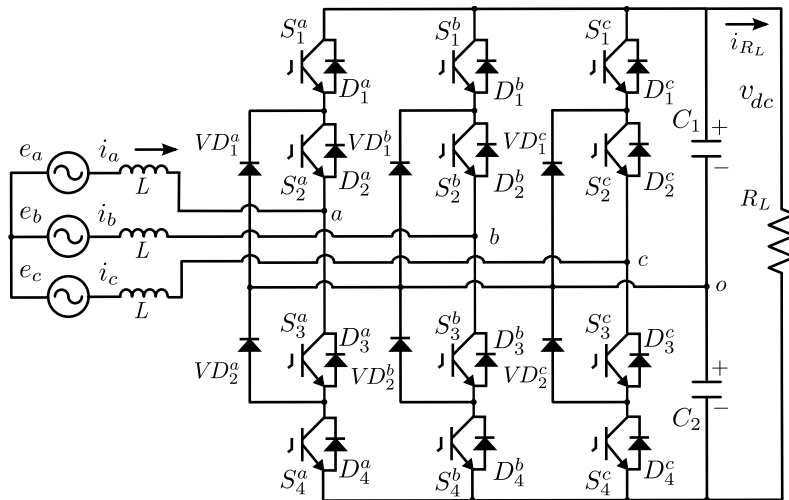


Figure 2.1: The topology of 3L-NPC power converter.

analysis:

- The power devices IGBTs constituting the midpoint-clamping NPC circuit possess ideal characteristics: no inertia, no internal losses, instantaneous switching capability, and ideal diode behavior without considering reverse recovery processes.
- The parallel capacitors C_1 and C_2 on the dc side have sufficiently large values. The dc-voltage v_{dc} is ripple-free and unaffected by load variations.
- The converter circuit has ideal loads, inductors, and capacitors, all considered ideal components.

The ac side inductance, denoted as L , filters out high-frequency harmonics. On the dc side, capacitors C_1 and C_2 help maintain the bus voltage and perform filtering functions. v_{dc} represents the dc bus voltage. Each converter phase consists of four switching tubes. Considering the arrangement and combination of switching states, each phase theoretically has nine combinations.

2.2 Switching State Analysis

Given the symmetry across the three phases of the 3L-NPC power converter, a -phase is exemplified to illustrate the switching states. This analysis applies equally to b -phase and c -phase. Figure 2.2 shows the topological structure of a -phase in a 3L-NPC power converter, in which a detailed analysis of the three switching states of a phase is provided. "1" indicates the on state, and "0" indicates the off state for switching tubes S_1^a , S_2^a , S_3^a , and S_4^a .

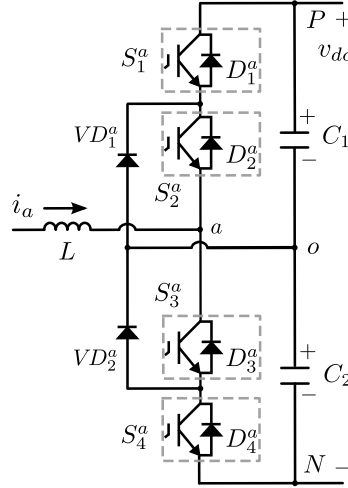


Figure 2.2: The topological structure of a -phase in 3L-NPC power converter.

In practice, switching tubes S_1^a and S_2^a , as well as S_3^a and S_4^a , are complementary. Concurrently, it must not be activated simultaneously to avert a short circuit. For the a -bridge arm, when the switches are in state P , it indicates that the upper two IGBTs on the bridge arm are conducting. The voltage v_{ao} at terminal a relative to the midpoint o of the converter is $\frac{v_{dc}}{2}$. Similarly, when the switches are in state n , it indicates that the lower two IGBTs are conducting, and at this point,

Table 2.1: Switch State of a -phase

Switch State	IGBT Swithc State				Voltage v_{ao}
	S_1^a	S_2^a	S_3^a	S_4^a	
P	1	1	0	0	$\frac{v_{dc}}{2}$
O	0	1	1	0	0
N	0	0	1	1	$-\frac{v_{dc}}{2}$

the voltage v_{ao} at terminal a relative to point o is $-\frac{v_{dc}}{2}$. Given these constraints, each phase effectively has three operational switching states, corresponding to distinct dc-voltage outputs: $\frac{v_{dc}}{2}$, 0, $-\frac{v_{dc}}{2}$, as shown in Table 2.1.

2.3 Commutation Process Analysis

Three operational states of the bridge arm are defined by the status of the four switches within the single-phase bridge arm. The current mode is analyzed below, taking into account the transition between these operational states. Current modes are divided into positive and negative, with current flowing from ac to dc being positive, and the reverse being negative. This results in three stable operational states and two dead zones during forward and reverse current flow, as shown in Figures 2.3 (a-e).

Conduct a modal analysis of the circuit under positive current $i_a > 0$ direction:

1. When the a -bridge arm operates in 1100, with the current direction $i_a > 0$ positive, current is conducted through the freewheeling diodes D_1^a and D_2^a . The current path is as follows: inductor $L \rightarrow$ freewheeling diode $D_2^a \rightarrow$ freewheeling diode $D_1^a \rightarrow$ capacitor $C_1 \rightarrow$ midpoint o of the capacitor. The current does not pass through the switching tubes S_1^a and S_2^a , as depicted in Figure 2.3(a).
2. Before Converting the working state of a -bridge arm from 1100 to 0110, a dead zone state of 0100 exists. Switching tube S_1^a is in the off state. Since no current flows through switching tube S_1^a in the previous state, switching its state does not affect circuit operation. At this moment, there is no change in the working state within the circuit, and the current flow direction remains unchanged: inductor $L \rightarrow$ freewheeling diode $D_2^a \rightarrow$ freewheeling diode $D_1^a \rightarrow$ capacitor $C_1 \rightarrow$ midpoint o of the capacitor. Diodes D_1^a and D_2^a serve the purpose of freewheeling, as shown in Figure 2.3(b).
3. The working state of the a -bridge arm transitions from 0100 to 0110, during which a commutation behavior occurs in the circuit. The current flow direction changes to inductor $L \rightarrow$ switching tube $S_3^a \rightarrow$ clamping diode $VD_2^a \rightarrow$ midpoint o of capacitor. During this process, the clamping diode VD_1^a is forcefully turned off. Although switching tube S_2^a remains open, no current flows through it, as depicted in Figure 2.3(c).
4. Before Converting the working state of the a -bridge arm from 0110 to 0011, a dead zone state of 0010 exists. At this point, switching tube S_2^a transitions from the on-state to the

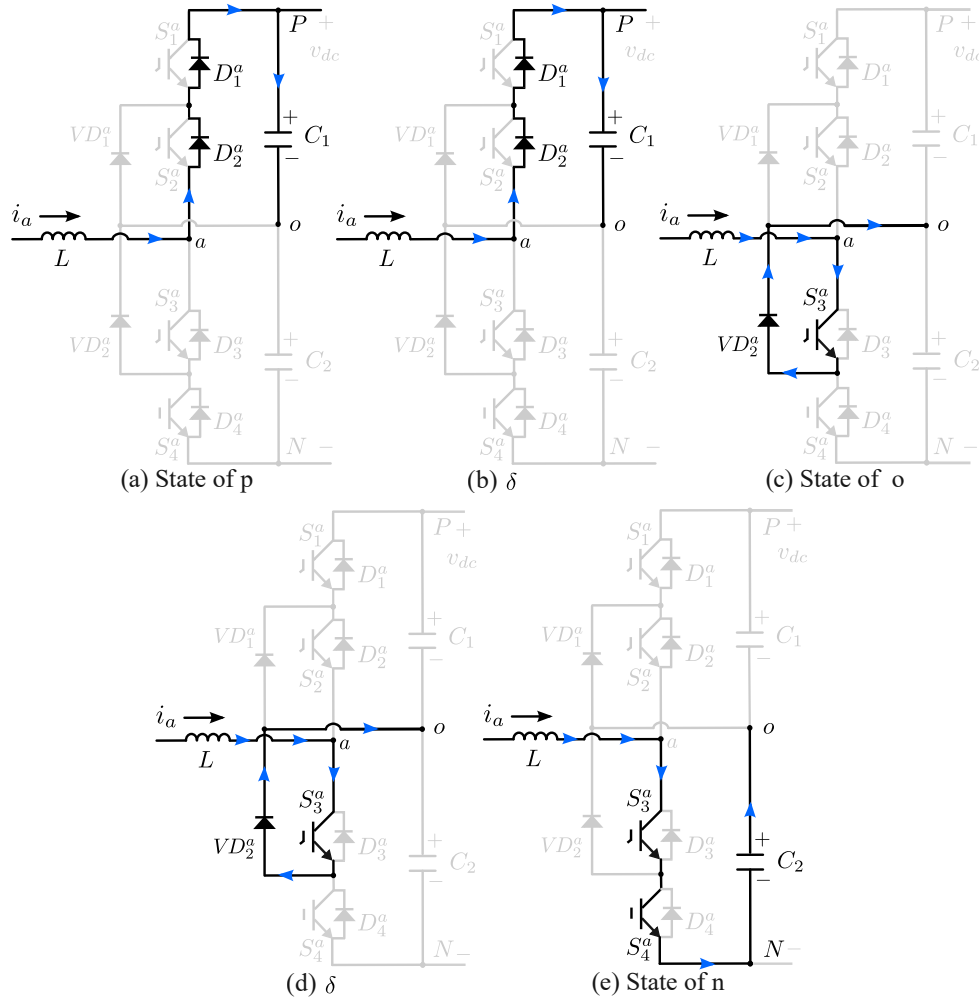


Figure 2.3: When $i_a > 0$, the process of a -phase current change.

off-state. Since no current flows through switching tube S_2^a in the previous state, switching its state does not affect circuit operation. The current flow direction remains unchanged: inductor $L \rightarrow$ switching tube $S_3^a \rightarrow$ clamping diode $VD_2^a \rightarrow$ midpoint o of capacitor, as depicted in Figure 2.3(d).

5. Converting the working state of a -bridge arm from 0010 to 0011, during which a commutation event occurs in the circuit. The current path is as follows: inductor $L \rightarrow$ switching tube $S_3^a \rightarrow$ switching tube $S_4^a \rightarrow$ capacitor $C_2 \rightarrow$ midpoint o of the capacitor. During this process, the clamp diode VD_2^a is forcefully turned off. At this time, neither the freewheeling diode nor the clamp diode has current flowing through it, as shown in Figure 2.3(e).

Analysis of circuit modes when the current direction is negative ($i_a < 0$) and results in three stable operational states and two dead zones, as shown in Figures 2.4 (a-e).

1. When the a -bridge arm exhibits a state of 1100, the current is negative ($i_a < 0$), flowing through the switching tubes S_1^a and S_2^a , with the current direction as circuit terminal $P \rightarrow$

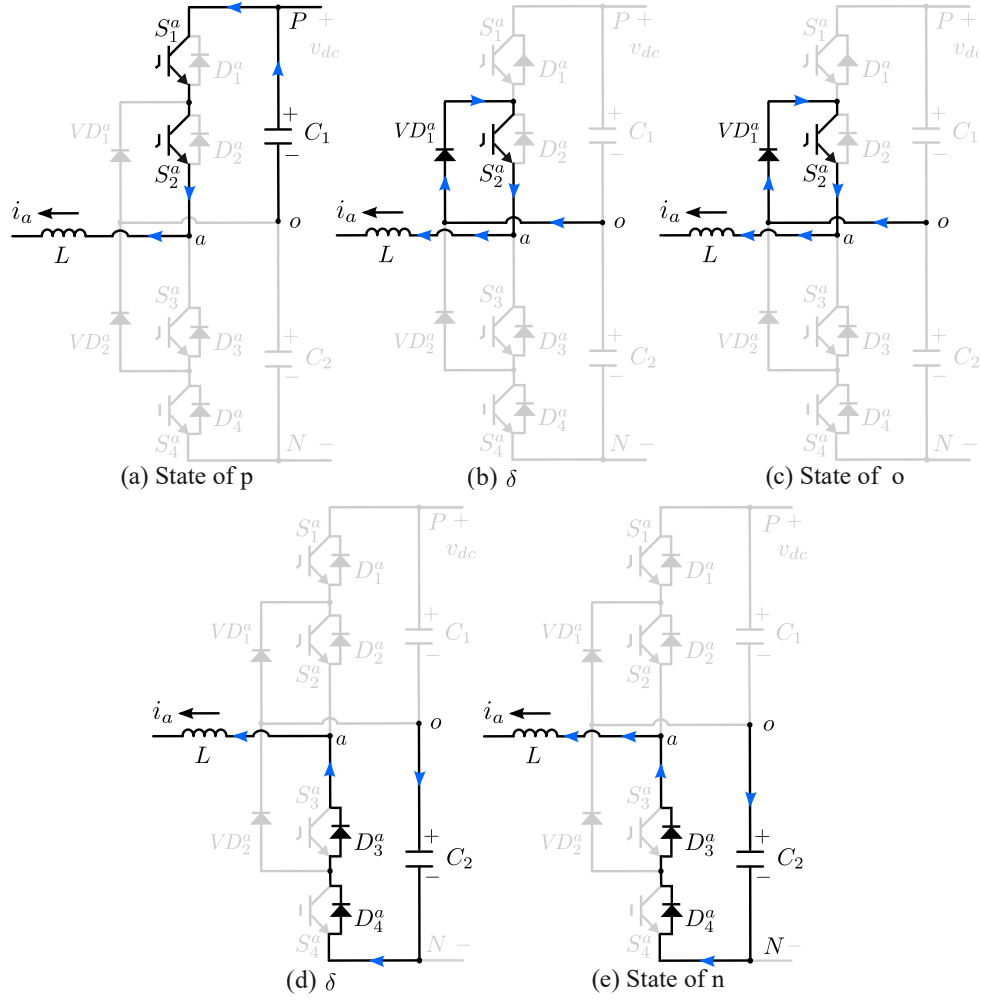


Figure 2.4: When $i_a < 0$, the process of a -phase current change.

switching tube $S_1^a \rightarrow$ switching tube $S_2^a \rightarrow$ inductor L . Under this working state, neither the freewheeling diode nor the clamp diode carries any current, as depicted in Figure 2.4(a).

2. Before the transition of the a -bridge arm working state from 1100 to 0110, there exists a dead zone state of 0100. At this point, switching tube S_2^a remains in an on-state while switching tube S_1^a transitions from an on-state to an off-state, resulting in commutation behavior in the circuit. At this time, the current flows through switching tube S_2^a and clamp diode VD_1^a . The current direction becomes midpoint of the capacitor $o \rightarrow$ clamp diode $VD_1^a \rightarrow$ switching tube $S_2^a \rightarrow$ inductor L , as illustrated in Figure 2.4(b).
3. During the transition of the a -bridge arm working state from 0100 to 0110. In both the 0100 state and the 0110 state, the current does not flow through the switching tube S_3^a . At this time, the state change of the switching tube S_1^a does not affect the circuit operation. The current direction remains unchanged, continuing as midpoint o of the capacitor $\rightarrow$ clamp diode $VD_1^a \rightarrow$ switching tube $S_2^a \rightarrow$ inductor L , as shown in Figure 2.4(c).

4. Before the transition of the a -bridge arm working state from 0110 to 0011, there also exists a dead zone state of 0010. At this point, switching tube S_2^a transitions from an on-state to an off-state, while switching tube S_3^a remains in an on-state, causing commutation behavior in the circuit. The current direction changes to midpoint of the capacitor $o \rightarrow$ capacitor $C_2 \rightarrow$ freewheeling diode $D_4^a \rightarrow$ freewheeling diode $D_3^a \rightarrow$ inductor L . During this process, clamp diode VD_2^a is forcibly turned off, and switching tube S_3^a does not carry any current, as depicted in Figure 2.4(d).
5. During the transition of the a -bridge arm working state from 0010 to 0011, neither switching tube S_3^a nor S_4^a carries current in the 0010 states nor the 0011 states. The state change of switching tube S_4^a at this point does not affect the circuit operation. The current direction remains unchanged, continuing as midpoint o of the capacitor $\rightarrow$ capacitor $C_2 \rightarrow$ freewheeling diode $D_4^a \rightarrow$ freewheeling diode $D_3^a \rightarrow$ inductor L . At this time, neither the switching tubes nor the clamp diodes carry any current, as shown in Figure 2.4(e).

2.4 3L-NPC Power Converter Modeling

Establishing the mathematical model of the 3L-NPC power converter is fundamental for studying its control methods. Before deriving the model of the 3L-NPC converter, the following assumptions are made to simplify the derivation process [103, 104]:

1. In the 3L-NPC power converter, the switches and diodes are ideal devices in all three bridge arms, neglecting the switching delay and forward conduction resistance of switches and diodes.
2. The ac-side inductance of the 3L-NPC power converter is assumed to be ideal, disregarding leakage inductance.
3. Both capacitance values on the dc-side of the 3L-NPC power converter are assumed to be identical.
4. The switching frequency of the switches in the 3L-NPC power converter is significantly higher than the fundamental frequency of the grid.
5. Equivalent series resistance, circuit impedance, and other losses in the 3L-NPC power converter are represented using equivalent resistances, neglecting stray inductances presented by components such as connecting wires and switches in the circuit.

2.4.1 Stationary Coordinate Model (abc -axis)

The equivalent circuit featuring ideal switches is presented in Figure 2.5.

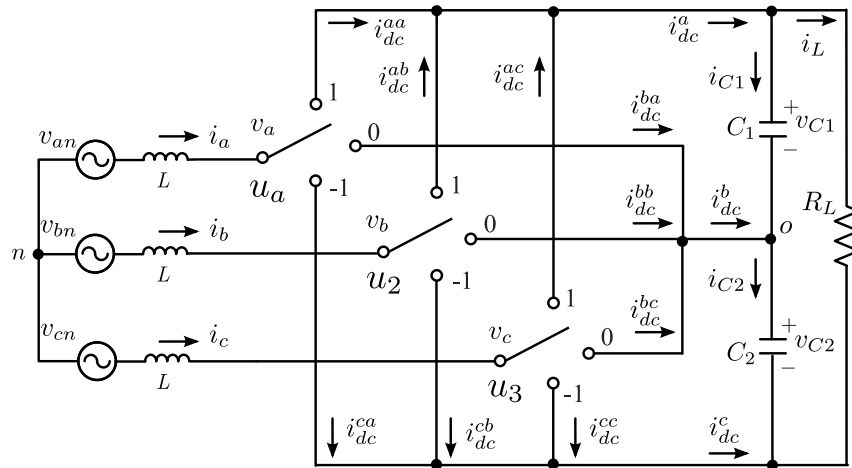


Figure 2.5: A 3L-NPC power converter with ideal switches.

2.4.1.1 Inductance Current Dynamics

Based on the Kirchhoff's Voltage Law (KVL) yields

$$\begin{cases} v_{an} = L \frac{di_a}{dt} + v_a + v_{on}, \\ v_{bn} = L \frac{di_b}{dt} + v_b + v_{on}, \\ v_{cn} = L \frac{di_c}{dt} + v_c + v_{on}, \end{cases} \quad (2.1)$$

where

v_{an}, v_{bn}, v_{cn}	grid voltage referred to "n";
$i_a, i_b, i_c,$	grid current frame;
$v_a, v_b, v_c,$	injected voltages referred to "o";
v_{on}	voltage at "o" referred to "n";
L	Filter inductor;
C	dc-link capacitor;
R	dc-load resistance;
$i_{dc}^a, i_{dc}^b, i_{dc}^c, i \in \{a, b, c\}$	current at top, middle and bottom connection points of capacitors;
i_{C1}, i_{C2}	capacitors currents;

Given that the source voltages are balanced and no fourth wire is utilized in this study, then the assumption can be obtained

$$v_{an} + v_{bn} + v_{cn} = 0, \quad (2.2)$$

$$i_a + i_b + i_c = 0. \quad (2.3)$$

Combining equations (2.1) yields

$$v_{on} = -\frac{1}{3}(v_a + v_b + v_c). \quad (2.4)$$

Substituting (2.4) into (2.1), one get

$$\begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \begin{bmatrix} L \frac{di_a}{dt} \\ L \frac{di_b}{dt} \\ L \frac{di_c}{dt} \end{bmatrix} + \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}. \quad (2.5)$$

The voltages v_i across switches, where $i \in \{a, b, c\}$, are defined concerning the reference point "o" and are determined by the switching positions u_i as follows:

$$v_i = \begin{cases} v_{C2}, u_i = 1, \\ 0, u_i = 0, \\ v_{C1}, u_i = -1, \end{cases} \quad (2.6)$$

According to equation (2.6), there are three permissible states, and the expression v_i (for every $i \in \{a, b, c\}$) of u_i is obtained. A single quadratic function can be formulated to suit all three states,

$$v_i = \frac{v_{C2} - v_{C1}}{2} u_i^2 + \frac{v_{C2} + v_{C1}}{2} u_i, \quad (2.7)$$

where $u_i \in \{-1, 0, 1\}$, $u_i^2 \in \{0, 1\}$.

In conclusion the inductors currents dynamics can be modelled as (2.5) where the voltages v_i are computed as in (2.7).

Remark 2.4.1 Under the consideration that $v_{C1} \cong v_{C2}$, (2.7) reduces to

$$v_i = \frac{v_{C2} + v_{C1}}{2} u_i, \quad i \in (1, 2, 3) \quad (2.8)$$

Thus, the subsystem (2.5) can be approximated as

$$\begin{bmatrix} L \frac{di_a}{dt} \\ L \frac{di_b}{dt} \\ L \frac{di_c}{dt} \end{bmatrix} = -\frac{v_{C1} + v_{C2}}{6} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} + \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} \quad (2.9)$$

2.4.1.2 Capacitance Voltage Dynamics

Using Kirchhoff's Current Law (KCL) in the diagram of Figure 2.5, we obtain the following relationships

$$\begin{aligned} i_{dc}^a &= i_{C2} + i_R, \\ i_{dc}^b &= i_{C1} - i_{C2}, \\ i_{dc}^c &= -i_{C1} - i_R, \\ 0 &= i_{dc}^a + i_{dc}^b + i_{dc}^c, \end{aligned}$$

where $i_{C1} = C\dot{v}_{C1}$, $i_{C2} = C\dot{v}_{C2}$, $i_R = \frac{v_{C1}+v_{C2}}{R}$, and $i_{dc}^i = i_{dc}^{ia} + i_{dc}^{ib} + i_{dc}^{ic}$, for every $i \in \{a, b, c\}$.

From these expressions, it can be obtained as,

$$i_{C2} + i_{C1} = i_{dc}^a - i_{dc}^c + 2i_R, \quad (2.10)$$

$$i_{C1} - i_{C2} = i_{dc}^b. \quad (2.11)$$

Expressions for i_{dc}^a , i_{dc}^b , and i_{dc}^c are approximated as quadratic functions following a similar procedure as before.

1) Current i_{dc}^a : We first establish the relationship between each output current $i_{dc}^{a,k}$ in terms of the three possible switch positions u_k , and for every $k \in \{a, b, c\}$. Thus, for a given $k \in \{a, b, c\}$ three points are known, and a quadratic function relating $i_{dc}^{a,k}$ and u_k can be designed to fit all three points.

The expression for the $k - th$ quadratic function is

$$i_{dc}^{a,k} = (u_k + 1) \frac{u_k i_k}{2}$$

Now, since $i_{dc}^a = \sum_{k \in \{a,b,c\}} i_{dc}^{a,k}$, we obtain the following expression

$$i_{dc}^a = (u_a + 1) \frac{u_a i_a}{2} + (u_b + 1) \frac{u_b i_b}{2} + (u_c + 1) \frac{u_c i_c}{2} \quad (2.12)$$

2) Current i_{dc}^b : In this case the expression for the quadratic function relating current $i_{dc}^{b,k}$ and u_k is

$$i_{dc}^{b,k} = (1 - u_k^2) i_k$$

From $i_{dc}^b = \sum_{k \in \{a,b,c\}} i_{dc}^{b,k}$, we obtain

$$\begin{aligned} i_{dc}^b &= (1 - u_a^2) i_a + (1 - u_b^2) i_b + (1 - u_c^2) i_c \\ &= -u_a^2 i_a - u_b^2 i_b - u_c^2 i_c \end{aligned} \quad (2.13)$$

where we used the fact that $i_a + i_b + i_c = 0$

3) Current i_{dc}^c : The case if i_{dc}^c is very similar to i_{dc}^a . In this case the expression of the quadratic function relating current $i_{dc}^{c,k}$ and u_k is

$$i_{dc}^{c,k} = (u_k - 1) \frac{u_k i_k}{2} \quad (2.14)$$

As before, from $i_{dc}^c = \sum_{k \in \{a,b,c\}} i_{dc}^{c,k}$, we obtain

$$i_{dc}^c = (u_a - 1) \frac{u_a i_a}{2} + (u_b - 1) \frac{u_b i_b}{2} + (u_c - 1) \frac{u_c i_c}{2} \quad (2.15)$$

Direct substitution of (2.12), (2.13) and (2.15) in (2.10) and (2.11) yields the following expressions

$$\begin{aligned} C(\dot{v}_{C1} + \dot{v}_{C2}) &= i_{dc}^a - i_{dc}^c - 2i_R, \\ &= u_a i_a + u_b i_b + u_c i_c - \frac{2(v_{C1} + v_{C2})}{R}, \end{aligned} \quad (2.16)$$

$$\begin{aligned} C(\dot{v}_{C1} - \dot{v}_{C2}) &= i_{dc}^b \\ &= -u_a^2 i_a - u_b^2 i_b - u_c^2 i_c \end{aligned} \quad (2.17)$$

Summarizing, the overall model of system is given by

$$\begin{bmatrix} L \frac{di_a}{dt} \\ L \frac{di_b}{dt} \\ L \frac{di_c}{dt} \end{bmatrix} = -\frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} + \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} \quad (2.18)$$

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \frac{v_{C1} - v_{C2}}{2} \begin{bmatrix} u_a^2 \\ u_b^2 \\ u_c^2 \end{bmatrix} + \frac{v_{C2} + v_{C1}}{2} \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} \quad (2.19)$$

$$C \frac{d}{dt}(v_{C1} + v_{C2}) = \begin{bmatrix} u_a & u_b & u_c \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \frac{2(v_{C1} + v_{C2})}{R} \quad (2.20)$$

$$C \frac{d}{dt}(v_{C1} - v_{C2}) = - \begin{bmatrix} u_a^2 & u_b^2 & u_c^2 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.21)$$

where define $v_{dc} = v_{C1} + v_{C2}$, $v_{dc}^o = v_{C1} - v_{C2}$, then (2.18)-(2.21) can be rewritten as,

$$L \dot{\mathbf{i}}_{abc} = -\frac{1}{6} \begin{pmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{pmatrix} \begin{pmatrix} u_a & u_a^2 \\ u_b & u_b^2 \\ u_c & u_c^2 \end{pmatrix} \begin{pmatrix} v_{dc} \\ v_{dc}^o \end{pmatrix} + \mathbf{v}_{abc} \quad (2.22)$$

$$\dot{v}_{dc} = \frac{1}{C} \mathbf{u}_{abc}^T \mathbf{i}_{abc} - \frac{2}{C} \frac{v_{dc}}{R} \quad (2.23)$$

$$\dot{v}_{dc}^o = -\frac{1}{C} (\mathbf{u}_{abc}^2)^T \mathbf{i}_{abc} \quad (2.24)$$

where $\mathbf{v}_{abc} = \{v_{an}, v_{bn}, v_{cn}\}^T$, $\mathbf{i}_{abc} = \{i_a, i_b, i_c\}^T$, $\mathbf{u}_{abc} = \{u_a, u_b, u_c\}^T$ is the three-phase duty cycle output by the controller.

2.4.2 Stationary Coordinate Model ($\alpha\beta\gamma$ -axis)

Vectors originally in abc reference frame are transformed into $\alpha\beta\gamma$ reference frame using the power-invariant transformation defined by the transformation matrix in

$$\{\cdot\}_{\alpha\beta\gamma} = T_{abc-\alpha\beta\gamma} \{\cdot\}_{abc}, \quad T_{abc-\alpha\beta\gamma} = \sqrt{\frac{2}{3}} \begin{pmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{pmatrix} \quad (2.25)$$

Applying this transformation to (2.22)-(2.24) yields the system dynamics in the $\alpha\beta\gamma$ reference frame

$$\dot{\mathbf{i}}_{\alpha\beta} = \frac{\mathbf{v}_{\alpha\beta}}{L} - \frac{1}{2L}v_{dc}\mathbf{u}_{\alpha\beta} + \frac{\mathbf{Y}}{L}, \quad (2.26)$$

$$\dot{v}_{dc} = \frac{\mathbf{u}_{\alpha\beta}^T \dot{\mathbf{i}}_{\alpha\beta}}{C} - \frac{2v_{dc}}{CR_L}, \quad (2.27)$$

$$\dot{v}_{dc}^o = \frac{2}{\sqrt{6}C}\mathbf{Z} + \frac{1}{\sqrt{6}C}\mathbf{u}_{\alpha\beta}^T \dot{\mathbf{i}}_{\alpha\beta} u_\gamma, \quad (2.28)$$

where $\mathbf{Y} = \begin{bmatrix} \frac{1}{2\sqrt{6}}(u_\beta^2 - u_\alpha^2) - \frac{1}{\sqrt{3}}u_\alpha u_\gamma \\ \frac{u_\alpha u_\beta}{\sqrt{6}} - \frac{u_\beta u_\gamma}{\sqrt{3}} \end{bmatrix} v_{dc}^o$, $\mathbf{Z} = \begin{bmatrix} u_\alpha^2 - u_\beta^2 & -u_\alpha u_\beta \end{bmatrix} \mathbf{i}_{\alpha\beta}$, $\mathbf{v}_{\alpha\beta} = [v_\alpha, v_\beta]^T$ grid voltage vector, $\mathbf{i}_{\alpha\beta} = [i_\alpha, i_\beta]^T$ is grid current vector, $\mathbf{u}_{\alpha\beta} = [u_\alpha, u_\beta]^T$ and u_γ are control signals.

After generating the control signals, they are transformed back to u_a, u_b, u_c and executed through a modulator. It can be observed that, due to the requirement for γ -axis control in capacitor voltage balancing, the fixed coordinate transformation here is a three-axis coordinate transformation.

It is important to note that, when the converter usually operates, v_{dc}^o will be close to zero, which can be assumed to be at least an order of magnitude lower than v_{dc} . On the other hand, the duty cycle $\mathbf{u}_{\alpha\beta}$ takes values in the interval $[-1, 1]$, so the third term on the right side of equation (2.26) will be at least two orders of magnitude lower than the second term.

Based on this consideration, equations (2.26) can be approximated as follows:

$$\dot{\mathbf{i}}_{\alpha\beta} = -\frac{\mathbf{v}_{\alpha\beta}}{L} - \frac{v_{dc}}{2L}\mathbf{u}_{\alpha\beta} \quad (2.29)$$

The following simplifies (2.28) from the perspective of instantaneous power. Based on the instantaneous active power p and reactive power q as follows,

$$p = \mathbf{v}_{\alpha\beta}^T \dot{\mathbf{i}}_{\alpha\beta}, q = \mathbf{v}_{\alpha\beta}^T \mathbf{J} \dot{\mathbf{i}}_{\alpha\beta}, \mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}. \quad (2.30)$$

Based on (2.29), the the dynamics of the (2.30) is expressed as,

$$L\dot{p} = \mathbf{v}_{\alpha\beta}^T (L\omega q \frac{\mathbf{v}_{\alpha\beta}}{\|\mathbf{v}_{\alpha\beta}\|^2} + \mathbf{v}_{\alpha\beta} - \frac{v_{dc}}{2}\mathbf{u}_{\alpha\beta}) \quad (2.31)$$

$$L\dot{q} = \mathbf{v}_{\alpha\beta}^T \mathbf{J}^T (-L\omega p \frac{\mathbf{J}\mathbf{v}_{\alpha\beta}}{\|\mathbf{v}_{\alpha\beta}\|^2} - \frac{v_{dc}}{2}\mathbf{u}_{\alpha\beta}) \quad (2.32)$$

Let $\dot{p} = 0, \dot{q} = 0$, the equivalent controller $\mathbf{u}_{\alpha\beta}^{eq}$ can be obtained,

$$\mathbf{u}_{\alpha\beta}^{eq} = \frac{2}{v_{dc}} \left\{ (\|\mathbf{v}_{\alpha\beta}\|^2 + \omega L q) \frac{\mathbf{v}_{\alpha\beta}}{\|\mathbf{v}_{\alpha\beta}\|^2} - (\omega L p) \frac{\mathbf{J}\mathbf{v}_{\alpha\beta}}{\|\mathbf{v}_{\alpha\beta}\|^2} \right\} \quad (2.33)$$

Consider the instantaneous power dynamics are faster than that of v_{dc}^o , let $p = p^*, q = q^*$, and substitute (2.33) into to (2.28), we can get:

$$\dot{v}_{dc}^o = \frac{2p^*}{\sqrt{6}Cv_{dc}}u_\gamma + \psi(t) \quad (2.34)$$

where

$$\psi(t) = \frac{2}{\sqrt{6}C}[(u_{\alpha}^{eq})^2 - (u_{\beta}^{eq})^2, -2u_{\alpha}^{eq}u_{\beta}^{eq}]\mathbf{i}_{\alpha\beta} \quad (2.35)$$

where $\psi(t)$ is regarded as the fundamental wave and the third harmonic.

Therefore, based on (2.27), (2.29), and (2.34), the simplified dynamic model of the 3L-NPC converter can be obtained as,

$$\dot{\mathbf{i}}_{\alpha\beta} = \frac{1}{L}\mathbf{v}_{\alpha\beta} - \frac{v_{dc}}{2L}\mathbf{u}_{\alpha\beta} \quad (2.36)$$

$$\dot{v}_{dc} = \frac{1}{2C}\mathbf{u}_{\alpha\beta}^T \dot{\mathbf{i}}_{\alpha\beta} - \frac{2}{CR_L}v_{dc} \quad (2.37)$$

$$\dot{v}_{dc}^o = \frac{2p^*}{\sqrt{6}v_{dc}}u_{\gamma} + \psi(t) \quad (2.38)$$

2.4.3 Rotating Coordinate Model ($dq0$ -axis)

Apply the Park transformation to the simplified $\alpha\beta$ -axis 3L-NPC converter dynamic model (2.36)-(2.38),

$$\{\cdot\}_{dq0} = T_{\alpha\beta\gamma-dq0}\{\cdot\}_{abc}, \quad T_{\alpha\beta\gamma-dq0} = \begin{bmatrix} \cos \omega t & \sin \omega t & 0 \\ -\sin \omega t & \cos \omega t & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad (2.39)$$

Modeling the system in the synchronous rotation coordinate system ($d, q, 0$) is possible as follows:

$$\dot{i}_d = \omega i_q + \frac{1}{L}v_d - \frac{v_{dc}}{2L}u_d \quad (2.40)$$

$$\dot{i}_q = -\omega i_d + \frac{1}{L}v_q - \frac{v_{dc}}{2L}u_q \quad (2.41)$$

$$\dot{v}_{dc} = \frac{1}{C}u_d i_d + \frac{1}{C}u_q i_q - \frac{2}{C}i_L \quad (2.42)$$

$$\dot{v}_{dc}^o = \frac{2p^*}{\sqrt{6}Cv_{dc}}u_0 + \psi(t) \quad (2.43)$$

where v_d and v_q are the grid voltage, i_d and i_q are the grid current, ω is the grid voltage angular frequency, u_d , u_q and u_0 are the control signals, $\psi(t)$ is the signal composed of the fundamental wave and the third-order harmonic, p^* is the active power reference value.

2.5 PWM Technology of 3L-NPC Power Converter

Pulse width modulation (PWM) technology uses the impulse area equivalence principle to convert ac signals into a sequence of pulse signals over time, allowing for adjustments in frequency and amplitude and the suppression of specific harmonics [105–107]. Multi-level PWM typically builds upon the foundation of two-level modulation techniques.

In 1977, Holtz introduced the three-level converter topology, serving as the prototype for the T-type three-level converter. By 1980, at the IEEE IAS annual conference, Japanese scholars had advanced the NPC-type three-level topology. Then several PWM strategies were also proposed, including carrier-based sinusoidal pulse width modulation (SPWM) [108,109], space vector pulse width modulation (SVPWM) [110,111], virtual space vector modulation (VSVPWM) [112,113], selective harmonic elimination pulse width modulation (SHEPWM) [114,115], and model predictive control, which directly generates PWM signals without carrier modulation, are among the PWM strategies identified [116,117]. Each method has advantages and disadvantages, requiring strategic selection for specific applications.

The three-level SPWM method evolves from the conventional two-level SPWM approach. The three-level SPWM method is an evolution of the traditional two-level SPWM approach. Each arm of the three-level NPC converter consists of four switching devices, requiring two sets of triangular carrier waves with identical amplitude and frequency to produce driving signals for the four switching tubes on the arm. When positioned above and below the time axis, these two sets of triangular carrier waves yield the driving pulse signals for the switching tubes when compared with the modulating wave signal. Typically, SPWM is categorized into two types: (1) In-phase carrier type, employing two sets of triangular carriers with identical phases for the upper and lower waves; (2) Phase-reversed carrier type, with a 180-degree phase difference between the upper and lower triangular carriers. The three-level SPWM strategy offers convenience and simplicity. It suits multi-level converters and structures like the diode-clamped, flying-capacitor-clamped, and cascaded multi-level types.

As PWM is not the central theme of this research, further details will not be elaborated upon.

2.6 Summary

This chapter provides an in-depth analysis of the operating principle of the 3L-NPC power converter, a mathematical model within the static three-phase coordinate system is established. The two-phase static synchronous and rotating reference frame models are obtained by Park-Clarke transformation. Moreover, a brief overview of the converter PWM technology is provided. The above foundation knowledge offers significant theoretical support and practical guidance for the control and application of the 3L-NPC power converter.

CHAPTER 3

Quasi-sliding Mode Power-Voltage Control for 3L-NPC Power Converters

High penetration of power converters may lead to power ripple, voltage swings, and weak anti-disturbances for microgrids. Figure 3.1 shows an overview of power converters in microgrids. Confronting these issues, this chapter proposes a robust control scheme, DSTO-embedded QISM (DSTO-QISM), for a 3L-NPC power converter system, dramatically enhancing power/voltage regulation performance and anti-disturbance capability. Attractive features: A fast convergence DSTO is deployed to offset multi-disturbances caused by parameter mismatches, unknown loads, current path changes, switch mode noise, self-compensating power/voltage tracking biases in QISM. To further mitigate power/voltage steady-state error and boost system robustness, a new quasi-integral sliding mode surface is built, inherently improving power/voltage tracking performance. Experimental data confirm the proposed control outperforms the discrete-time extended state observer-QISM (DESO-QISM), DSTO-quasi-sliding mode control (DSTO-QSM), and discrete-time proportional integral (DPI) control in power/voltage, grid current harmonics, and robustness.

3.1 Introduction

SMC is an efficient solution for complex power converter systems due to its fast dynamic response, insensitivity to model parameters, disturbance rejection, and design simplicity [56]. The SMC method uses sliding surface function based only on the input current error for single-ended primary-inductor converters in [118]. Nevertheless, the method suffers from chattering in the control input, which is not desirable control in practice. To circumvent it, [119] adopts an SMC approach by a boundary layer used for optimizing the sliding surface trajectory. As a result, the boundary layer-based SMC approach results in a trade-off between control performance and the elimination of chattering. A high-order sliding mode differentiator utilizes the high-order derivative of the system state as the control quantity, compensating for the harmonic currents

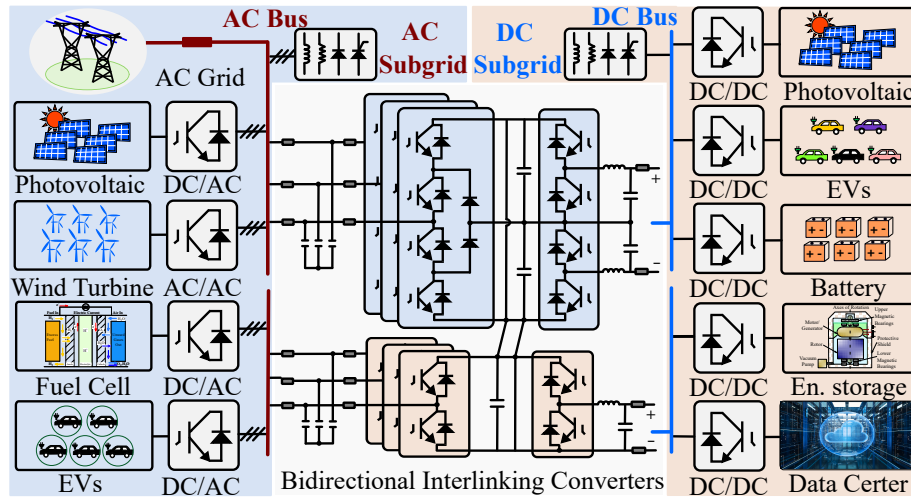


Figure 3.1: An overview of the power converters in microgrids.

in [120]. However, a high-order derivative of sliding variables inevitably affects the control method, which amplifies high-frequency components, resulting in undesirable current chattering. Besides, the control above schemes are designed to operate in a continuous-time domain.

Practically, digital signal processing achieves maximum reproducibility at the lowest possible cost for power conversion. Digital control relies on discrete-time systems. The discrete-time SMC systems are divided into discrete-time SMC (DSMC) [121] and discretized continuous-time SMC (CSMC) [122]. The discretized CSMC systems exhibit some intrinsic dynamic properties, such as periodic orbits and delta modulation effects, which may cause degradation of control performance and even system instability. Moreover, the stability criterion of CSMC is only a necessary condition for discrete-time stability, which explains why it cannot be converted from CSMC by simple equivalence [123]. A quasi-sliding mode (QSM) can be used for discrete-time sliding mode control systems, as it allows the system to confine the zigzag mode within a small band once it reaches the sliding mode [50]. QSMC is a sufficient and necessary condition for proving discrete-time system stability. Consequently, developing a QSMC-based stabilization is attractive due to its inherent robustness and guaranteed discretization stability.

An improved DSMC law is proposed to compensate the speed and current of the permanent magnet synchronous motor (PMSM) system for mismatch disturbances by exploiting a new QSM in [124]. As discussed in [125], a finite-set QSM predictive control is proposed to improve the out-voltage quality by introducing a quasi-sliding-mode function into the cost function. Inspired by the above methods, a new quasi-integral sliding mode control (QISMC) has been developed, which achieves high-precision control by adding the integral term to the QSM surface. However, large disturbances may result in system instability or require higher control gains that lead to the chattering. For this reason, it is necessary to design an observer to estimate part of disturbances as a feedforward compensation in control input, to enhance the voltage/power performance further.

Theoretically, numerous observer-based compensation disturbance methods have been proposed, such as resonant extended state observers [126], second-order fault-tolerant extended Kalman filter estimation [127], and sliding-mode observers [128]. A model-free predictive current control based on ESO for PMSM drives is proposed to reduce the computational time and

improve control performance in [129]. The ESO is considered as a estimator used to estimate external disturbances and uncertainties of PMSM systems. To improve the robustness of state observation, a discrete-time sliding mode observer (DSMO)-based compensation method is more attractive because its main features are similar to SMC. A DSMO is proposed for equivalent capacitor voltage control in modular multilevel converters, aiming to reduce the communication complexity in applications with several series-connected submodules in [130]. However, it is based on the frequency domain stability analysis of the DSMO, which relies on precise mathematical models and is susceptible to external disturbances and parameter uncertainties. In [20], an ESMDO not merely provides accurate multi-disturbance estimation, but is also equivalent to a built-in low-pass filter without introducing phase delay, thus significantly improving capacitance-voltage regulation. A discrete-time super-twisting observer (DSTO) is built into the control scheme to reduce the chattering further and enhance the robustness of power converter.

Concisely, motivated by observations, this chapter proposes a new robust control, DSTO-embedded-QISMC, for the 3L-NPC power converter, aiming at dramatically enhancing power and voltage regulation performance and anti-disturbance capability. Contributions include:

- 1) A fast convergence DSTO is constructed to offset multi-disturbances caused by parameter mismatches, unknown loads, current path changes, and switch mode noise, while self-compensating power and voltage tracking bias in QISMC, thereby attaining chattering suppression. The reason for chattering suppression is that the switching gain is only required to be designed greater than the bound of the multi-disturbances estimation error rather than that of the multi-disturbances.
- 2) To further mitigate the power/voltage steady-state error and improve the power converter system robustness, a new quasi-integral sliding mode surface is constructed, which essentially improves the power/voltage tracking performance.
- 3) Rigorous stability analysis of the discrete-time dynamics of power and voltage closed-loop systems provides a theoretical basis for the power converter system, achieving rapid convergence, high accuracy, and strong robustness.

3.2 System Description and Modeling

A 3L-NPC power converter topology is shown in Figure 3.2A, which connects the ac-link power source with three filter inductors and the dc-link with two capacitors. Notice that the instantaneous power of ac-link filter is not considered, the ac-link terminal power p' is equal to the grid side power p . Generally, the source bus connects photovoltaics, wind turbines, electric vehicles, etc. For simplification, the dc-link is connected to unknown equivalent loads. The list of nomenclatures of the power converter is presented in Table 3.1.

The 3L-NPC power converter is based on the Euler discretization model built from [19],

$$\mathbf{i}_{\alpha\beta,k+1} = \frac{\tau_s}{L} \mathbf{v}_{\alpha\beta,k} - \frac{\tau_s}{2L} v_{dc,k} \mathbf{u}_{\alpha\beta,k} + \frac{\tau_s}{L} \mathbf{M} + \mathbf{i}_{\alpha\beta,k}, \quad (3.1)$$

$$v_{dc,k+1} = -\frac{2\tau_s}{R_L C} v_{dc,k} + \frac{\tau_s}{C} \mathbf{u}_{\alpha\beta,k}^T \mathbf{i}_{\alpha\beta,k} + v_{dc,k}, \quad (3.2)$$

$$v_{dc,k+1}^o = \frac{\tau_s}{\sqrt{6}C} \mathbf{u}_{\alpha\beta,k}^T \mathbf{i}_{\alpha\beta,k} u_{\gamma,k} + \frac{2\tau_s}{\sqrt{6}C} \mathbf{N} + v_{dc,k}^o, \quad (3.3)$$



Table 3.1: Nomenclatures

Symbol	Description
$\mathbf{v}_{abc}$	$[v_a, v_b, v_c]^T$, grid voltage vector in abc frame.
$\mathbf{i}_{abc}$	$[i_a, i_b, i_c]^T$, grid current vector in abc frame.
$\mathbf{v}_{\alpha\beta,k}$	$[v_{\alpha,k}, v_{\beta,k}]^T$, grid voltage vector in $\alpha\beta$ -axis.
$\mathbf{i}_{\alpha\beta,k}$	$[i_{\alpha,k}, i_{\beta,k}]^T$, grid current vector in $\alpha\beta$ -axis.
$\mathbf{u}_{\alpha\beta\gamma,k}$	$[u_{\alpha,k}, u_{\beta,k}, u_{\gamma,k}]^T$, average duty cycles in $\alpha\beta$ -axis.
$\mathbf{VT}_l^n$	$n = \{a, b, c\}, l = \{1, 2, 3, 4\}$, insulated-gate bipolar transistors.
$\mathbf{VD}_l^n$	$n = \{a, b, c\}, l = \{1, 2\}$, clamping diodes.
$v_{dc,k}$	$v_{dc,k}^h + v_{dc,k}^l$, the sum of dc-link capacitor voltages.
$v_{dc,k}^o$	$v_{dc,k}^h - v_{dc,k}^l$, the difference of dc-link capacitor voltages.
R_L, i_L	dc-load resistance, dc-link load current.
L, C, ω	filter inductor, dc-link capacitor, grid voltage frequency.

where $\mathbf{M} = \begin{bmatrix} \frac{1}{2\sqrt{6}}(u_{\beta,k}^2 - u_{\alpha,k}^2) - \frac{1}{\sqrt{3}}u_{\alpha,k}u_{\gamma,k} & v_{dc,k}^o \\ \frac{u_{\alpha,k}u_{\beta,k}}{\sqrt{6}} - \frac{u_{\beta,k}u_{\gamma,k}}{\sqrt{3}} & \end{bmatrix} \mathbf{v}_{dc,k}^o$, $\mathbf{N} = [(u_{\alpha,k}^2 - u_{\beta,k}^2) - u_{\alpha,k}u_{\beta,k}] \mathbf{i}_{\alpha\beta,k}$, τ_s is sampling period, and $u_{\alpha,k}, u_{\beta,k}, u_{\gamma,k}$ are control signals.

Remark 3.2.1 Notice that source voltages are assumed to be symmetrical and balanced of the system (3.1)-(3.3), $u_{\alpha,k}, u_{\beta,k} \in [-1, 1]$ so mentioned $\mathbf{M}$ of (3.1) is lower than the second term of (3.1). $v_{dc,k}^o$ of equation (3.3) is lower by more than $v_{dc,k}$ one order of magnitude, due to $v_{dc,k}^o$ being close to zero. Therefore, considering the aforementioned, (3.1) can be approximated as,

$$\mathbf{i}_{\alpha\beta,k+1} = \frac{\tau_s}{L} \mathbf{v}_{\alpha\beta,k} - \frac{\tau_s}{2L} v_{dc,k} \mathbf{u}_{\alpha\beta,k} + \mathbf{i}_{\alpha\beta,k}. \quad (3.4)$$

Based on the instantaneous power theory [131], the active power p_k and reactive power q_k are defined as

$$p_k = \mathbf{i}_{\alpha\beta,k}^T \mathbf{v}_{\alpha\beta,k}, q_k = \mathbf{i}_{\alpha\beta,k}^T \mathbf{J} \mathbf{v}_{\alpha\beta,k}, \quad (3.5)$$

where $\mathbf{J} = [0, -1; 1, 0]$. Submitting (3.4) into (3.5), ones obtain

$$\begin{aligned} p_{k+1} &= -\frac{\tau_s}{2L} \mathbf{v}_{\alpha\beta,k}^T \mathbf{u}_{\alpha\beta,k} v_{dc,k} + \frac{\tau_s}{L} \|\mathbf{v}_{\alpha\beta,k}\|^2 \\ &\quad + \omega q_k \tau_s + p_k, \\ q_{k+1} &= -\frac{\tau_s v_{dc,k}}{2L} \mathbf{v}_{\alpha\beta,k}^T \mathbf{J} \mathbf{u}_{\alpha\beta,k} - \omega p_k \tau_s + q_k, \end{aligned} \quad (3.6)$$

where $\|\mathbf{v}_{\alpha\beta,k}\|^2 = \|v_{\alpha,k}\|^2 + \|v_{\beta,k}\|^2$.

The equilibrium point of (3.6) from [132] is obtained with $p_{k+1} = p_k = 0$ and $q_{k+1} = q_k = 0$ simultaneously:

$$\mathbf{u}_{\alpha\beta,k}^{eq} = \frac{2}{v_{dc,k}} \left\{ \left(1 + \frac{\omega L q_k}{\|\mathbf{v}_{\alpha\beta,k}\|^2} \right) - \frac{\omega L p_k}{\|\mathbf{v}_{\alpha\beta,k}\|^2} \mathbf{J} \right\} \mathbf{v}_{\alpha\beta,k}, \quad (3.7)$$

where $\mathbf{u}_{\alpha\beta,k}^{eq}$ represents the equilibrium point of the system in a steady state, i.e., instantaneous active and reactive power demanded by the power converter remain constant. Using the established equilibrium point in (3.7), the control architecture guarantees the instantaneous active and reactive power convergence to [132].

$$\mathbf{u}_{\alpha\beta,k} = \mathbf{u}_{\alpha\beta,k}^{eq} - \mu(e_{p,k})\mathbf{v}_{\alpha\beta,k} - \mu(e_{q,k})\mathbf{J}\mathbf{v}_{\alpha\beta,k}, \quad (3.8)$$

where $\mu(e_{p,k})$ and $\mu(e_{q,k})$ express the active power and reactive power controller, respectively.

3.3 Control Strategy Design

A description of the design procedure of the proposed DSTO-based QISMIC controller for the 3L-NPC power converter is presented in this section, as shown in Figure 3.2B.

3.3.1 Control Objectives

Three objectives need to be satisfied for the 3L-NPC power converter system: power tracking, voltage regulation, and capacitor voltage balancing. To achieve these objectives, a robust DSTO-QISMIC control structure is employed:

- In the voltage regulation loop, regulate capacitor voltages $v_{dc,k}$ to desired $v_{dc,k}^*$. The output of this loop is an active power reference p^* that goes into the power tracking loop.
- In the power tracking loop, force the active and reactive power p_k, q_k to track their references p_k^* and q_k^* , in which p_k^* is obtained from the voltage regulation loop, and q_k^* is given to change the power factor of the system.
- In the voltage balancing loop, maintain the balance of capacitor voltages $v_{dc,k}^o \rightarrow 0$. The outputs of both this loop and power tracking loop are combined to receive the final average duty cycle, controlling the power converter.

3.3.2 Power Tracking Loop

This subsection proposes a DSTO-based-QISMIC controller to accurately control instantaneous power ($p_k \rightarrow p_k^*, q_k \rightarrow q_k^*$), offsetting multi-disturbances (inductance mismatch, bias caused of current path changes and switch mode noise), and suppressing chattering feeding into the system. The power tracking errors are considered as

$$e_{p,k} = p_k^* - p_k, e_{q,k} = q_k^* - q_k, \quad (3.9)$$

where p_k^*, q_k^* are reference of active power p_k and reactive power q_k , respectively. The sampling period τ_s is very small, i.e., $p_k^* = p_{k+1}^*, q_k^* = q_{k+1}^*$.

Combining (3.6) and (3.9), Ones express as

$$e_{p,k+1} = -\omega q_k \tau_s + \frac{\mathbf{v}_{\alpha\beta,k}^T v_{dc,k} \tau_s}{2L} \mathbf{u}_{\alpha\beta,k} - \frac{\|\mathbf{v}_{\alpha\beta,k}\|^2 \tau_s}{L} + e_{p,k}, \quad (3.10)$$

$$e_{q,k+1} = \omega p_k \tau_s + \frac{\mathbf{v}_{\alpha\beta,k}^T \mathbf{J}^T v_{dc,k} \tau_s}{2L} \mathbf{u}_{\alpha\beta,k} + e_{q,k}. \quad (3.11)$$

Considering the multi-disturbances of the 3L-NPC power converter system in the practical plant, (3.10)-(3.11) become as,

$$e_{p,k+1} = -\omega q_k \tau_s + \frac{\mathbf{v}_{\alpha\beta,k}^T v_{dc,k} \tau_s}{2L} \mathbf{u}_{\alpha\beta,k} - \frac{\|\mathbf{v}_{\alpha\beta,k}\|^2 \tau_s}{L} + g_{p,k} + e_{p,k}, \quad (3.12)$$

$$e_{q,k+1} = \omega p_k \tau_s + \frac{\mathbf{v}_{\alpha\beta,k}^T \mathbf{J}^T v_{dc,k} \tau_s}{2L} \mathbf{u}_{\alpha\beta,k} + g_{q,k} + e_{q,k}. \quad (3.13)$$

where $g_{p,k}$ and $g_{q,k}$ refer to disturbances including inductance mismatch, bias cause by current path changes and switch mode noise. The sampling period τ_s is covered by the $g_{p,k}$ and $g_{q,k}$.

Substituting (3.8) into (3.12) and (3.13), respectively, ones get,

$$e_{j,k+1} = -O_{j,k} \mu(e_{j,k}) + g_{j,k} + e_{j,k}, \quad (3.14)$$

where $j \in \{p, q\}$, $O_{j,k} = \frac{v_{dc,k} \|\mathbf{v}_{\alpha\beta,k}\|^2 \tau_s}{2L}$.

The estimation system (3.14) can be expressed as,

$$\hat{e}_{j,k+1} = -O_{j,k} \mu(e_{j,k}) + \hat{g}_{j,k} + \hat{e}_{j,k}, \quad (3.15)$$

where $\hat{g}_{j,k}$ is the estimate of the $g_{j,k}$.

To deal with the disturbances $\hat{g}_{j,k}$ of (3.15), a DSTO is developed that allows fast and robust power estimation.

3.3.2.1 Discrete-time State Observer(DSTO) Design

The defined DSTO error is $\tilde{e}_{j,k} = e_{j,k} - \hat{e}_{j,k}$, and one gets,

$$\tilde{e}_{j,k+1} = -\hat{g}_{j,k} + g_{j,k} + \tilde{e}_{j,k}, \quad (3.16)$$

obviously, if $\tilde{e}_{j,k+1} = \tilde{e}_{j,k} = 0$ is satisfied, such that $\hat{g}_{j,k}$ can estimate completely $g_{j,k}$. Then the DSTO can be designed as

$$\begin{cases} \hat{g}_{j,k} = \lambda_j |\tilde{e}_{j,k}|^{\frac{1}{2}} \text{sign}(\tilde{e}_{j,k}) + \Phi_{j,k}, \\ \Phi_{j,k+1} = \chi_j \text{sign}(\tilde{e}_{j,k}) \tau_s + \Phi_{j,k}, \end{cases} \quad (3.17)$$

where gains λ_j and χ_j are positive value. $\Phi_{j,k}$ aims at eliminating the DSTO steady-state error.

Further, to obtain indirect $\hat{g}_{j,k} \rightarrow g_{j,k}$, it is equivalent to proving the following equation,

$$\begin{cases} \tilde{e}_{j,k+1} = \lambda_j |\tilde{e}_{j,k}|^{\frac{1}{2}} \text{sign}(\tilde{e}_{j,k}) \tau_s + \Phi_{j,k} \tau_s + \tilde{e}_{j,k}, \\ \Phi_{j,k+1} = \chi_j \text{sign}(\tilde{e}_{j,k}) \tau_s + \Phi_{j,k}. \end{cases} \quad (3.18)$$

Theorem 3.3.1 *Considering the power estimation system (3.18), supposed that $\lambda_j > 0$ and $\chi_j > 0$, if there exists a positive-definite matrix $\mathbf{P}_j = \mathbf{P}_j^T$ for given $\mathbf{Q}_j = \mathbf{Q}_j^T > 0$ and $\mathbf{H}_j = \mathbf{H}_j^T > 0$, satisfying*

$$\mathbf{A}_j^T(\mathbf{P}_j + \mathbf{P}_j \mathbf{H}_j^{-1} \mathbf{P}_j) \mathbf{A}_j - (1 - \zeta_j) \mathbf{P}_j + \mathbf{Q}_j \leq 0, \quad (3.19)$$

then the system (3.18) converges asymptotically to a ball $\mathbb{B}_j = \{\tilde{e}_{j,k} : \|\tilde{e}_{j,k}\|^2 < r_j\}$, and with a radius

$$r_j = \frac{h_j}{1 - \zeta_j}, \quad (3.20)$$

where $0 < \zeta_j < 1$, $h_j = \bar{\sigma}_{j2} + \frac{1}{4} \bar{\sigma}_{j1}^2 \|Q_j^{-1}\|^2$, $\bar{\sigma}_{j1} = \sigma_{j1} + \sigma_{j2}^2 \varpi_j$, $\bar{\sigma}_{j2} = \sigma_{j3} + \sigma_{j2}^2 \varpi_j^{-1}$, $\sigma_{j1} = \lambda_j^2 \tau_s^2 z_{j11}$, $\sigma_{j2} = \chi_j \lambda_j \tau_s^2 z_{j12}$, $\sigma_{j3} = \lambda_j^2 \tau_s^2 z_{j22}$, $\varpi_j \in R^+$, $\mathbf{Z}_j = [\mathbf{H}_j + \mathbf{P}_j] = [z_{j11}, z_{j12}; z_{j21}, z_{j22}]$, $\mathbf{P}_j$, $\mathbf{H}_j$, $\mathbf{Q}_j$ and $\mathbf{Z}_j \in \mathbb{R}^{2 \times 2}$.

Proof (Theorem 3.3.1), the (3.18) can be represented as

$$\Theta_{j,k+1} = \mathbf{A}_j \Theta_{j,k} + \mathbf{B}_{j,k} \text{sign}(\tilde{e}_{j,k}), \quad (3.21)$$

where $\Theta_{j,k} = [\tilde{e}_{j,k}, \Phi_{j,k}]^T$, $\mathbf{A}_j = [1, \tau_s; 0, 1]$, $\mathbf{B}_{j,k} = [\lambda_j |\tilde{e}_{j,k}|^{\frac{1}{2}} \tau_s, \chi_j \tau_s]^T$.

Considering (3.21), a Lyapunov function is selected as,

$$V_{j,k} = \Theta_{j,k}^T \mathbf{P}_j \Theta_{j,k} = \|\Theta_{j,k}\|_{\mathbf{P}_j}^2. \quad (3.22)$$

where $\mathbf{P}_j = \mathbf{P}_j^T \in \mathbb{R}^{2 \times 2}$ is positive definite matrix.

The difference equation of (3.22) can be expressed,

$$\begin{aligned} \Delta V_{j,k} &= V_{j,k+1} - V_{j,k} \\ &= \Theta_{j,k}^T (\mathbf{A}_j^T \mathbf{P}_j \mathbf{A}_j - \mathbf{P}_j) \Theta_{j,k} + \mathbf{B}_{j,k}^T \mathbf{P}_j \mathbf{B}_{j,k} \\ &\quad + 2 \Theta_{j,k}^T \mathbf{A}_j^T \mathbf{P}_j \mathbf{B}_{j,k} \text{sign}(\tilde{e}_{j,k}). \end{aligned} \quad (3.23)$$

Using the inequality,

$$\begin{aligned} 2 \Theta_{j,k}^T \mathbf{A}_j^T \mathbf{P}_j \mathbf{B}_{j,k} \text{sign}(\tilde{e}_{j,k}) &\leq \Theta_{j,k}^T \mathbf{A}_j^T \mathbf{P}_j \mathbf{H}_j^{-1} \mathbf{P}_j \mathbf{A}_j \Theta_{j,k} \\ &\quad + \mathbf{B}_{j,k}^T \mathbf{H}_j \mathbf{B}_{j,k}, \end{aligned} \quad (3.24)$$

where $\mathbf{H}_j = \mathbf{H}_j^T \in \mathbb{R}^{2 \times 2} > 0$.

The equation (3.23) can be represented as,

$$\begin{aligned} \Delta V_{j,k} &\leq \Theta_{j,k}^T [\mathbf{A}_j^T (\mathbf{P}_j + \mathbf{P}_j \mathbf{H}_j^{-1} \mathbf{P}_j) \mathbf{A}_j - (1 - \zeta_j) \mathbf{P}_j] \Theta_{j,k} \\ &\quad + \mathbf{B}_{j,k}^T (\mathbf{H}_j + \mathbf{P}_j) \mathbf{B}_{j,k} - \zeta_j \Theta_{j,k}^T \mathbf{P}_j \Theta_{j,k}, \end{aligned} \quad (3.25)$$

where $\zeta_j \in (0, 1)$.

Setting $\mathbf{Z}_j = \mathbf{H}_j + \mathbf{P}_j = [z_{j11}, z_{j12}; z_{j21}, z_{j22}]$,

$$\mathbf{B}_{j,k}^T \mathbf{Z}_j \mathbf{B}_{j,k} = \sigma_{j1} |\tilde{e}_{j,k}| + 2 \sigma_{j2} |\tilde{e}_{j,k}|^{\frac{1}{2}} + \sigma_{j3}, \quad (3.26)$$

where $\sigma_{j1} = \lambda_j^2 \tau_s^2 z_{j11}$, $\sigma_{j2} = \chi_j \lambda_j \tau_s^2 z_{j12}$, $\sigma_{j3} = \chi_j^2 \tau_s^2 z_{j22}$.

If there exists a positive definite matrix $\mathbf{Q}_j = \mathbf{Q}_j^T \in \mathbb{R}^{2 \times 2}$, that satisfies the LMI given by $\mathbf{A}_j^T (\mathbf{P}_j + \mathbf{P}_j \mathbf{H}_j^{-1} \mathbf{P}_j) \mathbf{A}_j - (1 - \zeta_j) \mathbf{P}_j = -\mathbf{Q}_j$, then $\Delta V_{j,k}$ is transformed into

$$\begin{aligned} \Delta V_{j,k} = & -\|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{Q}_j}^2 + \sigma_{j1} |\tilde{e}_{j,k}| + 2\sigma_{j2} |\tilde{e}_{j,k}|^{\frac{1}{2}} \\ & + \sigma_{j3} - \zeta_j \|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{P}_j}^2. \end{aligned} \quad (3.27)$$

For any $\varpi_j \in \mathbb{R}^+$, (3.27) can be written as

$$\begin{aligned} \Delta V_{j,k} \leq & -\|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{Q}_j}^2 + \bar{\sigma}_{j1} |\tilde{e}_{j,k}| \\ & + \bar{\sigma}_{j2} - \zeta_j \|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{P}_j}^2, \end{aligned} \quad (3.28)$$

where $\bar{\sigma}_{j1} = \sigma_{j1} + \sigma_{j2}^2 \varpi_j$, $\bar{\sigma}_{j2} = \sigma_{j3} + \sigma_{j2}^2 \varpi_j$.

Further, (3.28) can be presented as,

$$\begin{aligned} \Delta V_{j,k} \leq & -\|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{Q}_j}^2 + \bar{\sigma}_{j1} \|\mathbf{Q}_j \mathbf{Q}_j^{-1} \boldsymbol{\Theta}_{j,k}\|_1^2 \\ & + \bar{\sigma}_{j2} - \zeta_j \|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{P}_j}^2 \\ = & -\left(\|\mathbf{Q}_j \boldsymbol{\Theta}_{j,k}\| - \frac{1}{2} \bar{\sigma}_{j1} \|\mathbf{Q}_j^{-1}\|\right)^T \\ & \times \left(\|\mathbf{Q}_j \boldsymbol{\Theta}_{j,k}\| - \frac{1}{2} \bar{\sigma}_{j1} \|\mathbf{Q}_j^{-1}\|\right) \\ & - \zeta_j \|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{P}_j}^2 + \bar{\sigma}_{j2} + \frac{1}{4} \bar{\sigma}_{j1}^2 \|\mathbf{Q}_j^{-1}\|^2, \end{aligned} \quad (3.29)$$

and then (3.29) can be obtained as,

$$\Delta V_{j,k} \leq -\zeta_j \|\boldsymbol{\Theta}_{j,k}\|_{\mathbf{P}_j}^2 + h_j, \quad (3.30)$$

with $h_j = \bar{\sigma}_{j2} + \frac{1}{4} \bar{\sigma}_{j1}^2 \|\mathbf{Q}_j^{-1}\|^2$.

Based on (3.30), it can be obtained as,

$$V_{j,k+1} \leq -(1 - \zeta_j)^k V_{j,k_0} + \sum_{m=1}^k (1 - \zeta_j)^{(m-1)} h_j, \quad (3.31)$$

One has $\overline{\lim}_{k \rightarrow \infty} V_{j,k} \leq \frac{h_j}{1 - \zeta_j}$, and the radius of the region of convergence of the DSTO is defined as $r_j \leq \frac{h_j}{1 - \zeta_j}$.

It can be concluded that (3.18) exhibits asymptotic stability within the region of convergence $\mathbb{B}_j = \{\tilde{e}_{j,k} : \|\tilde{e}_{j,k}\|^2 < r_j\}$ as $k \rightarrow \infty$, that is $\boldsymbol{\Theta}_{j,k} = 0$. Thus, $\tilde{e}_{j,k} = 0$, i.e., $\tilde{e}_{j,k+1} = 0$ with the sampling period τ_s is very small. ■

Based on *Theorem 3.3.1*, it can be concluded that $\hat{g}_j \rightarrow g_j$. As a result, DSTO has strong robustness and a fast convergence rate against disturbances. Then the power estimation system (3.14) can be reexpressed as

$$e_{j,k+1} = -O_{j,k} \mu(e_{j,k}) + \hat{g}_{j,k} + e_{j,k}. \quad (3.32)$$

To further enhance robustness of power tracking, a QISM surface is constructed as follows.

3.3.2.2 Quasi-integral Sliding Mode (QISM) Surface Design

A QISM surface is designed based on instantaneous power tracking errors of the (3.9),

$$\begin{cases} s_{j,k} = \rho_j e_{j,k} + \varrho_j \delta_{j,k}, \\ \delta_{j,k} = \delta_{j,k-1} + e_{j,k}, \\ \delta_{j,k_0} = -\frac{\rho_j}{\varrho_j} e_{j,k}, \end{cases} \quad (3.33)$$

where $j \in \{p, q\}$, and $s_{j,k}$ is quasi-integral sliding function; ρ_j and ϱ_j are positive values; $\delta_{j,k}$ aims at eliminating the power steady-state error; δ_{j,k_0} is the initial value of (3.33).

Remark 3.3.1 *The initial value of the integral term, denoted as δ_{j,k_0} , plays a crucial role in determining the initial trajectory of (3.33). In order to eliminate the initial approaching phase and enhance system robustness against uncertainties and disturbances, $\delta_{j,k_0} = -\frac{\rho_j}{\varrho_j} e_{j,k}$ leads to $s_{j,k} = 0$. As a result, δ_{j,k_0} is located on the QISM surface.*

3.3.2.3 Quasi-integral Sliding Mode Control (QISM) Law Design

Combined (3.14) and (3.33), one gets

$$s_{j,k+1} = -\varphi_j (O_{j,k} \mu(e_{j,k}) - g_{j,k} - e_{j,k}) + \varrho_j \delta_{j,k}, \quad (3.34)$$

where $\varphi_j = \rho_j + \varrho_j$. Then one can obtain that

$$s_{j,k+1} - s_{j,k} = -\varphi_j (O_{j,k} \mu(e_{j,k}) - g_{j,k}) + \varrho_j e_{j,k}, \quad (3.35)$$

A discrete-time equivalent control law is utilized to derive the QISM, i.e., $s_{j,k+1} - s_{j,k} = 0$, then the equivalent control law of QISM is obtained by,

$$\mu(e_{j,k})^{eq} = \frac{\varrho_j}{\varphi_j O_{j,k}} e_{j,k} + \frac{1}{O_{j,k}} \hat{g}_{j,k}, \quad (3.36)$$

where $\hat{g}_{j,k} \rightarrow g_{j,k}$ by DSTO, and the switch control law of QISM is designed as follows,

$$\mu(e_{j,k})^{sw} = \frac{1}{O_{j,k}} \kappa_{jsw} \text{sign}(s_{j,k}), \quad (3.37)$$

where κ_{jsw} is a positive switching gain.

Combined (3.36) and (3.37), the DSTO-QISM is designed as,

$$\mu(e_{j,k}) = \frac{1}{O_{j,k}} \left(\frac{\varrho_j}{\varphi_j} e_{j,k} + \hat{g}_{j,k} + \kappa_{jsw} \text{sign}(s_{j,k}) \right). \quad (3.38)$$

Putting (3.38) into (3.35), one can easily arrive at

$$s_{j,k+1} - s_{j,k} = \varphi_j (g_{j,k} - \hat{g}_{j,k} - \kappa_{jsw} \text{sign}(s_{j,k})). \quad (3.39)$$

3.3.2.4 QISM Stability Analysis

Theorem 3.3.2 Consider the system (3.39), let φ_j and κ_{jsw} are given positive constant. If there exist $\kappa_{jsw} > \Delta_g$ such that $|s_{j,k+1}| < |s_{j,k}|$, then the system (3.39) is global uniformly exponential stable with the upper bound of the lumped disturbance estimation error $|g_{j,k} - \hat{g}_{j,k}| \leq \Delta_g$.

Proof (Theorem 3.3.2), For $\forall k > 0$, satisfying $\kappa_{jsw} > \Delta_g$, if $s_{j,k} > 0$, (3.39) becomes,

$$s_{j,k+1} = s_{j,k} + \varphi_j(\Delta_g - \kappa_{jsw}) < s_{j,k}, \quad (3.40)$$

where Δ_g is the upper bound of lumped disturbance estimation error, i.e., $\Delta_g \geq |g_{j,k} - \hat{g}_{j,k}|$.

For $\forall k > 0$ and $\kappa_{jsw} > \Delta_g$, if $s_{j,k} < 0$, (3.39) obtains as,

$$s_{j,k+1} = s_{j,k} + \varphi_j(-\Delta_g + \kappa_{jsw}) > s_{j,k}, \quad (3.41)$$

Therefore, combined (3.40) and (3.41), we have

$$|s_{j,k+1}| < |s_{j,k}|. \quad (3.42)$$

It can be concluded that (3.42) satisfies the necessary and sufficient condition of the existence of QISM. ■

Therefore, the state of system (3.35) using the proposed method can finally enter the QISM, indicating that the system is stable. i.e., power tracking errors $e_{j,k} \rightarrow 0$, that is, active power $p \rightarrow p^*$ and reactive power $q \rightarrow q^*$.

3.3.3 Voltage Regulation Loop

A DSTO-based QISM method can ensure accurate and rapid DC-link voltage control while canceling multiple disturbances (including unknown loads, capacitance mismatches, bias caused by current path changes, and switch mode noise) and suppressing chattering. Assume the equivalent dc-load as an external disturbance, and rewrite (3.2) follows,

$$z_{k+1} = \frac{2\tau_s}{C} p_k^* - \frac{\tau_s}{C} d_k + z_k, \quad (3.43)$$

where $z_k = \frac{1}{2} v_{dc,k}^2$, $p_k^* = \frac{1}{2} \mathbf{i}_{\alpha\beta,k}^{*T} \mathbf{u}_{\alpha\beta,k} v_{dc,k}$, and d_k is considered as the multi-disturbances, including capacitance mismatch, unknown dc-load R_L , bias caused by current path changes and switch mode noise.

(3.43) can be represented,

$$z_{k+1} = 2E p_k^* - E d_k + z_k, \quad (3.44)$$

where $E = \frac{\tau_s}{C}$. Define voltage regulation error.

$$e_k = z_k - z_k^*, \quad (3.45)$$

where z_k^* is the desired value for z_k . Since the sampling period τ_s is very small, then $z_k^* = z_{k+1}^*$.

Submitting (3.44) into (3.45),

$$e_{k+1} = 2E p_k^* - E d_k + e_k. \quad (3.46)$$

Based on the (3.46), the voltage estimation system is obtained,

$$\hat{e}_{k+1} = 2E p_k^* - E \hat{d}_k + \hat{e}_k. \quad (3.47)$$

where $\hat{d}_k$ is the estimated value of d_k . To estimate the multi-disturbances $\hat{d}_k$ of (3.47), a DSTO is designed as follows.

3.3.3.1 DSTO Design

Defined the DSTO error is regarded as $\tilde{e}_k = e_k - \hat{e}_k$, it can be obtained

$$\tilde{e}_{k+1} = E(\hat{d}_k - d_k) + \tilde{e}_k, \quad (3.48)$$

obviously, if $\tilde{e}_{k+1} = \tilde{e}_k = 0$, such that $\hat{d}_k = d_k$. Then the DSTO can be designed as based on (3.21)-(3.31),

$$\begin{cases} \hat{d}_k = -\epsilon |\tilde{e}_k|^{\frac{1}{2}} \text{sign}(\tilde{e}_k) + \Gamma_k \\ \Gamma_{k+1} = -\varepsilon \text{sign}(\tilde{e}_k) \tau_s + \Gamma_k, \end{cases} \quad (3.49)$$

where ϵ and ε are positive value. Γ_k aims at eliminating the steady-state error.

3.3.3.2 QISM Surface Design

To improve the robustness of dc-link voltage, a QISM surface is built,

$$\begin{cases} s_k = \iota e_k + \nu f_k, \\ f_k = f_{k-1} + e_k, \\ f_{k_0} = -\frac{\iota}{\nu} e_k, \end{cases} \quad (3.50)$$

where ι and ν are positive value; s_k is a quasi-integral sliding mode function; f_k aims at eliminating the steady-state error; f_{k_0} is the initial value.

3.3.3.3 QISM Law Design

Similar to (3.34) to (3.37), DSTO-QISM of dc-link voltage is designed as,

$$p_k^* = -\frac{\nu}{2E\nu} e_k + \frac{\hat{d}_k}{2} - \kappa \text{sign}(s_k), \quad (3.51)$$

where $\nu = \iota + \nu$, κ is a positive switching gain.

The proof procedure can be obtained with similar arguments to those described in *Theorem 3.3.1* and *3.3.2*, respectively. As a result, it can be concluded that the system (3.51) is stable by utilizing the proposed control approach DSTO-based QISM, i.e., $z_k \rightarrow z_k^*$, that is v_{dc} can track v_{dc}^* .

3.3.4 Voltage Balancing Loop

To balance the dc-link capacitor voltage, a simple discrete-time proportional integral (DPI) method is implemented to maintain the capacitor voltage difference v_{dc}^o close to zero. Then (3.3) can be rewritten as,

$$v_{dc,k+1}^o = \frac{2p_k^* \tau_s}{C\sqrt{6}v_{dc,k}} u_{\gamma,k} + \frac{1}{C} \Xi, \quad (3.52)$$

where $u_{\gamma,k}$ is control input, and Ξ is unknown disturbances.

3.3.4.1 Discrete-time Proportional-Integral (DPI) Design

A DPI is implemented as,

$$u_{\gamma,k} = k_{bp}\tilde{v}_{dc,k}^o + k_{bi} \sum_{k=0}^n \tilde{v}_{dc,k}^o, \quad (3.53)$$

where $k_{bp} > 0$ and $k_{bi} > 0$ are control gains; $\tilde{v}_{dc,k}^o = v_{dc,k}^o - v_{dc,k}^{o*}$, in which $v_{dc,k}^{o*}$ is reference of $v_{dc,k}^o$.

3.4 Discrete-time Super-Twisting Observer-based Quasi-Integral Sliding Mode Control (DSTO-QISM) Tuning for the Power Tracking Loop

3.4.1 DSTO Parameters Tuning Process

The following *Lemma 3.4.1* is given for subsequent developments.

Lemma 3.4.1 (see [133]): *Consider the nonlinear system (3.17). Given positive scalars λ_j , χ_j , if there exists $\lambda_j^* > 0$, such that $\lambda_j > \lambda_j^*$, then the nonlinear system (3.17) exhibits asymptotic stability within finite-time intervals, and the convergence time cannot be less than $\frac{\Phi_{j,k_0}}{\chi_j}$, where Φ_{j,k_0} is the initial value of $\Phi_{j,k}$. In fact, one can select $\lambda_j > \max\{\sqrt{\chi_j/2m}, 2(\chi_j^2 m/n^3)^{1/4}\}$, where m is the mathematical constant, and n is a positive constant satisfying $0 < n < 1$, if $\Phi_{j,k_0} = 0$. If $\lambda_j \rightarrow \infty$, the convergence time tends to 0.*

Therefore, based on *Lemma 3.4.1*, the following procedures can be employed for tuning the DSTO parameters: (1) Noticing that $\Phi_{j,k}$ does not affect the stability of the system, it is possible to set $\Phi_{j,k_0} = 0$ for simplicity; (2) Taking into account the parameter $\chi_j > 0$; (3) Gradually increasing the value of λ_j can be employed to reduce the settling time; (4) Raising the value of χ_j can further reduce the settling time, but it is important to be cautious not to make it too large to avoid excessive steady-state error in the system.

Certainly, increasing the values of λ_j and χ_j can potentially result in overshoot. Nevertheless, it is essential to emphasize that achieving the optimal parameters for DSTO remains an open problem. While one approach has been provided for fine-tuning the control parameters in the proposed scheme, a trial-and-error process is still essential to attain the highest performance from the power converter.

3.4.2 QISM Surface Parameters Tuning Process

The closed dynamics of the power tracking errors are derived by substituting (3.38) into (3.14) as

$$e_{j,k+1} = -\frac{\rho_j}{\varphi_j} e_{j,k}^* + e_{j,k} + \text{QISM}_k, \quad (3.54)$$

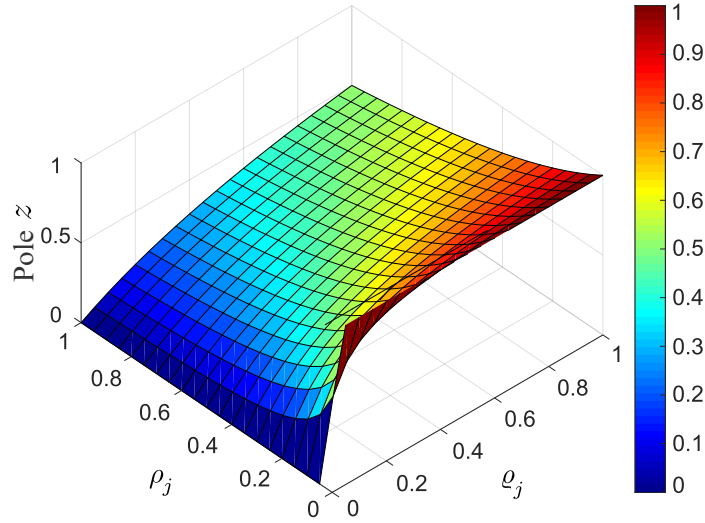


Figure 3.3: Relationship between parameters ρ_j , q_j and closed-loop pole z .

where $e_{j,k}$ influenced by (3.38), can be denoted as $e_{j,k}^*$, representing the reference value of $e_{j,k}$, which should be close to zero. $QISM_k = g_{j,k} - \hat{g}_{j,k} - \kappa_{jsw} \text{sign}(s_{j,k})$ is a very small external disturbance associated with the QISM band, and can be ignored in the steady state. Subsequently, the z -domain power transfer function of the (3.54) is expressed as

$$\frac{e_j(z)}{e_j^*(z)} = \frac{\rho_j}{(1-z)\varphi_j}, \quad (3.55)$$

with closed-loop pole as

$$z = 1 - \frac{1}{\varphi_j} = 1 - \frac{1}{\rho_j + q_j} \in (0, 1), \quad (3.56)$$

where $\varphi_j = \rho_j + q_j$. The relationship between parameters ρ_j and q_j , and the pole z are shown in Figure 3.3. One can infer from (3.56) that achieving the desired control bandwidth can be obtained by using a pole placement strategy [134]. A trade-off should be made in the selection of parameters. For simplicity, ρ_j can be set to 1, making q_j the sole design parameter. A faster dynamic response can be achieved when the pole is placed closer to the origin with a larger q_j , while positioning the pole nearer to the unit circle enhances robustness.

3.4.3 QISM Parameter Tuning Process

Based on the stability condition of *Theorem 3.3.2* is as

$$\kappa_{jsw} > \Delta_g, \quad (3.57)$$

where $\Delta_g \geq |g_{j,k} - \hat{g}_{j,k}|$ is the upper bound of multi-disturbance estimation error. It can be inferred that the switching gain κ_{jsw} influenced by the multi-disturbances. Therefore, by selecting an appropriate κ_{jsw} that satisfies (3.57), the stability of the power QISM can still be guaranteed with control saturation. Under κ_{jsw} , the system states would gradually be driven to the sliding surface. The proposed power controller would undergo a de-saturation process,

and the system described by (3.14) will remain stable. Certainly, a trade-off should be made in selecting parameters κ_{jsw} . The smaller the value of κ_{jsw} , the less chattering there is, while a larger κ_{jsw} value results in more chattering. Consequently, the choice of the κ_{jsw} value must be coordinated with the observer compensation gain. Continuous tuning is required until satisfactory control performance.

3.5 DSTO-QISMIC Tuning for the Voltage Regulation Loop

The voltage regulation DSTO-QISMIC parameters are adjusted the same way as power tracking loop controller parameters. Due to space constraints, this process will not be described in detail here.

3.6 Robustness and Chattering Suppression for the Power Tracking Loop

If only a QISMIC is designed based on the dynamic model (3.14), neglecting the multi-disturbance compensation by setting $\hat{g}_{j,k} = 0$ in (3.38), then the proposed DSTO-QISMIC method degrades into a QISMIC method. Following is an explanation of how the proposed DSTO-QISMIC method is able to suppress chattering and improve robustness in power tracking loop.

Assumed that multi-disturbance in (3.14) is bounded with

$$|g_{j,k}| < \bar{\Delta}_g. \quad (3.58)$$

Note that the sampling period τ_s is covered by $g_{j,k}$, and $\bar{\Delta}_g$ is the upper bound of the multi-disturbances.

Based on (3.39), the QISMIC dynamics of power are obtained,

$$s_{j,k+1} = \varphi_j(g_{j,k} - \hat{g}_{j,k} - \kappa_{jsw} \text{sign}(s_{j,k})) + s_{j,k}. \quad (3.59)$$

Setting $\hat{g}_{j,k} = 0$ in (3.59), then the stability condition of the QISMIC is obtained by (3.58).

Lemma 3.6.1 (see [133]) *Suppose that the function $\Phi_{j,k}$ is uniformly bounded and condition (3.17) holds. Then, the observation algorithm (3.17) with parameters ensures the convergence of the estimation errors to the domain $|\tilde{g}_{j,k}| \leq l_1 \tau_s^2$, $|\Phi_{j,k}| \leq l_2 \tau_s$, where $\tilde{g}_{j,k} = \hat{g}_{j,k} - g_{j,k}$, and l_1 and l_2 are some positive constants, depending on the observer parameters.*

According to Lemma 3.6.1,

$$|\hat{g}_{j,k} - g_{j,k}| \leq l_1 \tau_s^2 = \Delta_g, \quad (3.60)$$

where $l_1 > 0$, and Δ_g is the upper bound of the multi-disturbance estimation error of DSTO. Comparing (3.58) and (3.60), it becomes evident that due to the small value of τ_s , Δ_g is smaller than $\bar{\Delta}_g$. This indicates that the proposed DSTO-QISMIC method effectively suppresses large disturbances into small estimation errors. Consequently, a smaller κ_{jsw} can be chosen to against the same multi-disturbances, essentially mitigating the chattering and enhancing steady-state

accuracy. Using the same κ_{jsw} , the QISM without DSTO would achieve poor robustness or may even lead to system instability.

The voltage regulation DSTO-QISM parameters are adjusted the same way as power tracking loop controller parameters. Due to space constraints, this process will not be described in detail here. The robustness and chattering suppression of DSTO-QISM for the voltage regulation loop are similar to the above. This analysis process will not be detailed here due to space limitations.

3.7 Experimental Results

This section validates the effectiveness of the proposed control scheme on a laboratory-built 3L-NPC power converter prototype, as shown in Figure 3.4. The proposed control scheme is implemented using the IMPERIX control platform. Parameters of the experimental platform are listed in Table 3.2, DSTO-QISM, discrete-time extended state observer-based-QISM (DESO-QISM), DSTO-based quasi sliding mode control (DSTO-QSMC), and discrete-time PI (DPI) control parameters are shown in Table 3.3. A comprehensive set of guiding principles is applied to comparative experimental analysis in order to ensure maximum objectivity and accuracy (DSTO-QISM, DESO-QISM, DSTO-QSMC, DPI) in the comparative analysis of control methods ¹.

Table 3.2: Parameters Used for Experiment

Description	Parameter	Value	Unites
Sampling frequency	f_s	10	kHz
Switching frequency	f_{sw}	10	kHz
Grid frequency	f	50	Hz
Grid phase voltage	v_{abc}	110	V
Phase inductor	L	5	mH
Dc-link capacitor	C	1551	uF
Load resistance	R_L	80 $\leftrightarrow$ 120	Ω
Reference output voltage	v_{dc}^*	370 $\leftrightarrow$ 450	V
Reactive power	q^*	0 $\leftrightarrow$ 2000	Var

3.7.1 An Overall Evaluation Comparison

To better validate the superiority of the proposed control scheme in overall control performance, the proposed DSTO-QISM control scheme is compared to the DESO-QISM, DSTO-QSMC,

¹These principles encompass: (1) Rigorous testing under identical conditions, including consistent hardware configurations and systems, as well as identical initial conditions to eliminate external variables. (2) Utilizing a fixed-variable analysis to accurately assess the impact of specific variables on the 3L-NPC power converter system. (3) Evaluation under the same input conditions (i.e., the same control goals). (4) The adoption of common performance metrics (stability, response speed, control error) to ensure consistency throughout the analysis. (5) Conducting diverse experiments to reveal the performance of these control methods across various real-world scenarios. The experimental comparisons are carried out with due attention to objectivity, both quantitatively and qualitatively, to be sure the assessments are accurate, reliable, and practical.

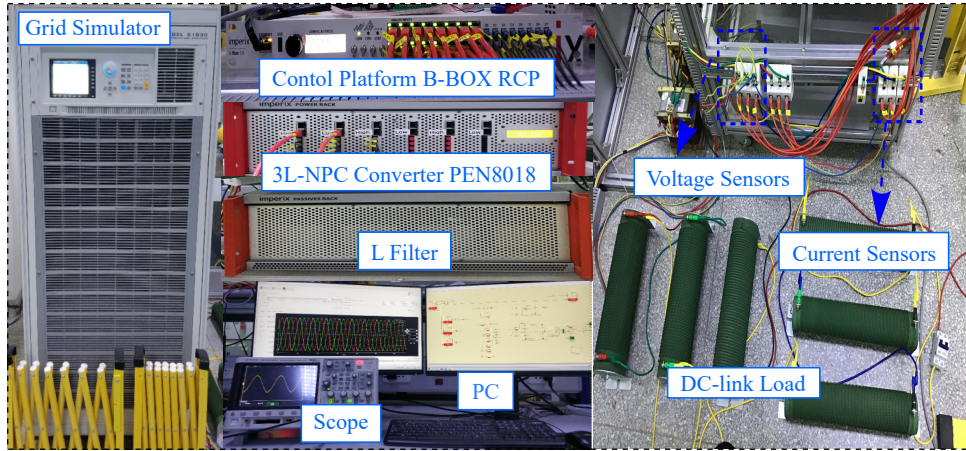


Figure 3.4: The lab-constructed 3L-NPC power converter prototype of Chap. 3.

Table 3.3: Controllers Parameters

Control Scheme		Power Tracking Loop	Voltage Regulation Loop	Voltage Balancing Loop
				DPI Control [†]
DSTO-QISM (Proposed)	DSTO $\diamond$	$\lambda_p = \lambda_q = 1e-4$ $\chi_p = \chi_q = 10$	$\iota = -5$ $\varepsilon = -6.4e-4$	$\begin{cases} k_{bp} = 0.5 \\ k_{bi} = 5e-3 \end{cases}$ [†]
	QISM $*$	$\rho_p = \rho_q = 0.22$ $\varrho_p = \varrho_q = 100$ $k_{psw} = k_{qsw} = 9e-4$	$\epsilon = 800$ $v = 100$ $k = 0.3$	
DESO-QISM	DESO	$\phi_p = \phi_q = \begin{bmatrix} 8e-1 \\ 8e-4 \end{bmatrix}$	$\eta = \begin{bmatrix} 4.8e3 \\ -5e1 \end{bmatrix}$	the same [†]
	QISM $*$	the same	the same	
DSTO-QSM	DSTO $\diamond$	the same	the same	the same [†]
	QSM	$\tau_p = \tau_q = 1e-3$ $\varsigma_{psw} = \varsigma_{qsw} = 9e-4$	$k_{z1} = 2e-2$ $k_{z2} = 3e-1$	
DPI	DP	$k_{pp} = k_{pq} = 4e-6$	$k_{vp} = 0.13$	the same [†]
	DI	$k_{ip} = k_{iq} = 1e-3$	$k_{vi} = 5$	

$\diamond$ indicates that the DSTO parameters remain consistent.

$*$ indicates that the DISMC parameters are the same.

[†] indicates that the DPI control parameters are consistent.

and DPI schemes. From Figure 3.5, it can be observed that the proposed control scheme outperforms the DESO-QISM, DSTO-QSM, and DPI methods in terms of transient/steady-state responses, including dc-link voltage, active power, reactive power, grid current quality (THDs) and chattering suppression. Specifically, the proposed control method had less overshoot/undershoot, faster transient response speed, smaller chattering, and lower THDs when regulating the dc-link voltage and tracking power. More detailed comparisons refer to Table 3.4. The underlying reason is that DSTO combines the robustness of sliding mode control with a high-order sliding mode control law to effectively address issues such as chattering, multi-disturbances (including load uncertainties, parameter variations, and bias caused by current path changes and switch mode noise) for power converter system, achieving rapid and accurate state estimation. Additionally, it self-compensates power/voltage bias in QISM, leading to significant improvements in voltage/power control performance. In QISM, the integral component serves the purpose of alleviating steady-state errors and enhancing system robustness compared with QSM. Overall, DSTO-QISM exhibits a good robustness and control performance.

Table 3.4: Comparison Of Steady/Transient States Based On Figure3.5

Figure3.5 (Indexes)		Control Schemes			
	Steady/ Transient States	DSTO-QISM	DESO-QISM	DSTO-QSM	DPI
Voltage (V)	Over-shooting(V)	0	4.5	3.3	6.2
	Over-settling Times(s)	0.055	0.073	0.109	0.084
	Enhancement(%) [‡]	-	24.6	49.5	34.5
	Under-shooting(V)	0	5.2	3.8	6.8
	Under-settling Times(s)	0.045	0.082	0.095	0.102
	Enhancement(%) [‡]	-	45.1	52.6	55.9
Power (W)	Over-settling Times(s)	0.0047	0.175	0.055	0.089
	Enhancement(%) [‡]	-	73.1	14.5	47.2
	Under-settling Times(s)	0.036	0.088	0.065	0.103
	Enhancement(%) [‡]	-	59	44.6	65
Power (Var)	Chattering	50	77	100	180
	$q^*=1500$ (t=0.7s)	-	35.1	50	72.2
Grid (A)	THDs(%)	3.0575	3.4062	3.4198	3.5382
	Enhancement(%) [‡]	-	10.2	10.6	13.6

[‡] indicates that the percentage improvement of DSTO-QISM relative to DESO-QISM, DSTO-QSM, and DPI, respectively.

3.7.2 Comparative Voltage Regulation Performance

The effectiveness and superiority of DSTO-QISM voltage regulation are verified by comparing three kinds of voltage regulators (DESO-QISM, DSTO-QSM, DPI), and a comparative

Table 3.5: Comparison Of Steady/Transient States Based Of Figure3.6

Figure3.6 (Indexes)	Control Schemes			
Steady/ Transient States	DSTO-QISM C	DESO-QISM C	DSTO-QSM C	DPI
Under-shooting(V)	0	5.2	0	8.2
Over-shooting(V)	0	4.8	0	6.5
Under-settling Times(s)	0.03	0.07	0.1	0.09
Enhancement(%) [♡]	-	57.1	70	66.7
Over-settling Times(s)	0.04	0.07	0.06	0.09
Enhancement(%) [♡]	-	42.9	33.3	55.6
Ripple(V)	0.19	0.2	1.48	0.21
Mean Error(V) [§]	0.0144	0.0322	0.2409	0.0461
Mean Square Error(V) [§]	0.1346	0.1416	0.2999	0.1668
Variance Error(V) [§]	0.0184	0.0190	0.0319	0.0257

[♡] indicates that the improvement of DSTO-QISM C relative to DESO-QISM C, DSTO-QSM C, and DPI, respectively.

[§] indicates that times is 0.2s-0.25s.

Table 3.6: Comparison Of Steady/Transient States Based Of Figure3.7

Figure3.7 (Indexes)	Control Schemes			
Steady/ Transient States	DSTO-QISM C	DESO-QISM C	DSTO-QSM C	DPI
Over-shooting(Var)	0	200	0	132
Over-settling Times(s)	0.008	0.024	0.016	0.027
Enhancement(%) [‡]	-	66.7	50	70.4
Under-shooting(Var)	0	200	0	288
Under-settling Times(s)	0.009	0.021	0.013	0.029
Enhancement(%) [‡]	-	57.1	30.8	69
Chattering(Var)	86	130	130	144
Enhancement(%) [‡]	-	33.8	33.8	40.3
THDs(%)	2.1422	2.3403	2.3124	2.3335
Enhancement(%) [‡]	-	8.5	7.4	8.2
Mean Error(Var) [‡]	0.016	0.3362	0.0333	0.0218
Mean Square Error(Var) [‡]	0.1526	0.7143	0.1602	0.6134
Variance Error(Var) [‡]	0.0231	0.3980	0.0246	0.3765

[‡] indicates that the improvement of DSTO-QISM C relative to DESO-QISM C, DSTO-QSM C, and DPI, respectively.

[‡] indicates that times is 0.3s-0.4s.

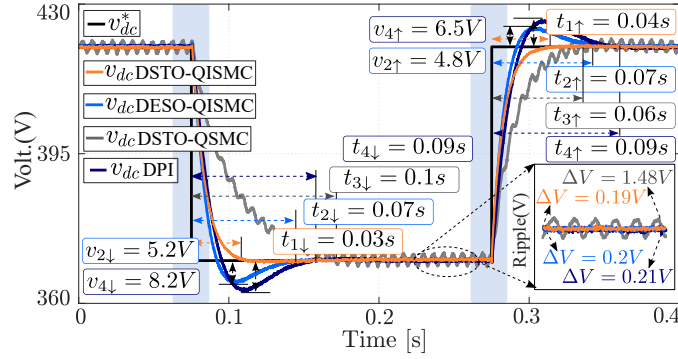


Figure 3.6: Comparative capacitor voltage regulation performance under voltage reference v_{dc}^* from $420\text{V} \rightarrow 370\text{V} \rightarrow 420\text{V}$, $R_L=120\Omega$, and $q^* = 0$.

analysis is performed in Figure 3.6. The DPI method is employed for the power tracking loop and balancing voltage loop during these tests. It proves that the proposed control method achieves better transient and steady-state performances. Accordingly, the proposed method exhibits minor voltage amplitude variations, smaller over- and undershooting, better chattering suppression, and faster convergence properties than DESO-QISM, DSTO-QSM, and DPI methods. Meanwhile, Table 3.5 provides a more detailed dynamic and static comparison of voltage regulation performance, enabling an evaluation of effectiveness and feasibility of the proposal. The main reason is that DSTO is capable of effectively estimating multi-disturbances as the voltage reference changes, so as to further offset QISM power/voltage bias, thereby improving voltage regulation performance. DSTO is combined with QISM to further suppress chattering and increase the robustness of voltage regulation.

3.7.3 Comparative Power Tracking Performance

To evaluate the effectiveness of the proposed DSTO-QISM method in power tracking, comparisons are made with DESO-QISM, DSTO-QSM, and DPI, respectively, as depicted in Figure 3.7. The DPI method is employed for the voltage regulation loop and balancing voltage loop during these tests. As depicted in Figure 3.7, the proposed method exhibits superior performance in reactive power dynamics, reduced power chattering, and improved grid-current quality compared to DESO-QISM, DSTO-QSM, and DPI methods. It is attributed to the robustness achieved by DSTO against uncertainties and external disturbances as opposed to DESO and DPI methods, and QISM shows superior chattering suppression compared to QSM. Additionally, the proposed solution achieves lower harmonics, meaning fewer noises in the power tracking loop. The more detailed performance evaluation indicators of reactive power are shown in Table 3.6. It can be seen that the proposed method exhibits better performance than DESO-QISM, DSTO-QSM, and DPI methods. DSTO is robust against multiple disturbances compared to DESO and DPI. Moreover, QISM exhibits better chattering suppression compared to QSM.

3.7.4 Evaluating Control Performance via Trajectory Motion

The voltage/power control performance of the proposed DSTO-QISM and DESO-QISM methods are evaluated in the presence of unknown interference, parameter mismatch, and measurement noise through voltage/power smooth trajectory and travel distance on the QISM surface. Figure 3.8 shows trajectories of voltage regulation, and the working mode is to adopt

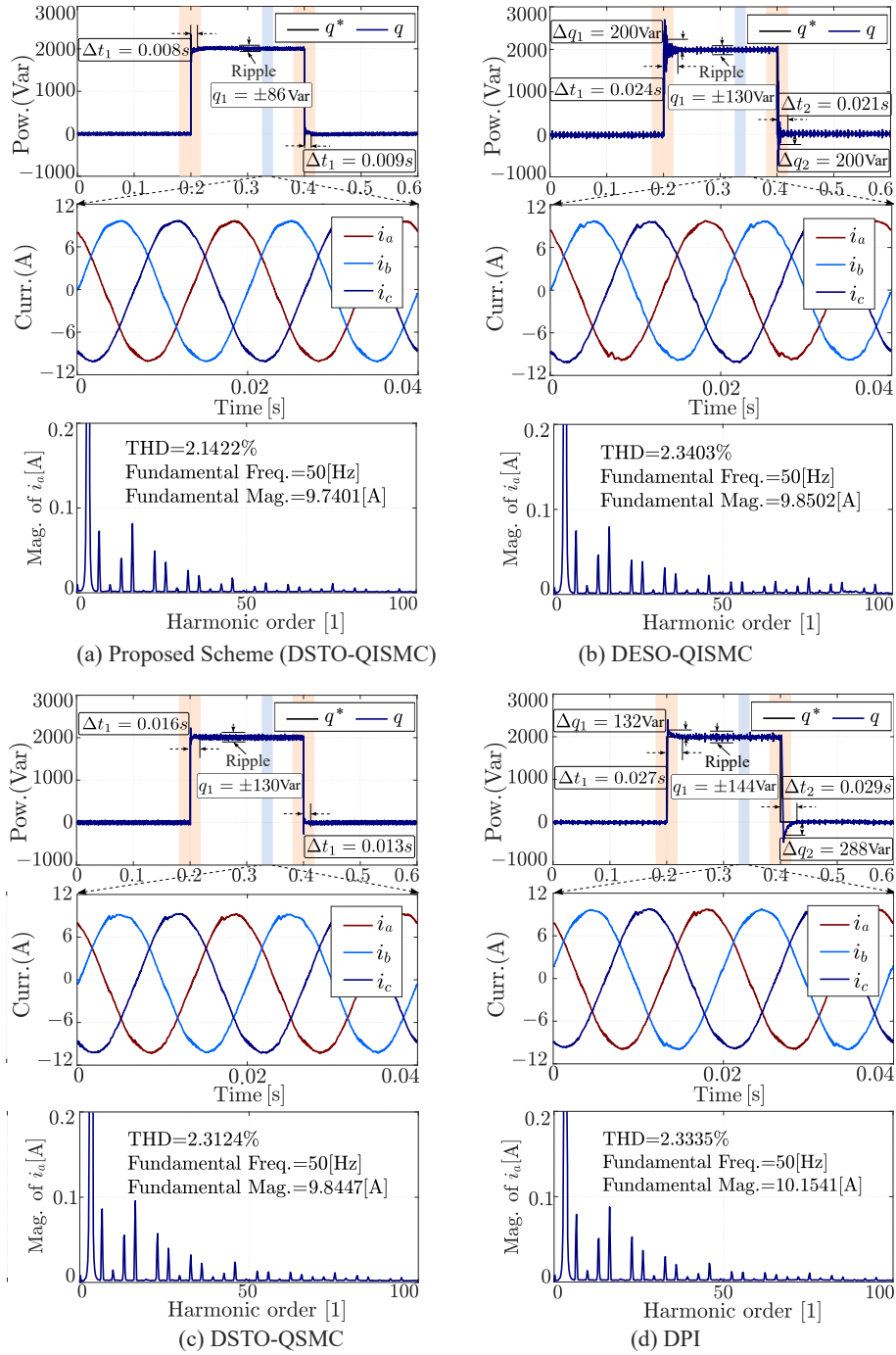


Figure 3.7: Comparative power tracking performance with q^* from 0 to 2000Var back to 0, $v_{dc}^*=400V$, and $R_L=120\Omega$.

DPI control in both power tracking and voltage balancing loops. Figure 3.9 illustrates trajectories of power tracking, where DPI control is employed in both the voltage regulation and voltage balancing loops.

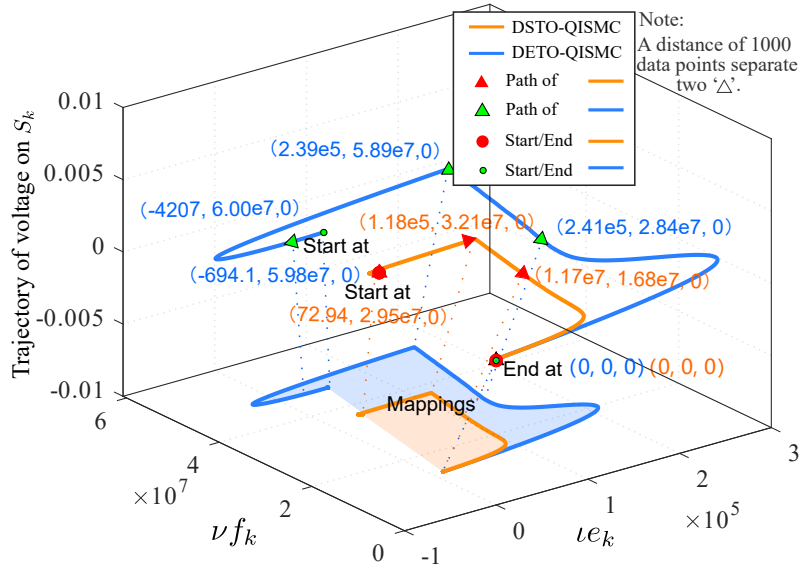


Figure 3.8: The trajectory of voltage regulation under the voltage reference v_{dc}^* from $420V \rightarrow 370V \rightarrow 420V$, $R_L=120\Omega$, and $q^*=0$.

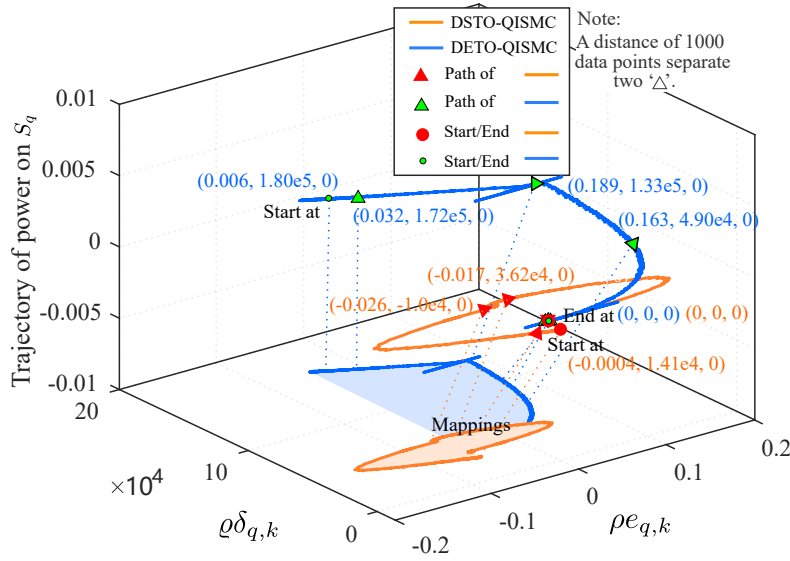


Figure 3.9: The trajectory of power tracking under the reactive power reference q^* from $0 \rightarrow 2000Var \rightarrow 0$, $R_L=120\Omega$, and $v_{dc}^*=400V$.

3.7.4.1 Voltage/Power Smooth Trajectory on the QISM Surface

The proposed DSTO-QISM method exhibits the stability of the voltage regulation and power tracking, as illustrated in Figs. 3.8 and 3.9, compared to the DESO-QISM method. Notably, the smoother the trajectory of the proposed DSTO-QISM method on the QISM surface, the better the stability performance of the control system. Notably, the smoother the trajectory of the proposed DSTO-QISM method on the quasi-integral sliding mode surface, the better the stability performance of the control system due to DSTO strong robustness for QISM bias compensate. Essentially, chattering is suppressed by the reduced variation of the sliding variable

that the system maintains while in sliding mode.

3.7.4.2 Voltage/Power Travel Distance on the QISM Surface

Figs. 3.8 and 3.9 show the distance traveled from the initial to final states on the QISM surface, illustrating that the DSTO-QISM provides more precise voltage/power control than the DESO-QISM. Essentially, when the sliding mode has reached zero, the smaller final distance associated with the sliding mode variable, the better control accuracy and the more robust system. It can be concluded that the DSTO-QISM control exhibits superior stability, accuracy, and rapidity in regulating voltage and tracking power. Essentially, when the sliding mode reaches zero, the smaller the final point distance associated with the sliding mode variable indicates the better the control accuracy and the more robust the system. This is primarily due to DSTO exhibiting asymptotic convergence, whereas DESO demonstrates bounded convergence.

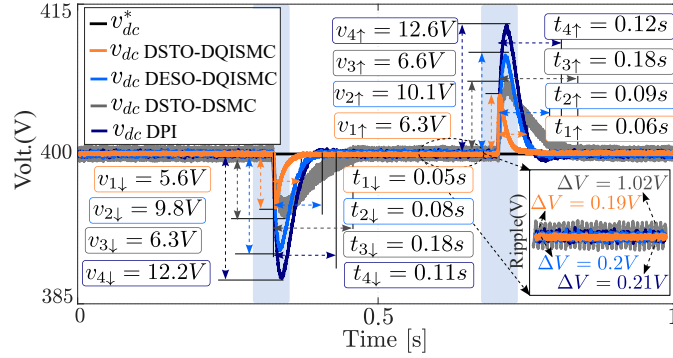


Figure 3.10: Comparative voltage regulation performance under dc-load R_L from 120Ω to 80Ω back to 120Ω , $v_{dc}^* = 400V$, and $q^* = 0$.

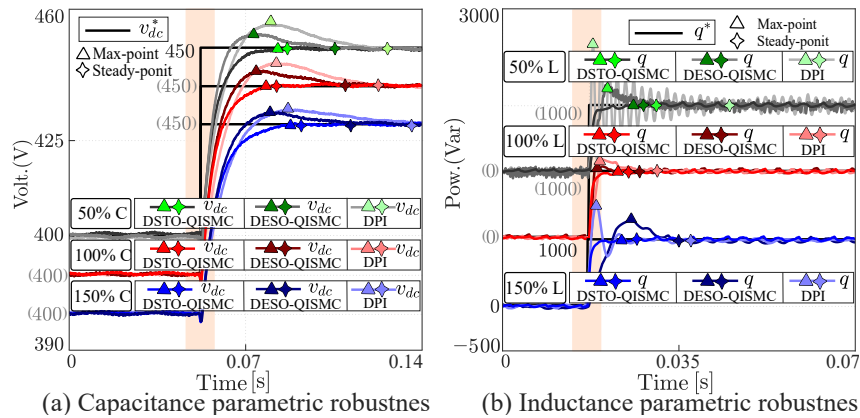


Figure 3.11: Capacitance parametric robustness comparison (transient-state) under (a) v_{dc}^* from 400V to 450V and R_L set at 120Ω with $q^* = 0$; (b) q^* from 0 to 1000Var, v_{dc}^* set at 400V with R_L sustained at 120Ω .

3.7.5 Robustness Evaluation with External Disturbance

A test bench is carried out to validate the robustness against external disturbance variations of the proposed method compared to DESO-QISM, DSTO-QSM, and DPI methods, as shown in Figure 3.10. The DPI method is employed for the voltage power tracking and balancing voltage loops during these tests. It can be concluded that the proposed voltage regulation method is superior to DESO-QISM, DSTO-QSM, and DPI methods in terms of voltage amplitude variations, chattering suppression, and convergence properties. As a result, the proposed method performs exceptionally well, indicating its excellence robust as a voltage regulation method. The critical factor behind this success lies in its ability to effectively estimate multiple disturbances arising from changes in the reference voltage. This capability compensates QISM power/voltage variations, resulting in a significant enhancement in voltage regulation. Consequently, DSTO-QISM demonstrates exceptional voltage regulation capabilities, even in scenarios characterized by load uncertainties.

3.7.6 Performance Evaluation under Parameter Mismatch

A performance evaluation under parameter mismatch is conducted by utilizing the DSTO-QISM, DESO-QISM, and DPI control methods. The evaluation involves varying the capacitance parameter (C) at 100% C , 50% C , and 150% C of the nominal capacitance and the inductance parameter (L) at 100% L , 50% L , and 150% L of the nominal inductance. The working condition requires the implementation of consistent control schemes in the voltage regulation loop, power tracking loop, and voltage balancing loop, specifically, DSTO-QISM, DESO-QISM, and DPI control, respectively.

3.7.6.1 Transient-state Performance with Parameter Mismatch

Figure 3.11 (a) and (b) illustrate the performance of the proposed DSTO-QISM, DSTO-QISM, DESO-QISM, and DPI control methods under capacitance/inductance parameter mismatch for voltage and power tracking, respectively. The results show that the proposed DSTO-QISM method exhibits greater robustness to capacitance/inductance mismatch compared to the DESO-QISM and DPI methods, across varying voltage regulation and power tracking conditions. Because DSTO is especially effective at mitigating ISM voltage and power bias, even in the presence of capacitance and inductance mismatches, leading to notable improvements in voltage and power tracking capabilities.

3.7.6.2 Steady-state Performance with Parameter Mismatch

Figs. 3.12 (a)-(f) describe comparative analyses of the steady-state voltage/power performance among the proposed DSTO-QISM, DESO-QISM, and DPI control methods under capacitance/inductance mismatch. The proposed method exhibits root mean square (RMS) voltage values within the range of 0.1204 $\rightarrow$ 0.1249, which shows more minor deviations between the voltage reference and actual voltage than the DESO-QISM (RMS value: 0.1224 $\rightarrow$ 0.1275) and DPI (RMS value: 0.2045 $\rightarrow$ 0.2198) methods, resulting in better parameter robustness against capacitance parameter mismatches as shown in Figs. 3.12 (a), (b), and (c). Additionally, the deviation of the power tracking process is experimentally studied based on DSTO-QISM,

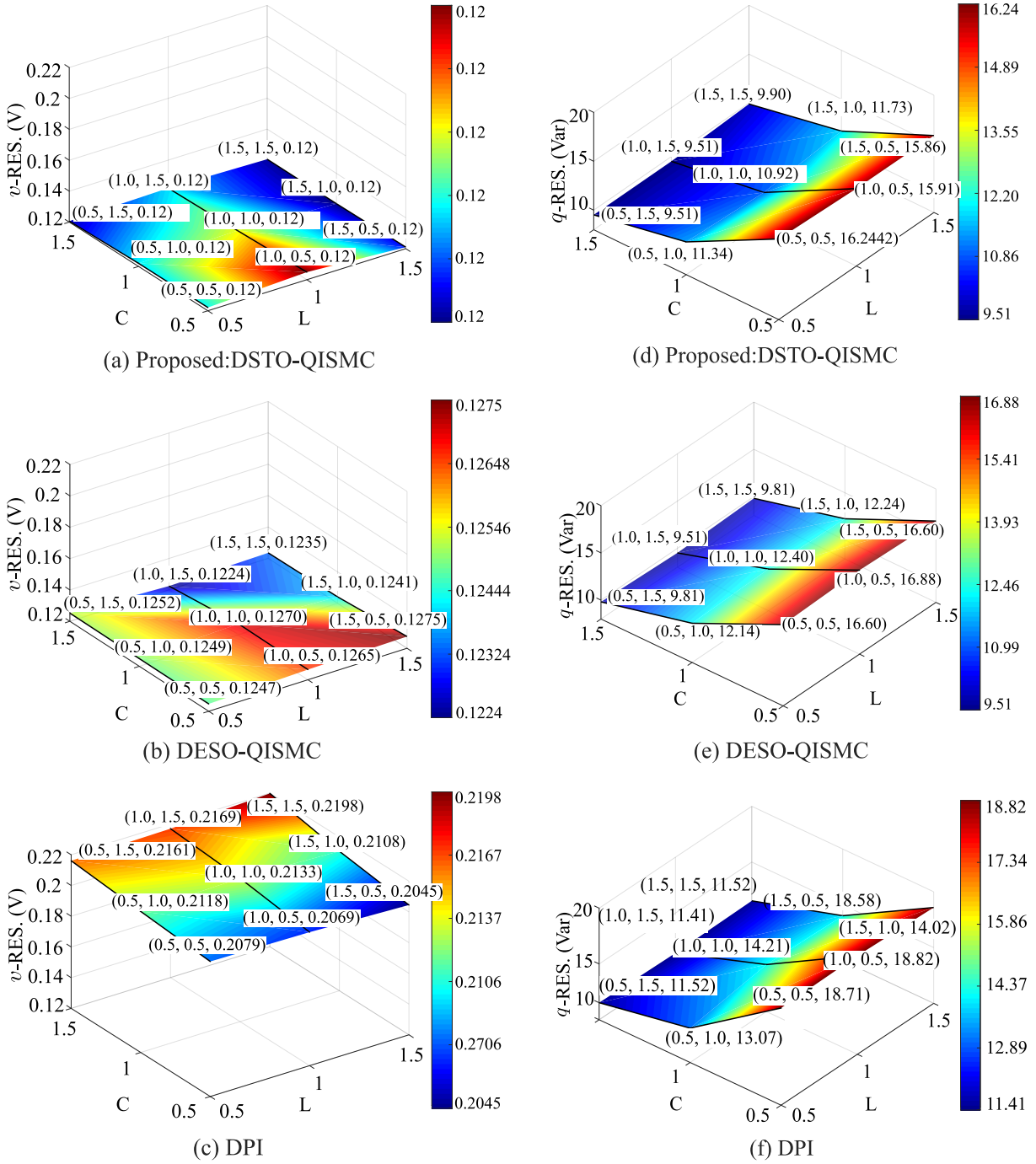


Figure 3.12: Parameter mismatch comparison (steady-state) under v_{dc}^* is 400V and R_L is 120Ω and $q^* = 0$. Capacitance mismatch: (a) DSTO-QISM, (b) DESO-QISM, and (c) DPI; Inductance mismatch: (d) DSTO-QISM, (e) DESO-QISM and (f) DPI.

DESO-QISM, and DPI control methods, as shown in Figs. 3.12 (d), (e), and (f). It can be concluded that the proposed DSTO-QISM control method is more robust with a smaller power RMS value (range: $9.5170 \rightarrow 16.2422$) against inductance parameter mismatches, in which the range power RMS value of DESO-QISM and DPI are (range: $9.5196 \rightarrow 16.8868$) and (range: $11.4133 \rightarrow 18.8286$), respectively. As a result, the proposed DSTO-QISM control scheme can dramatically have anti-disturbance capability against capacitance/inductance parameter mismatch. Two factors account for these: DSTO is more robust to multi-disturbances for compensating voltage/power QISM. QISM can adapt and correct for cumulative error, ensuring better tracking accuracy and improved disturbance rejection.

3.8 Summary

This chapter proposed a new DSTO-based QISM scheme for the 3L-NPC power converter in microgrids to address power ripple, voltage swings, and weak anti-disturbance, raising power/voltage regulation performance and anti-disturbance capability. A DSTO is used to offset multi-disturbances caused by parameter mismatches, unknown loads, current path changes, switch mode noise, and self-compensating power/voltage tracking biases in QISM. To further mitigate power/voltage steady-state error and boost system robustness, a new quasi-integral sliding mode surface is built, inherently improving power/voltage tracking performance. Rigorous stability analysis of the discrete-time dynamics of the corresponding power/voltage system provides a theoretical basis for the 3L-NPC power converter system. The experimental tests show that the proposed control outperforms the DESO-QISM, DSTO-QISM, and DPI control under different scenarios.

CHAPTER 4

Adaptive Sliding Mode Power-Voltage Control for 3L-NPC Power Converters

This chapter proposes a robust adaptive sliding mode control for the 3L-NPC power converter with multi-disturbances of microgrids. Figure 4.1 illustrates a simplified schematic of power converters in microgrids. Specifically, an AO-P voltage controller is proposed to accurately and quickly regulate dc-voltage in real-time identifying unknown equivalent dc-loads, compensating for the active power reference. To track the reference, a DO-ISMC is adopted to dramatically enhance the power tracking performance in case of parameter mismatches and bias caused by current path changes and switch mode noise. Additionally, an SMO-P control is established to balance dc-link (*two*) capacitors, rejecting harmonic injection and power ripple behaviors. Experimental data confirm that the proposed control scheme outperforms STO-ISMC, STA, and PI control schemes in the transient/steady state operations in terms of dynamic response, grid current harmonic distortions, and robustness.

4.1 Introduction

Some advanced control methods with strong anti-disturbance performance are proposed to achieve better dynamic response and steady-state performance in power converter systems, including AC [135], event-triggered optimal control [136] and artificial neural network control [137].

A novel robust model predictive control framework with the merits of fuzzy logic is proposed for power converter to enhance anti-disturbance capacities under different scenarios in [138]. To decrease harmonic components in the 3L-NPC inverter, an SMC approach is used in [139]. However, since disturbances are rejected by FBC rather than FFC compensation control, these control systems can not react immediately and quickly enough in the case of severe disturbances. To circumvent the limits of FBC approaches, FFC is proposed to directly and actively attenuate disturbances [140]. However, its applications are limited due to unmeasurable/difficult-to-detect

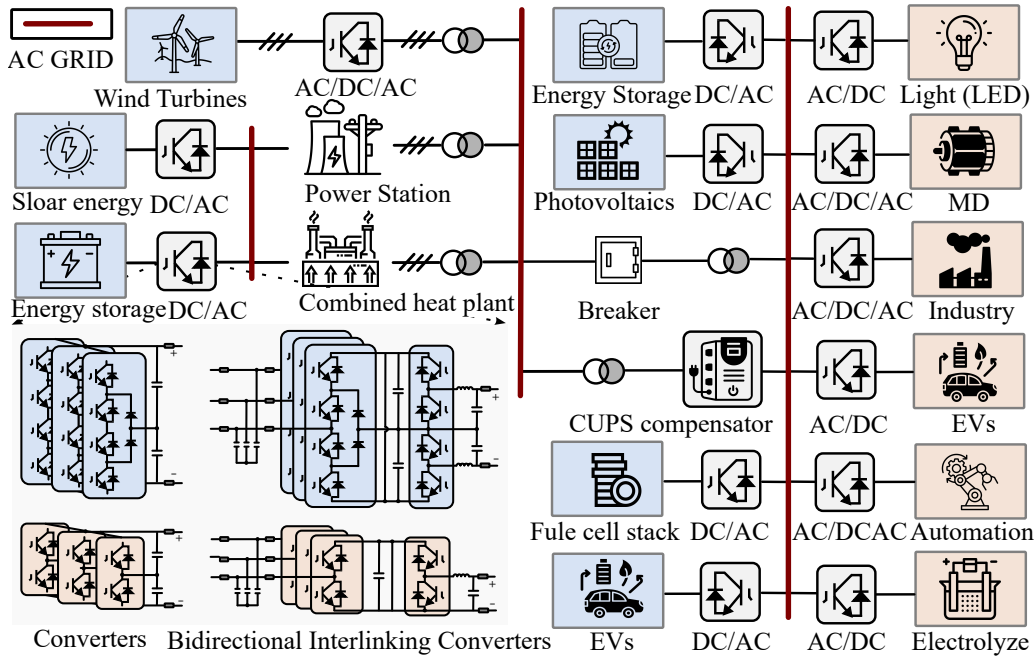


Figure 4.1: Diagram of power converters in microgrids.

disturbances or prohibitively expensive sensors. To make use of the benefits of FFC in disturbance rejection while also overcoming its drawbacks as stated above, OBC has received lots of interest in power converters, thanks to its flexible integrated control [20, 141].

To achieve accurate speed tracking performance of permanent magnet synchronous motor, an ESO-based SMC method is designed to improve disturbance rejection ability, especially, in face of internal parameter variations and external load changes in [142]. An ESO-based dc-link voltage controller is proposed in [143] to improve the output voltage performance. Although ESO has a good anti-disturbance ability against time-invariant disturbance, the disturbance can not be achieved through asymptotic estimating. A LO based PI controller is constructed to address the imbalance of the capacitor voltages and the presence of an ac-voltage low-frequency oscillation in the dc-link of the NPC power converter [144]. Note that the LO is a full-order state observer. A ROO- H_∞ controller is designed to balance the capacitor voltage in [95]. Since the nature of the ROO is still a LO, the high gain can result in a peaking phenomenon where the initial estimate error is considerably large. Therefore, an SMO is available which brings one estimated error to zero in finite time, even in the presence of measurement noise and the other states have the error that behaves similarly to the error in a LO after peaking has subsided [100]. Furthermore, the SMO boasts rapid convergence, exceptional disturbance rejection, and straightforward implementation. On the other hand, an AO fundamentally serves as an estimation mechanism where both the parameters and state variables of the system are simultaneously estimated. This process enhances the accuracy and robustness of state estimation, particularly with system changes and uncertainties. The selection between these two techniques hinges on system requirements, characteristics, and estimation objectives.

Furthermore, the power tracking control plays a critical role in the stable operation of power converter. Thus, SMC, as a robust control with a strong ability to counteract disturbances and uncertainties, has been extensively applied in power tracking to improve power performance [56].

Traditional SMC is merely insensitive to matched disturbances, as a consequence the application range is limited [145]. To attenuate mismatched disturbances, many researchers are concentrating on designing the specific sliding surface for uncertain systems. The integral sliding surface is one of the most popular methods, which can drive the state to the desired equilibrium with disturbances [146]. Although the ISMC method can handle the mismatched disturbances in a robust way, the chattering problem remains a serious problem to be addressed. It is feasible to utilize the STA one of the SOSMC methods to suppress the chattering effects [72]. However, a very large disturbance may result in require higher STA controller gains that lead to system chattering or system instability. By incorporating disturbance observers into the framework of ISMC, one can suppress chattering and deal with disturbances, resulting in a more robust controller. Employing a PO to estimate fast disturbance states, this approach is applied to a doubly-fed induction generator-based wind turbine, as presented by [147]. However, the PO must assume certain dynamic properties for all physical variables. In [97], UIOs are employed for each power system region to track the dynamic states in real-time operation, but this necessitates specific conditions; otherwise, accomplishing asymptotic tracking of disturbances remains challenging. To reduce constraint issues, a new filter is devised for an improved an EID-based estimator in [98]. It is evident that this approach will increase the design complexity. As part of this study, a DO is employed to estimate uncertainties and disturbances of the power tracking, improving its stability and performance. In this way, the DO can utilize estimated disturbances to feed-forward compensate the ISMC while reducing chattering in the ISMC. The DO-ISMC ultimately enhances the robustness and performance of the power tracking, which is an outstanding strategy for power tracking control.

Driven by the aforementioned considerations, this work presents an robust adaptive sliding mode control scheme in a multi-loop framework for the 3L-NPC power converter in microgrids. Contributions are summarized as follows.

- 1) An AO-P voltage controller is proposed to accurately and quickly regulate dc-link voltage with unknown equivalent dc-loads. With AO, the unknown disturbance is estimated in real-time, enabling compensated voltage inputs to be improved significantly. Besides, the whole proof process of stability analysis is strictly performed.
- 2) Power tracking controllers based on DO-ISMC are designed to dramatically enhance the power tracking performance. It is robust against parameter mismatches and bias caused by current path changes and switch model noise. The DO is applied to offset the disturbances, further alleviating the chattering caused by ISMC.
- 3) An SMO-P voltage balancing controller is constructed for balancing (*two*) capacitors. The SMO is utilized to estimate sinusoidal disturbances, preventing harmonic injections and power ripples caused by ac voltage low-frequency oscillations. Compared to the LO, the SMO dynamic can be reduced by one in sliding mode. This adjustment significantly minimizes delays, thus enhancing the observer's ability to estimate and compensate for disturbances.

4.2 System Description

The topology of the 3L-NPC power converter is shown in Figure 4.2A, in which a three-phase ac power source with inductors is connected to the ac side. The dc-link is connected to application scenarios, including photovoltaics, wind turbines, and electric vehicles, etc. Note that the dc-link

load change, as a disturbance, can result in abrupt voltage (active power) fluctuation. To illustrate this situation, the dc-link is connected to an unknown equivalent resistive load for simplification in the following modeling and analysis. The list of parameters of the power converter is presented in the Table 4.1.

Table 4.1: List of Nomenclatures

Symbol	Description
$\mathbf{e}_{abc}$	$[e_a, e_b, e_c]^T$, grid voltage vector in abc frame
$\mathbf{i}_{abc}$	$[i_a, i_b, i_c]^T$, grid current vector in abc frame
$\mathbf{v}_{\alpha\beta}$	$[v_\alpha, v_\beta]^T$, grid voltage vector in $\alpha\beta$ -axis.
$\mathbf{i}_{\alpha\beta}$	$[i_\alpha, i_\beta]^T$, grid current vector in $\alpha\beta$ -axis.
$\mathbf{u}_{\alpha\beta\gamma}$	$[u_\alpha, u_\beta, u_\gamma]^T$, average duty cycles in $\alpha\beta$ -axis.
VT_m^n	$n = \{a, b, c\}$, $m = \{1, 2, 3, 4\}$, insulated-gate bipolar transistors
VD_l^n	$n = \{a, b, c\}$, $l = \{1, 2\}$, clamping diodes
v_{dc}	$v_{dc}^h + v_{dc}^l$, the sum of dc-link capacitor voltages.
v_{dc}^o	$v_{dc}^h - v_{dc}^l$, the difference of dc-link capacitor voltages.
L, C	linear inductor, dc-link capacitor
R_L, i_L	total dc load resistance, dc-link load current
ω	angular frequency of grid voltage

Based on the 3L-NPC power converter topology [19], it can be obtained in stationary $\alpha\beta$ coordinates,

$$\dot{\mathbf{i}}_{\alpha\beta} = \frac{\mathbf{v}_{\alpha\beta}}{L} - \frac{1}{2L}v_{dc}\mathbf{u}_{\alpha\beta} + \frac{\mathbf{Z}}{L}, \quad (4.1)$$

$$\dot{v}_{dc} = \frac{\mathbf{u}_{\alpha\beta}^T \mathbf{i}_{\alpha\beta}}{C} - \frac{2v_{dc}}{CR_L}, \quad (4.2)$$

$$\dot{v}_{dc}^o = \frac{2}{\sqrt{6}C}\mathbf{N} + \frac{1}{\sqrt{6}C}\mathbf{u}_{\alpha\beta}^T \mathbf{i}_{\alpha\beta} u_\gamma, \quad (4.3)$$

where $\mathbf{Z} = [(\frac{1}{2\sqrt{6}}(u_\beta^2 - u_\alpha^2) - \frac{1}{\sqrt{3}}u_\alpha u_\gamma), \frac{u_\alpha u_\beta}{\sqrt{6}} - \frac{u_\beta u_\gamma}{\sqrt{3}}]^T v_{dc}^o$, $\mathbf{N} = [u_\alpha^2 - u_\beta^2, -u_\alpha u_\beta] \mathbf{i}_{\alpha\beta}$, and $\mathbf{u}_{\alpha\beta}$, u_γ are control signals.

Notice that the following assumptions are essential for the 3L-NPC power converter system (4.1)-(4.3) to operate correctly.

Assumption 4.2.1 *The source voltages are symmetrical and balanced.*

Assumption 4.2.2 *The averaged duty cycles u_α and u_β are defined as values in $[-1, 1]$ range so mentioned $\mathbf{Z}$ of (4.1) is two order of magnitude lower than the second term of (4.1).*

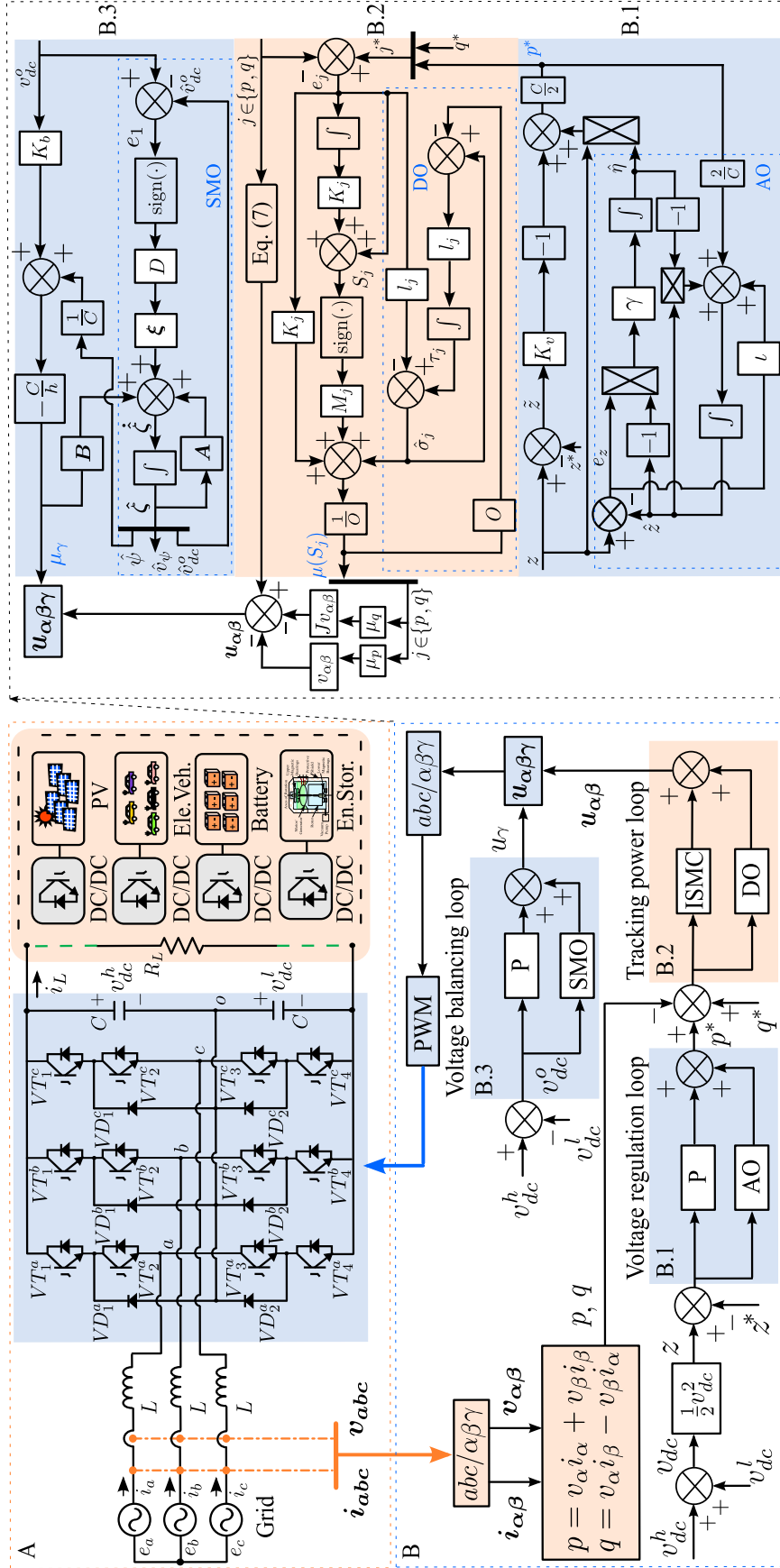


Figure 4.2: A cascade control scheme for the grid-connected 3L-NPC power converter with L filter.

Assumption 4.2.3 v_{dc}^o of equation (4.3) is lower by more than v_{dc} one order of magnitude, because v_{dc}^o must be near 0.

With the consideration of (4.1), the inductance current dynamics can be approximated,

$$\dot{\mathbf{i}}_{\alpha\beta} = \frac{\mathbf{v}_{\alpha\beta}}{L} - \frac{v_{dc}}{2L} \mathbf{u}_{\alpha\beta}, \quad (4.4)$$

Based on the instantaneous power theory, the instantaneous active and reactive powers p, q are defined as,

$$p = \mathbf{i}_{\alpha\beta}^T \mathbf{v}_{\alpha\beta}, \quad q = \mathbf{i}_{\alpha\beta}^T \mathbf{J} \mathbf{v}_{\alpha\beta}. \quad (4.5)$$

where $\mathbf{J} = [0, -1; 1, 0]$.

The derivative of (4.5) is obtained,

$$\begin{aligned} \dot{p} &= -\frac{1}{2L} \mathbf{v}_{\alpha\beta}^T \mathbf{u}_{\alpha\beta} v_{dc} + \frac{1}{L} \|\mathbf{v}_{\alpha\beta}\|^2 + \omega q, \\ \dot{q} &= -\frac{1}{2L} \mathbf{v}_{\alpha\beta}^T \mathbf{J}^T \mathbf{u}_{\alpha\beta} v_{dc} - \omega p, \end{aligned} \quad (4.6)$$

where $\|\mathbf{v}_{\alpha\beta}\|^2 = \|v_\alpha\|^2 + \|v_\beta\|^2$

The equilibrium point is obtained by [132] to address the (4.6) with $\dot{p} = 0$ and $\dot{q} = 0$, i.e.,

$$\mathbf{u}_{\alpha\beta}^{eq} = \frac{2}{v_{dc}} \left\{ \left(1 + \frac{\omega L q}{\|\mathbf{v}_{\alpha\beta}\|^2} \right) \mathbf{v}_{\alpha\beta} - \left(\frac{\omega L p}{\|\mathbf{v}_{\alpha\beta}\|^2} \right) \mathbf{J} \mathbf{v}_{\alpha\beta} \right\}. \quad (4.7)$$

where $\mathbf{u}_{\alpha\beta}^{eq}$ represents the equilibrium point of the system in steady state, i.e., instantaneous active and reactive power demanded by the power converter remain constant.

4.3 Control Strategy Design

In this section, a cascade control scheme is based on the approximated model derived from the analysis of the 3L-NPC power converter dynamics. Figure 4.2B presents the cascade control structure diagram of the 3L-NPC power converter.

4.3.1 Control Objectives

Three control goals need to be satisfied for the 3L-NPC power converter system: power tracking, voltage regulation, and capacitor voltage balancing. To achieve these goals, a robust adaptive sliding mode control is employed:

- In the voltage regulation loop, the dc-link voltage v_{dc} is regulated to follow its reference v_{dc}^* . The active power reference p^* , serving as the output of this loop, is the input of power tracking loop (see Figure 4.2. B.1);
- In the power tracking loop, the active and reactive powers, denoted as p and q , are forced to track their references p^* and q^* . Here, p^* is derived from the voltage regulation loop, while q^* is given to adjust the system power factor (see Figure 4.2. B.2);

- In the voltage balancing loop, the goal is to keep the capacitor voltage difference $v_{dc}^o \rightarrow 0$. The outputs from both this loop and the power tracking loop determine the final average duty cycle, controlling the power converter (see Figure 4.2. B.3).

4.3.2 Voltage Regulation Loop

In this subsection, an AO-P controller is devoted to regulating the sum of capacitor voltage v_{dc} to achieve its desired value in presence of the external disturbance.

Assumption 4.3.1 *The instantaneous power dynamics are much faster than the dc-link voltage dynamics, i.e., $p \approx p^*$.*

Assumption 4.3.2 *The equivalent load R_L , which is considered as external disturbance with slowing time varying.*

On the basis of the Assumptions 4.3.1 and 4.3.2, the dynamic of (4.2) can be rewritten as,

$$\dot{z} = -\eta z + \frac{2}{C}p^*, \quad (4.8)$$

where $z = \frac{1}{2}v_{dc}^2$, the desired active power $p^* = \frac{v_{dc}}{2}\mathbf{i}_{\alpha\beta}^{*T}\mathbf{u}_{\alpha\beta}$, C is known capacitance parameter, and $\eta = \frac{4}{R_L C}$ is regarded as an unknown parameter, with R_L representing the unknown dc-load.

4.3.2.1 Adaptive observer (AO) Design

To estimate η , based on (4.8), an AO is designed,

$$\dot{\hat{z}} = -\hat{\eta}\hat{z} + \frac{2}{C}p^* + \iota(z - \hat{z}), \quad (4.9)$$

where $\hat{z}$ and $\hat{\eta}$ are the estimate values of z , η , respectively. ι is the AO gain. Initially, when t_0 , $\hat{z}_{t_0} = 0$.

Next, through Lyapunov stability theory, it is proved that z and η can be asymptotically estimated by the AO, as follows.

Define the estimation errors $e_z = z - \hat{z}$, and $\tilde{\eta} = \eta - \hat{\eta}$. Their time-derivation can be expressed as,

$$\begin{cases} \dot{e}_z = (-\eta - \iota)e_z - \tilde{\eta}\hat{z}, \\ \dot{\tilde{\eta}} = \dot{\eta} - \dot{\hat{\eta}}, \end{cases} \quad (4.10)$$

Definition 4.3.1 is given for follow-up developments.

Definition 4.3.1 ([148]) *For the system (4.10), a set $\mathcal{X}$ is said to be invariant if every trajectory that starts from a point in $\mathcal{X}$ will remain in $\mathcal{X}$ for all future time, i.e.,*

$$\{e_z, \tilde{\eta}/e_z = 0, \tilde{\eta} = 0\} \in \mathcal{X} \Rightarrow \{e_z, \tilde{\eta}\} \in \mathcal{X}, \forall t \geq t_0. \quad (4.11)$$

Next, the stability condition for system (10) is presented in the following theorem, using adaptive technology.

Theorem 4.3.1 Consider the system (4.10), given $\gamma > 0$ and $\iota > 0$, if there exists a Lyapunov function $V_{(e_z, \tilde{\eta})}$ of (4.12) is a positive-definite function with a continuous first partial derivate $\dot{V}_{(e_z, \tilde{\eta})} \leq 0$ of (4.13), merely designing the adaptive law of (4.14), in finite region $B_R \subset \mathbb{R}$. Let $F = \{e_z, \tilde{\eta}/e_z = 0\} \subset B_R$ be set of all points where $\dot{V}_{(e_z, \tilde{\eta})} = 0$. Let $\mathcal{K} = \{e_z, \tilde{\eta}/e_z = 0, \tilde{\eta} = 0\}$ be the largest invariant set in F . Then, the every solution of e_z and $\tilde{\eta}$, starting in B_R approaches $\mathcal{K}$ as $t \rightarrow \infty$, such that, all trajectories of e_z and $\tilde{\eta}$ will asymptotically converge to 0 as $t \rightarrow \infty$.

Proof The Lyapunov function is constructed as,

$$V_{(e_z, \tilde{\eta})} = \frac{1}{2}e_z^2 + \frac{1}{2\gamma}\tilde{\eta}^2, \quad (4.12)$$

where γ is a positive constant. The time-derivation of the Lyapunov function (4.12) becomes,

$$\begin{aligned} \dot{V}_{(e_z, \tilde{\eta})} &= e_z \dot{e}_z + \frac{1}{\gamma}(\eta - \hat{\eta})(\dot{\eta} - \dot{\hat{\eta}}) \\ &= e_z [(-\eta - \iota)e_z - \tilde{\eta}\hat{z}] + \frac{1}{\gamma}(\eta - \hat{\eta})(-\dot{\hat{\eta}}) \\ &= (-\eta - \iota)e_z^2 - \tilde{\eta}\hat{z}e_z - \frac{1}{\gamma}\tilde{\eta}\dot{\hat{\eta}}. \end{aligned} \quad (4.13)$$

According to (4.13), the adaptive law is designed as follows

$$\dot{\hat{\eta}} = -\hat{z}e_z\gamma. \quad (4.14)$$

Then (4.13) becomes

$$\dot{V}_{(e_z, \tilde{\eta})} = (-\eta - \iota)e_z^2 \leq 0, \quad (4.15)$$

as $\eta > 0$ and $\iota > 0$. Since $\dot{V}_{(e_z, \tilde{\eta})}$ is negative semi-definite, then the origin is only stable in the Lyapunov sense, but not asymptotically stable as one would expect for the voltage regulation system. Let F be a set of all point where $\dot{V}_{(e_z, \tilde{\eta})} = 0$. Then

$$F = \{\{e_z, \tilde{\eta}\} \in \mathbb{R} : \dot{V}_{(e_z, \tilde{\eta})} = 0 \Rightarrow e_z = 0\}, \quad (4.16)$$

is a collection of all points that lie on the $\tilde{\eta}$ -axis. It follows that any point on this axis must satisfy

$$e_z \dot{e}_z + \frac{1}{\gamma}\tilde{\eta} = 0. \quad (4.17)$$

Based on (4.17), one way to find an invariant set is to evaluate the derivative of the function $V_{(e_z, \tilde{\eta})}$ that describes F and set it equal to zero. Hence,

$$\dot{e}_z = 0, \quad (4.18)$$

which is satisfied if and only if $\tilde{\eta} = 0$. Thus, the invariant set $\mathcal{K} \in F$ is a set that contains only the $(e_z = 0, \tilde{\eta} = 0)$, i.e., $\mathcal{K} = \{e_z = 0, \tilde{\eta} = 0\}$. However, $\mathcal{K}$ happens to be an origin. Then according to *Defination 4.3.1*, all trajectories of e_z and $\tilde{\eta}$ will converge to the 0, as $t \rightarrow \infty$, i.e., the system (4.10) is asymptotically stable. ■

Therefore, based on the proof, yields $\hat{z} \rightarrow z$ and $\hat{\eta} \rightarrow \eta$. The stability and convergence of AO are presented globally. The unknown dc-load is estimated completely in real-time.

Remark 4.3.1 *The AO is intended to estimate the state variable η . To estimate $\hat{\eta}$, utilizing the output signal z is essential, where z can be measured. Estimating $\hat{z}$ can deduce the state variable $\hat{\eta}$. While it is possible to directly estimate the state variable η using a DO, bounded convergence may pose challenges. Conversely, the AO can estimate η asymptotically. Additionally, although $\hat{z}$ can be estimated, the actual utilization in the subsequent voltage controller involves the measured value z . Compared to the estimated value $\hat{z}$, the measured value z is more genuine and reliable.*

Based on the AO, a composite AO-P controller is designed to regulate v_{dc} . Define the voltage error $\tilde{z} = z - z^*$, where $z^* = \frac{1}{2}v_{dc}^{*2}$. Then, it can be obtained that

$$\dot{\tilde{z}} = -\eta z + \frac{2}{C}p^*. \quad (4.19)$$

Proposition 4.3.1 *Consider the system (4.19) and AO estimation system (4.9) with well defined disturbance-to-output relative degree, then the closed-loop system under the composite controller (that is, AO-P) designed by (4.20) is asymptotically stable in the sense that for $K_v > 0$, such that $\tilde{z}$ are converge to 0 as $t \rightarrow \infty$.*

Proof

4.3.2.2 AO-based Proportional (AO-P) Voltage Controller Design

Based on (4.19), the AO-P voltage controller can be designed,

$$p^* = \frac{C}{2}(-K_v \tilde{z} + \hat{\eta} z), \quad (4.20)$$

where the positive constant K_v is the gain of p^* to be selected. Putting controller (4.20) into (4.19),

$$\dot{\tilde{z}} = -K_v \tilde{z} - (\eta - \hat{\eta})z, \quad (4.21)$$

is obtained, due to $\hat{\eta} \rightarrow \eta$ and if $K_v > 0$, then the voltage regulation error $\tilde{z}$ converges 0, as $t \rightarrow \infty$. ■

Therefore, the composite controller AO-P can regulate the capacitor voltages $v_{dc} \rightarrow v_{dc}^*$ and real-time unknown dc-loads.

4.3.3 Power Tracking Loop

A DO-ISMC power controller is applied to force the active power $p \rightarrow p^*$ and reactive power $q \rightarrow q^*$, where the DO is designed to estimate parameter mismatches, unmodeled dynamics, measurement errors, and unknown disturbances caused by current path changes and switch mode noise, simultaneously, suppressing chattering of the ISMC control.

The power tracking errors e_p and e_q are defined as,

$$e_p = p^* - p, \quad e_q = q^* - q, \quad (4.22)$$

where p^* is the reference of the active power from the voltage regulation output and q^* is the reference of the reactive power which can determine desired power factor.

Using (4.6), the time-derivative (4.22) can be expressed as,

$$\begin{aligned}\dot{e}_p &= -\omega q + \frac{\mathbf{v}_{\alpha\beta}^T v_{dc}}{2L} \mathbf{u}_{\alpha\beta} - \frac{\|\mathbf{v}_{\alpha\beta}\|^2}{L} + \dot{p}^*, \\ \dot{e}_q &= \omega p + \frac{(\mathbf{J}\mathbf{v}_{\alpha\beta})^T v_{dc}}{2L} \mathbf{u}_{\alpha\beta} + \dot{q}^*.\end{aligned}\quad (4.23)$$

Note that the above system is an ideal model. Nevertheless, in practical applications, converters are affected by some factors such as parameter mismatches, measurement errors, unmodeled dynamics, and bias caused by current path changes and switch mode noise, etc. Hence, it is necessary to establish a more accurate model that closely reflects the real system behavior. The system (4.23) can be rewritten,

$$\begin{aligned}\dot{e}_p &= -(\omega + \Delta_\omega)q + \left(\frac{1}{2L} + \Delta_L\right)\mathbf{v}_{\alpha\beta}^T v_{dc} \mathbf{u}_{\alpha\beta} \\ &\quad - \left(\frac{1}{L} + \Delta_L\right)\|\mathbf{v}_{\alpha\beta}\|^2 + \dot{p}^* + d_p, \\ \dot{e}_q &= (\omega + \Delta_\omega)p + \left(\frac{1}{2L} + \Delta_L\right)(\mathbf{J}\mathbf{v}_{\alpha\beta})^T v_{dc} \mathbf{u}_{\alpha\beta} \\ &\quad + \dot{q}^* + d_q,\end{aligned}\quad (4.24)$$

where Δ_ω and Δ_L are the parameter deviations, $\dot{p}^*$ and $\dot{q}^*$ are power dynamic references (regarded as unmodeled dynamics), and d_p and d_q denote measurement noise and bias arising from the current path changes and switch mode noise.

Then (4.24) can be rewritten as,

$$\begin{aligned}\dot{e}_p &= -\omega q + \frac{\mathbf{v}_{\alpha\beta}^T v_{dc}}{2L} \mathbf{u}_{\alpha\beta} - \frac{\|\mathbf{v}_{\alpha\beta}\|^2}{L} + \sigma_p, \\ \dot{e}_q &= \omega p + \frac{(\mathbf{J}\mathbf{v}_{\alpha\beta})^T v_{dc}}{2L} \mathbf{u}_{\alpha\beta} + \sigma_q,\end{aligned}\quad (4.25)$$

where

$$\begin{aligned}\sigma_p &= -\Delta_\omega q + \Delta_L \mathbf{v}_{\alpha\beta}^T v_{dc} \mathbf{u}_{\alpha\beta} - \Delta_L \|\mathbf{v}_{\alpha\beta}\|^2 + \dot{p}^* + d_p, \\ \sigma_q &= \Delta_\omega p + \Delta_L (\mathbf{J}\mathbf{v}_{\alpha\beta})^T v_{dc} \mathbf{u}_{\alpha\beta} + \dot{q}^* + d_q.\end{aligned}\quad (4.26)$$

Using the established equilibrium point in (4.7), the average duty cycle control signal $\mathbf{u}_{\alpha\beta}$ is designed as,

$$\mathbf{u}_{\alpha\beta} = \mathbf{u}_{\alpha\beta}^{eq} - \mu(S_p)\mathbf{v}_{\alpha\beta} - \mu(S_q)\mathbf{J}\mathbf{v}_{\alpha\beta}, \quad (4.27)$$

where $\mu(S_p)$ and $\mu(S_q)$ express the active and reactive power controllers, i.e., achieving $p \rightarrow p^*$ and $q \rightarrow q^*$, respectively.

Substituting (4.27) into (4.25), one can obtain,

$$\dot{e}_j = -O\mu(S_j) + \sigma_j, \quad (4.28)$$

where $j \in \{p, q\}$, $\mu(S_j)$ is controller and σ_j is the lumped disturbances, and $O = \frac{v_{dc}}{2L}\|\mathbf{v}_{\alpha\beta}\|^2$.

4.3.3.1 Disturbance Observer (DO) Design

Based on the system (4.28), a DO is designed to estimate σ_j ,

$$\begin{cases} \hat{\sigma}_j = \tau_j - l_j e_j, \\ \dot{\tau}_j = l_j (-O\mu(S_j) + \hat{\sigma}_j), \end{cases} \quad (4.29)$$

where $\hat{\sigma}_j$ is the estimation of lumped disturbances σ_j , τ_j is the auxiliary variable, and l_j is the DO gain.

Assumption 4.3.3 *Both the lumped disturbance σ_j and its derivative $\dot{\sigma}_j$ are bounded, i.e., $\|\sigma_j\| \leq \Gamma$ and $\|\dot{\sigma}_j\| \leq \aleph$.*

Theorem 4.3.2 *Suppose that Assumption 4.3.3 is satisfied for system (4.28). The estimate $\hat{\sigma}_j$ yielded by DO (4.29) can ultimately bounded estimate the lumped disturbances σ_j and converge in a region $\Omega = \{\tilde{\sigma}_j : \|\tilde{\sigma}_j\| \leq -\frac{\aleph}{l_j}\}$, where estimated error $\tilde{\sigma}_j = \sigma_j - \hat{\sigma}_j$, if the DO gain $l_j < 0$ in (4.29) is chosen.*

Proof The disturbance estimation error of (4.29) is defined as $\tilde{\sigma}_j = \hat{\sigma}_j - \sigma_j$. The dynamic of $\tilde{\sigma}_j$ can be obtained from (4.28)-(4.29),

$$\dot{\tilde{\sigma}}_j = l_j \tilde{\sigma}_j - \dot{\sigma}_j. \quad (4.30)$$

Construct a Lyapunov function,

$$V_{\tilde{\sigma}_j} = \frac{1}{2} \tilde{\sigma}_j^2. \quad (4.31)$$

Differentiating $V_{\tilde{\sigma}_j}$ and using (4.30), it can be obtained that

$$\dot{V}_{\tilde{\sigma}_j} = l_j \tilde{\sigma}_j^2 - \tilde{\sigma}_j \dot{\sigma}_j. \quad (4.32)$$

According to (4.30)-(4.32), one gets

$$\begin{aligned} \dot{V}_{\tilde{\sigma}_j} &\leq l_j \|\tilde{\sigma}_j\|^2 + \|\tilde{\sigma}_j\| \dot{\sigma}_j \\ &\leq \|\tilde{\sigma}_j\| (l_j \|\tilde{\sigma}_j\| + \dot{\sigma}_j). \end{aligned} \quad (4.33)$$

Therefore, after a sufficiently long time,

$$\|\tilde{\sigma}_j\| \leq -\frac{\aleph}{l_j}, (l_j < 0), \quad (4.34)$$

then the estimation error is ultimately bounded. Thus, it can be verified that the DO error system (4.30) is ultimately bounded stable convergence in a region $\Omega = \{\tilde{\sigma}_j : \|\tilde{\sigma}_j\| \leq -\frac{\aleph}{l_j} (l_j < 0)\}$ as $t \rightarrow \infty$, i.e., $\hat{\sigma}_j$ can estimate σ_j . ■

As a result, the DO can ultimately bounded estimate the lumped disturbances σ_j .

The following *Corollary 4.3.1* can be drawn if the lumped disturbances σ_j of Assumption 4.3.3 have constant values in steady-state, i.e., $\lim_{t \rightarrow \infty} \dot{\sigma}_j(t) = 0$.

Corollary 4.3.1 *Considering system (4.28), the estimate $\hat{\sigma}_j$ yielded by DO (4.29) can asymptotically track the lumped disturbances σ_j if the observer gain $l_j < 0$ in (4.29) is chosen.*

Proof According to (4.30)-(4.32) of the proof process in Theorem 4.3.2, the derivative of $\dot{V}_{\tilde{\sigma}_j}$ can be obtained when $\lim_{t \rightarrow \infty} \dot{\sigma}_j(t) = 0$, namely,

$$\dot{V}_{\tilde{\sigma}_j} = l_j \tilde{\sigma}_j^2, \quad (4.35)$$

merely satisfying $l_j < 0$, it can be obtained that $\dot{V}_{\tilde{\sigma}_j} < 0$, such that, $\tilde{\sigma}_j \rightarrow 0$ as $t \rightarrow \infty$. Thus, it can be verified that the error system (4.30) is asymptotically stable convergence to 0. ■

It can be concluded that $\tilde{\sigma}_j$ asymptotically converges to 0 as $t \rightarrow \infty$, i.e., $\hat{\sigma}_j$ can estimate σ_j , thereby countering the effects of lumped disturbances.

Remark 4.3.2 *According to Corollary 4.3.1, it appears that achieving asymptotic tracking of the DO is only feasible under the assumption that $\lim_{t \rightarrow \infty} \dot{\sigma}_j(t) = 0$. Nevertheless, the DO also effectively tracks fast time-varying disturbances, provided that the dynamics of the observer are significantly faster than those of the disturbances [149]. Additionally, Theorem 4.3.2 can be regarded as an extension of Corollary 4.3.1 when the dynamics of lumped disturbances cannot be ignored, i.e., $\lim_{t \rightarrow \infty} \dot{\sigma}_j(t) \neq 0$. In other words, Theorem 4.3.2 imposes fewer constraints than Corollary 4.3.1, allowing it to remain valid in addressing more complex disturbance scenarios. Hence, it is advisable to consider disturbance estimation using Theorem 4.3.2 for subsequent control design.*

Consequently, a robust ISMC power controller design will be predicated upon Theorem 4.3.2. Define the integral sliding function as follows,

$$S_j = e_j + K_j \int_0^t e_j(\tau) d\tau, \quad (4.36)$$

where K_j is the positive coefficient. The time-derivative of (4.36) becomes,

$$\dot{S}_j = -O\mu(S_j) + K_j e_j + \sigma_j. \quad (4.37)$$

Theorem 4.3.3 *Suppose the Theorem 4.3.2 is satisfied for the system (4.28). Considering system (4.28) under proposed DO-ISMC controller (4.38), the closed-loop system is asymptotically stable if the control gains $K_j > 0$, $M_j > -\frac{N}{l_j}$, and DO gain $l_j < 0$ in (4.38) is designed.*

Proof

4.3.3.2 DO-based Integral Sliding Mode Control (DO-ISMC) Power Controller Design

A composite DO-ISMC controller is designed as,

$$\mu(S_j) = \frac{1}{O} (K_j e_j + M_j \text{sign}(S_j) + \hat{\sigma}_j). \quad (4.38)$$

Putting the controllers (4.38) into (4.37) gives,

$$\dot{S}_j = -M_j \text{sign}(S_j) - \tilde{\sigma}_j. \quad (4.39)$$

Construct the following Lyapunov function,

$$V_j = \frac{1}{2} S_j^2. \quad (4.40)$$

The time-derivative of Lyapunov function (4.40) is,

$$\begin{aligned} \dot{V}_j &= S_j \dot{S}_j \\ &= -M_j S_j \text{sign}(S_j) - \tilde{\sigma}_j S_j, \\ &\leq \|S_j\|(-M_j + \tilde{\sigma}_j) \\ &\leq \|S_j\|(-M_j + \frac{\aleph}{-l_j}). \end{aligned} \quad (4.41)$$

Taking into account the given condition $M_j > -\frac{\aleph}{l_j}$, it can be derived from (4.41) that system states will reach the defined integral sliding surface $S_j = 0$ in finite time.

Then, substituting (4.38) into (4.28) yields

$$\dot{e}_j = -K_j e_j - M_j \text{sign}(S_j) - \tilde{\sigma}_j. \quad (4.42)$$

Obviously, if $K_j > 0$, then $e_j \rightarrow 0$ as $t \rightarrow \infty$, such that the system (4.28) will converge to the desired equilibrium point asymptotically with the proposed DO-ISM method. ■

Based on the Theorem 4.3.3, it is obtained that the DO-ISM method can drive power to track its reference.

Remark 4.3.3 *The reason why the proposed composite controller DO-ISM can suppress the lumped disturbance and chattering is explained below. Generally, if the DO is not employed ($\hat{\sigma}_j = 0$ in (4.38)), the proposed controller degrades into a nominal ISMC controller. To guarantee power tracking stability, the switching gain of the nominal ISMC and DO-ISM have to be designed such that $M_j > 0$ and $M_j > -\frac{\aleph}{l_j}$, ($l_j < 0$), respectively. For the DO-ISM, since the disturbance has been estimated by DO, the magnitude of estimation error $|\tilde{\sigma}_j|$, which is expected to converge to 0, can be kept much smaller than the magnitude of disturbance σ_j . Thus the switching gain of the proposed DO-ISM method can be designed much smaller than the ISMC method, alleviating the chattering problem. In other words, using the same K_j , nominal ISMC without DO would achieve worse robustness than the proposed DO-ISM controller. Hence, enhanced disturbance rejection ability is obtained using the proposed DO-ISM controller.*

4.3.4 Voltage Balancing Loop

An SMO-P controller is designed to balance the dc-link capacitor voltage, avoiding low-frequency voltage ripple of the system presented in the neutral point of the dc-link.

Assumption 4.3.4 *The instantaneous power dynamics are much faster than the dc-link capacitor voltage difference dynamics $\dot{v}_{dc}^o$, i.e., $p \simeq p^*$, and $q \simeq q^*$.*

Based on *Assumptions 4.2.3 and 4.3.4*, the dynamic of the dc-link capacitor voltage difference v_{dc}^o can be expressed as,

$$\dot{v}_{dc}^o = \frac{2p^*}{C\sqrt{6}v_{dc}}u_\gamma + \frac{1}{C}\psi, \quad (4.43)$$

where u_γ is the control signal, ψ is the sinusoidal disturbance [144], as expressed,

$$\psi = \chi \sin(3\omega t + \varphi), \quad (4.44)$$

where amplitude χ and phase φ are the unknown parameters, ω is angular frequency of grid voltage, and t is sampling time.

Introducing two new state variables, disturbances ψ and its time-derivative v_ψ , based on equations (4.43) and (4.44), the augmented system can be formulated as follows:

$$\begin{cases} \dot{v}_{dc}^o = \frac{h}{C}u_\gamma + \frac{1}{C}\psi, \\ \dot{\psi} = v_\psi, \\ \dot{v}_\psi = -9\omega^2\psi, \end{cases} \quad (4.45)$$

where $h = \frac{2p^*}{\sqrt{6}v_{dc}}$. Here, it is worth noting that ψ is viewed as a state of the augmented system. In other words, it is possible to utilize an SMO for estimation purposes.

According to (4.45), the state space can be rebuilt as follows,

$$\begin{cases} \dot{\zeta} = \mathbf{A}\zeta + \mathbf{B}u_\gamma, \\ y = \mathbf{E}\zeta, \end{cases} \quad (4.46)$$

where $\zeta = [v_{dc}^o, \psi, v_\psi]^T$ and y represent system states and measurable output, respectively.

$$\mathbf{A}, \mathbf{B}, \mathbf{E} \text{ are constant matrices, which are shown as: } \mathbf{A} = \left[\begin{array}{c|cc} \underbrace{0}_{\mathbf{A}_{11}} & \underbrace{\frac{1}{C} \ 0}_{\mathbf{A}_{12}} \\ \hline 0 & 0 \ 1 \\ 0 & \underbrace{-9\omega^2 \ 0}_{\mathbf{A}_{22}} \\ \hline \underbrace{}_{\mathbf{A}_{21}} & \end{array} \right], \mathbf{B} =$$

$[\frac{h}{C}, 0, 0]^T$, and $\mathbf{E} = [1, 0, 0]$.

Then the system (4.46) can be reexpressed as,

$$\begin{cases} \begin{bmatrix} \dot{v}_{dc}^o \\ \dot{\zeta}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{A}_{11} & \mathbf{A}_{12} \\ \mathbf{A}_{21} & \mathbf{A}_{22} \end{bmatrix} \begin{bmatrix} v_{dc}^o \\ \zeta_2 \end{bmatrix} + \begin{bmatrix} \mathbf{B}_1 \\ \mathbf{B}_2 \end{bmatrix} u_\gamma, \\ y = v_{dc}^o, \end{cases} \quad (4.47)$$

where $\zeta_2 = [\psi, v_\psi]^T$, $\mathbf{A}_{11} = 0$, $\mathbf{A}_{12} = [\frac{1}{C}, 0]$, $\mathbf{A}_{21} = [0, 0]^T$, $\mathbf{A}_{22} = [0, 1; -9\omega^2, 0]$, $\mathbf{B}_1 = \frac{h}{C}$, $\mathbf{B}_2 = [0, 0]^T$.

4.3.4.1 Sliding Mode Observer (SMO) Design

In order to estimate the state vector ψ for the system (4.47), an SMO can be constructed as,

$$\dot{\hat{\zeta}} = \mathbf{A}\hat{\zeta} + \mathbf{B}u_\gamma + \xi D \text{sign}(e_1), \quad (4.48)$$

where $\hat{\zeta}$ is the estimated state of ζ , $e_1 = v_{dc}^o - \hat{v}_{dc}^o$ is set as the sliding mode variable, D is a positive constant and $\xi \in \mathbb{R}^{3 \times 1}$ is the SMO gain.

Define observed error $e = \zeta - \hat{\zeta}$. The dynamic of e is,

$$\dot{e} = \dot{\zeta} - \dot{\hat{\zeta}}, \quad (4.49)$$

where $\dot{e}$, $\dot{\zeta}$, and $\dot{\hat{\zeta}} \in \mathbb{R}^{3 \times 1}$.

Substituting (4.46) and (4.48) into (4.49), one can get,

$$\dot{e} = \begin{bmatrix} \mathbf{A}_{11} & \mathbf{A}_{12} \\ \mathbf{A}_{21} & \mathbf{A}_{22} \end{bmatrix} e - \xi D \text{sign}(e_1). \quad (4.50)$$

Based on (4.50), set $e = [e_1, e_2]$, $e_2 \in \mathbb{R}^{2 \times 1}$. $\xi = [1, \xi_2]$, $\xi_2 \in \mathbb{R}^{2 \times 1}$. Then, the equation (4.50) can be rewritten,

$$\begin{bmatrix} \dot{e}_1 \\ \dot{e}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{A}_{11} & \mathbf{A}_{12} \\ \mathbf{A}_{21} & \mathbf{A}_{22} \end{bmatrix} \begin{bmatrix} e_1 \\ e_2 \end{bmatrix} - \begin{bmatrix} 1 \\ \xi_2 \end{bmatrix} D \text{sign}(e_1), \quad (4.51)$$

where $e_2 = [\zeta_2 - \hat{\zeta}_2] = \begin{bmatrix} \varphi - \hat{\varphi} \\ v_\varphi - \hat{v}_\varphi \end{bmatrix}$.

Theorem 4.3.4 *Considering the SMO error system (4.51), suppose that $D > \text{Max} \{ |A_{11}e_1 + A_{12}e_2| \}$, it means the state trajectories can reach the predefined sliding mode surface in the finite time driven by equivalent control law $v_{eq} = D \text{sign}(e_1)$, as well as if there exists an SMO reaching gain $[1, \xi_2]$, where $\xi_2 = P^{-1}W$ ($P = P^T > 0$, $P \in \mathbb{R}^{2 \times 2}$, $W \in \mathbb{R}^{2 \times 1}$), satisfying the following LMI of (4.59), where $\Phi = PA_{22} - WA_{12}$ ($\Phi \in \mathbb{R}^{2 \times 2}$), it ensures that the estimate error e_2 asymptotically converges to zero as $t \rightarrow \infty$.*

Proof From (4.51), the dynamic of e_1 can be written as,

$$\dot{e}_1 = A_{11}e_1 + A_{12}e_2 - D \text{sign}(e_1). \quad (4.52)$$

When $e_1 \dot{e}_1 = e_1(A_{11}e_1 + A_{12}e_2) - D|e_1|$, one can select $D > \text{Max} \{ |A_{11}e_1 + A_{12}e_2| \}$ to ensure that $e_1 \dot{e}_1 < 0$. This means the state trajectories can reach the predefined sliding mode surface in the finite time. Hence, it can be obtained that $e_1 = \dot{e}_1 = 0$ and discontinuous control $D \text{sign}(e_1)$ can be replaced with the equivalent continuous control v_{eq} .

(4.52) can be rewritten,

$$\dot{e}_1 = A_{11}e_1 + A_{12}e_2 - v_{eq}, \quad (4.53)$$

since $e_1 = \dot{e}_1 = 0$, then yields $v_{eq} = \mathbf{A}_{12}\mathbf{e}_2$.

Based on $\mathbf{e}_2 = \hat{\boldsymbol{\zeta}}_2 - \boldsymbol{\zeta}_2$, the time-derivative $\dot{\mathbf{e}}_2$ can be obtained,

$$\dot{\mathbf{e}}_2 = \dot{\hat{\boldsymbol{\zeta}}_2} - \dot{\boldsymbol{\zeta}}_2 = \mathbf{A}_{21}\mathbf{e}_1 + \mathbf{A}_{22}\mathbf{e}_2 - \boldsymbol{\xi}_2 D\text{sign}(\mathbf{e}_1). \quad (4.54)$$

Since $v_{eq} = D\text{sign}(\mathbf{e}_1) = \mathbf{A}_{12}\mathbf{e}_2$, then the (4.54) becomes,

$$\dot{\mathbf{e}}_2 = (\mathbf{A}_{22} - \boldsymbol{\xi}_2 \mathbf{A}_{12})\mathbf{e}_2. \quad (4.55)$$

It can be observed that (4.55) can be stabilized in exactly the same way as a typical linear state observer when $\mathbf{A}_{12}$ is viewed as the output matrix. Thus, from this point of view, it can be concluded that the order of the observer dynamics are reduced by one when the system enters the sliding mode. The Lyapunov function is chosen for system (4.55) as follows,

$$V_{e_2} = \mathbf{e}_2^T \mathbf{P} \mathbf{e}_2, \quad (4.56)$$

where $\mathbf{P} = \mathbf{P}^T \in \mathbb{R}^{2 \times 2} > 0$. The time-derivative of (4.56) is,

$$\begin{aligned} \dot{V}_{e_2} &= \dot{\mathbf{e}}_2^T \mathbf{P} \mathbf{e}_2 + \mathbf{e}_2^T \mathbf{P} \dot{\mathbf{e}}_2 \\ &= \mathbf{e}_2^T \left((\mathbf{P} \mathbf{A}_{22})^T - (\mathbf{P} \boldsymbol{\xi}_2 \mathbf{A}_{12})^T \right. \\ &\quad \left. + \mathbf{P} \mathbf{A}_{22} - \mathbf{P} \boldsymbol{\xi}_2 \mathbf{A}_{12} \right) \mathbf{e}_2. \end{aligned} \quad (4.57)$$

Define $\mathbf{W} = \mathbf{P} \boldsymbol{\xi}_2$ ($\mathbf{W} \in \mathbb{R}^{2 \times 1}$), then the switching gain $\boldsymbol{\xi}_2 = \mathbf{P}^{-1} \mathbf{W}$ ($\boldsymbol{\xi}_2 \in \mathbb{R}^{2 \times 1}$). Then (4.57) can be rewritten as,

$$\begin{aligned} \dot{V}_{e_2} &= \mathbf{e}_2^T (\mathbf{A}_{22}^T \mathbf{P} - \mathbf{A}_{12}^T \mathbf{W}^T + \mathbf{P} \mathbf{A}_{22} - \mathbf{W} \mathbf{A}_{12}) \mathbf{e}_2, \\ &= \mathbf{e}_2^T (\boldsymbol{\Phi}^T + \boldsymbol{\Phi}) \mathbf{e}_2. \end{aligned} \quad (4.58)$$

If $\boldsymbol{\Phi}^T + \boldsymbol{\Phi} < 0$, such that the LMI of (4.58) is,

$$\begin{bmatrix} \boldsymbol{\Phi} & 0 \\ 0 & \boldsymbol{\Phi}^T + \boldsymbol{\Phi} \end{bmatrix} < 0 \quad (4.59)$$

where $\boldsymbol{\Phi} = \mathbf{P} \mathbf{A}_{22} - \mathbf{W} \mathbf{A}_{12}$ ($\boldsymbol{\Phi} \in \mathbb{R}^{2 \times 2}$). It can be concluded that if (4.59) is held, then $\dot{V}_{e_2} < 0$, that means $\mathbf{e}_2 \rightarrow 0$ as $t \rightarrow \infty$, that is the system (4.55) is asymptotically stable. Conclusively, $\mathbf{e}_2$ can converge to 0 as $t \rightarrow \infty$, i.e., $\hat{\boldsymbol{\zeta}}_2 \rightarrow \boldsymbol{\zeta}_2$, the proposed SMO can drive $\hat{\psi}$ to estimate ψ . ■

As a result, the sinusoidal disturbance is estimated.

4.3.4.2 SMO-based Proportional (SMO-P) Voltage Balancing Controller Design

Proposition 4.3.2 Consider the system (4.43) and SMO estimation system (4.48) with well defined disturbance-to-output relative degree, then the closed-loop system under the composite controller (that is, SMO-P) designed by (4.60), is semiglobally exponentially stable in the sense that for $K_b > 0$, such that v_{dc}^o of the system (4.43) converges to 0 as $t \rightarrow \infty$.

Proof Based on the $\hat{\psi}$ estimated by SMO, a composite controller SMO-P can be designed as

$$u_\gamma = -\frac{C}{h}(K_b v_{dc}^o + \frac{1}{C}\hat{\psi}). \quad (4.60)$$

Putting the controller (4.60) into (4.45) gets,

$$\dot{v}_{dc}^o = -K_b v_{dc}^o + \frac{1}{C}(\psi - \hat{\psi}). \quad (4.61)$$

It can be observed that when $\hat{\psi} \rightarrow \psi$, the dc-link capacitor voltage difference $v_{dc}^o \rightarrow 0$ as $t \rightarrow \infty$, if $K_b > 0$. ■

The SMO-P control method can drive v_{dc}^o close to zero, demonstrating its ability to balance capacitor voltage.

4.4 AO-P Control Tuning for Voltage Regulation

Based on the Theorem 4.3.1 and Proposition 4.3.1, the AO-P voltage control parameters depend on:

1. The adaptive law parameter $\gamma > 0$ of (4.14) is selected to accommodate the rate of changes in unknown disturbance. A larger γ accelerates the convergence speed of the voltage regulation system; however, excessively large may introduce instability to the system;
2. AO gain $\iota > 0$ of (4.9) is defined to adjust the sensitivity of AO for estimation of unknown system states. Larger ι values result in AO achieving faster convergence rates but may introduce more noise;
3. The parameter γ is typically chosen based on the sampling period for initial adjustment, while ι is tuned according to the magnitude of disturbances. The debugging process involves a trade-off considering the functions of both ι and γ ;
4. The AO-P control gain $K_v > 0$ of (4.20) is chosen to balance the speed and stability of the voltage response. Initially, a smaller initial value K_v is selected, gradually increasing it. Combined with the selected AO gain ι and γ , this process requires continuous tuning until satisfactory voltage regulation performance is achieved.

4.5 DO-ISM Control Tuning for Voltage Regulation

The power tracking parameters of DO-ISM are predicated upon:

1. DO gain l_j of (4.29) governs its convergence speed. When the condition $\|\tilde{\sigma}_j\| \leq \frac{\aleph}{-l_j}$ ($l_j < 0$) is met, Theorem 4.3.2 indicates that ultimately bounded estimation becomes attainable; In essence, a larger l_j enhances the anti-disturbance capability of power tracking. Crucially, it is essential to consider the magnitude of the disturbances;

2. According to Theorem 4.3.3, ensuring that DO-ISMC gains of (4.38) K_j and M_j satisfy the conditions in (4.41) and (4.42), specifically $M_j > -\frac{\kappa}{l_j}$ and $K_j > 0$, enables rapid and accurate power tracking. Increasing K_j enhances power tracking speed, but excessively large K_j may introduce noise. A larger M_j can enhance chattering, whereas a smaller M_j can reduce chattering but may slow down convergence. It is typically recommended to choose M_j based on how strong the nonlinear terms are in the power tracking loop. Generally, M_j is less than K_j ;
3. Due to the precise estimation of DO, the estimation error $\tilde{\sigma}_j$ is significantly smaller than the actual magnitude of disturbances σ_j . Consequently, the gain M_j in the DO-ISMC method can be relatively smaller yet effectively compensate for disturbances, successfully alleviating the chattering. As a result, the DO plays a crucial role in control systems, especially when it comes to addressing disturbances and enhancing system stability;
4. A trial and error for parameters l_j , M_j , and K_j are essential to achieve accurate power tracking performance.

4.6 SMO-P Control Tuning for Balancing Voltage

For the selection of D , ξ , K_b of SMO-P, the design parameter can be summarized as:

1. As a result of this theorem 4.3.4, for the SMO gain parameter D , it is sufficient to ensure that $D > \text{Max} \{|A_{11}e_1 + A_{12}e_2|\}$ to ensure that the estimated state e_1 converges to the actual state within a finite time; Increasing D can expedite reaching the sliding surface but may lead to the chattering if it is too large. An appropriate D should be determined based on the magnitude of the disturbance to be suppressed;
2. The SMO gain ξ ($\xi = [1, \xi_2]$, ξ_2 convergency law of the dynamic $\dot{e}_2$) is obtained by (4.59). When designing ξ_2 for SMO gain using pole placement, better resistance to sinusoidal perturbations can be obtained by placing the eigenvalues in the left half-plane and away from the imaginary axis while reducing the voltage difference in the balanced loop. The choice of ξ_2 too far from the imaginary axis may result in system instability.
3. As per Proposition 4.3.2, the SMO-P control gain K_b enhances the convergence rate of voltage balancing. Hence, a trial and error for parameters K_b , ξ , and D are essential to balance (two) capacitance voltages.

4.7 Experimental Results

This section verifies the effectiveness of the proposed control scheme on the lab-constructed 3L-NPC power converter prototype as well as comparing with the STO-ISMC derived from [150], STA, and PI control schemes, as shown in Figure 4.3. For the sake of providing a fair comparison, all control schemes are tuned to the best performance. The system parameters of the experimental platform are presented in Table 4.2. The proposed, STO-ISMC, STA, and PI control parameters are displayed in Table 4.3.

Table 4.2: Parameters Used for Experiment

Description	Parameter	Value	Unites
Sampling frequency	f_s	10	kHz
Switching frequency	f_{sw}	10	kHz
Grid frequency	f	50	Hz
Grid phase voltage	e_{abc}	50	V
Phase inductor	L	5	mH
Dc-link capacitor	C	1551	uF
Load resistance	R_L	$30 \leftrightarrow 40$	Ω
Reference output voltage	v_{dc}^*	$140 \leftrightarrow 160$	V

Table 4.3: Controller Parameters

Control Scheme	Voltage Regulation Loop	Power Tracking Loop	Voltage Balancing Loop
Proposed Scheme	AO: $\gamma = 5e-4$ $\iota = 4e2$ P: $K_v = 0.3$	DO: $l_p = l_q = -50$ ISMC: $M_p = M_q = 5$ $K_p = K_q = 60$	SMO: $D = 5e2$ $\xi = [1, 0.031, -1.376e3]^T$ P: $K_b = 0.2$
STO-ISMIC Scheme see [150]	STO: $\iota = -1e6$ $\lambda = -3e8$ ISMC: $\varrho = -5e4$ $k_z = 2e2$	STO: $m_p = m_q = -5e3$ $n_p = n_q = -3e5$ ISMC: $\rho_p = \rho_q = 4.5e5$ $k_p = k_q = 1e3$	PI: $k_{p1} = 5e-1$ $k_{i1} = 50$
STA Scheme	$\lambda_z = 2$ $\psi_z = 1e4$	$\lambda_p = \lambda_q = 1e-4$ $\psi_p = \psi_q = 1e-2$	$\lambda_b = 1e-2$ $\psi_b = 0.9$
PI Scheme	$k_{vp} = 0.1$ $k_{bp} = 5e-3$	$k_{pp} = k_{pq} = 2e-5$ $K_{pi} = k_{qi} = 5e-2$	$k_{bp} = 5e-3$ $k_{bi} = 5e-2$

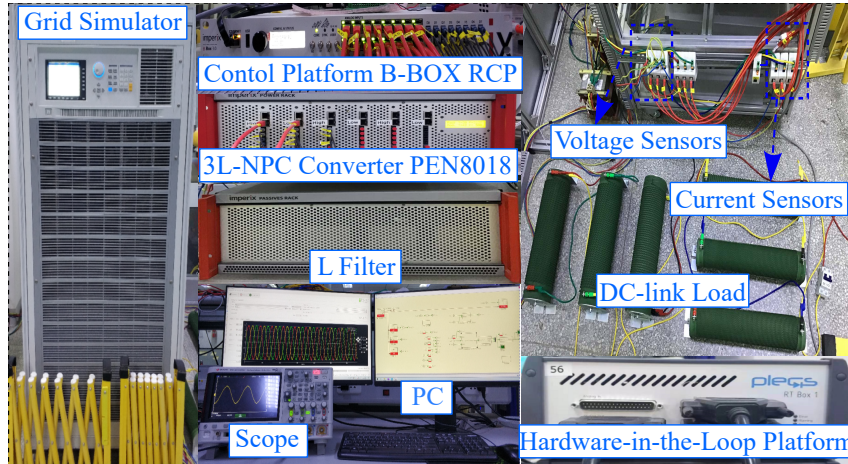


Figure 4.3: The constructed 3L-NPC power converter prototype of Chap. 4.

4.7.1 An Overall Evaluation Comparison

In this case, a comparative analysis among the STO-ISM, STA, and PI control schemes reveals that the proposed control scheme (AO-P, DO-ISM, SMO-P) exhibits superior robustness for the 3L-NPC converter system. The overall assessment and comparison results are illustrated in Figure 4.4. When the dc-link load is 30Ω , the reference dc-link voltage reference v_{dc}^* varies from 160V to 140V at 0.02s, from 140V back to 160V at 0.08s. Simultaneously, the reactive power reference transitions from 0 to 100Var at 0.13s. From all the above, the transient response of the proposed scheme is much better than the STO-ISM, STA, and PI methods. In detail, compared with the STO-ISM, STA, and PI methods, the proposed method presents less over/undershoot, faster response, and stronger robustness in terms of regulating the dc-link capacitor voltage, the tracking power, and the balancing capacitor voltage.

4.7.2 Comparison of Voltage Regulation Performance

Based on a comparison of STO-ISM, STA, and PI voltage control methods with varying dc-load R_L , it is demonstrated that the proposed AO-P control method exhibits superior voltage regulation performance. It is necessary to include both DO-ISM power control and SMO-P voltage balance control within all methods in order to ensure a fair assessment. The proposed AO-P method obtains smaller voltage amplitude variations and faster convergence properties compared with STO-ISM, STA and PI methods as shown in Figure 4.5. Meanwhile, the proposed control method can achieve better static and dynamic performance as shown in Table 4.4. Furthermore, the influence of adaptive gain γ on the performance behavior of voltage regulation is verified. The larger γ can result in faster dynamic responses tested in two cases: (1): the load start-up is shown in Figure 4.6 (a), and (2): the voltage start-up is shown in Figure 4.6 (b). It is worth noting that a too-large γ may induce voltage overshoot and even lead to an unstable operation. Consequently, a trade-off between γ and the voltage performance should be made.

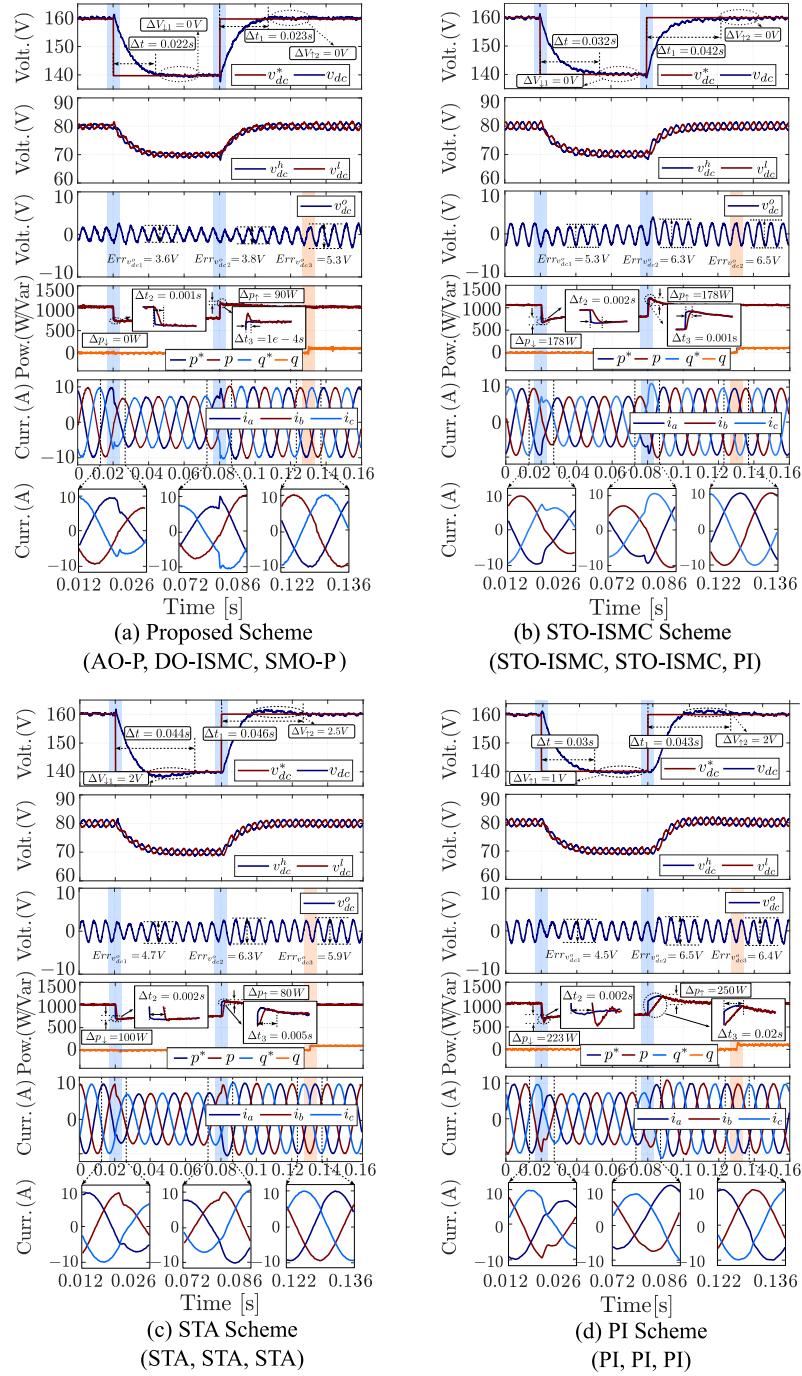


Figure 4.4: An overall evaluation comparison when the capacitor voltage reference v_{dc}^* ($160\text{V} \rightarrow 140\text{V} \rightarrow 160\text{V}$), and the q^* ($0 \rightarrow 100 \text{ Var}$) with $R_L = 30\Omega$.

4.7.3 Comparison of Power Tracking Performance

The effectiveness of the DO-ISM method is demonstrated through a comparison with the STO-ISM, STA, and PI power control methods, as illustrated in Figure 4.7. The results indicate superior performance of the DO-ISM method in power tracking, especially when facing various disturbances such as parameter mismatches, unmodeled dynamics, measurement noise, and bias

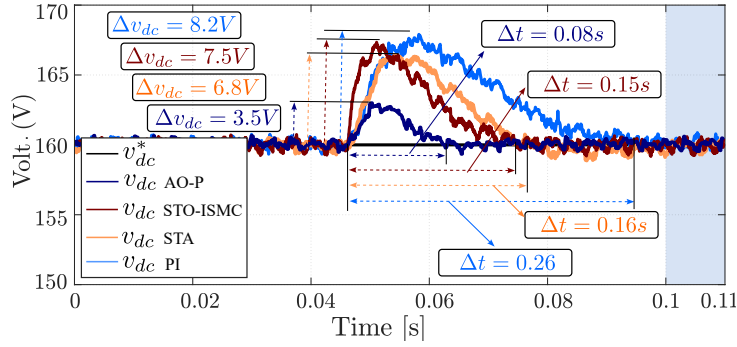


Figure 4.5: Voltage regulation comparisons are conducted under varying dc-load R_L ($30\Omega \rightarrow 40\Omega$) with $v_{dc}^* = 160\text{V}$ and $q^* = 0\text{Var}$.

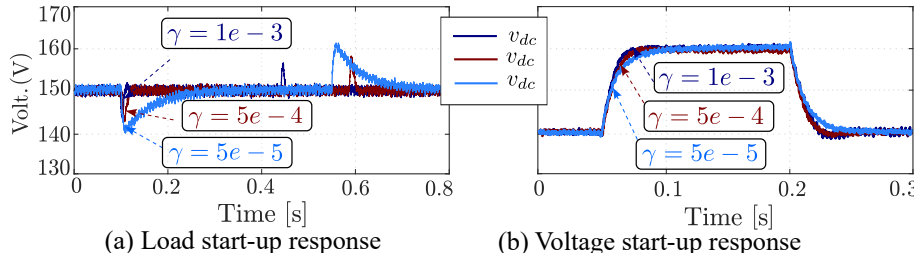


Figure 4.6: Dynamic response voltage v_{dc} with different AO gain γ : (a) when R_L is $40\Omega \rightarrow 30\Omega$ with $v_{dc}^* = 150\text{V}$, $q^* = 0\text{Var}$; and (b) when v_{dc}^* is $140\text{V} \rightarrow 160\text{V}$ with $R_L = 30\Omega$ and $q^* = 0\text{Var}$.

caused by current path changes and switch mode noise. All the control methods above employ P-AO in the voltage regulation loop and SMO-P in the voltage balance loop. As shown in Figure 4.7, the reactive power tracking presents a faster dynamic response using the proposed DO-ISM. From the perspective of current quality, the THD of i_a obtained by the proposed DO-ISM is 2.6693%, which is smaller than that of the STO-ISM (2.8398%), STA (3.0451%), and PI (3.2545%) methods. The performance indexes of reactive power are shown in Table 4.5. It can be seen that the proposed DO-ISM method exhibits better performance than the STO-ISM, STA, and PI methods.

4.7.4 Comparison of Voltage Balancing Performance

To examine voltage balanced performance, the comparison experiment results among the proposed SMO-P, STA, and PI methods are shown in Figure 4.8. The DO-ISM and AO-P control approaches are applied to the power tracking loop and voltage regulation loop, respectively. The SMO-P method presents smaller voltage difference amplitudes than the STA and PI methods. The proposed control SMO-P can yield better static and dynamic performance than the STA and PI methods as shown in Table 4.6. To further analyze the influence of the observer gain on the closed-loop system, the pole assignment method is adopted to choose SMO gain $\xi = [1, \xi_2]^T$. Three sets ξ_2 for SMO gains are obtained, i.e., $\xi_2' = [0.248, -1.368e3]^T$, $\xi_2'' = [-0.002, -1.378e3]^T$, and $\xi_2''' = [-0.155, -1.374e3]^T$. It should be noted that the ξ_2' is obtained by the minimum eigenvalue that is furthest away from the imaginary axis, ξ_2'' is the second and ξ_2''' is the closest. Observably, from Figure 4.9, the difference between the capacitor voltages is the smallest by

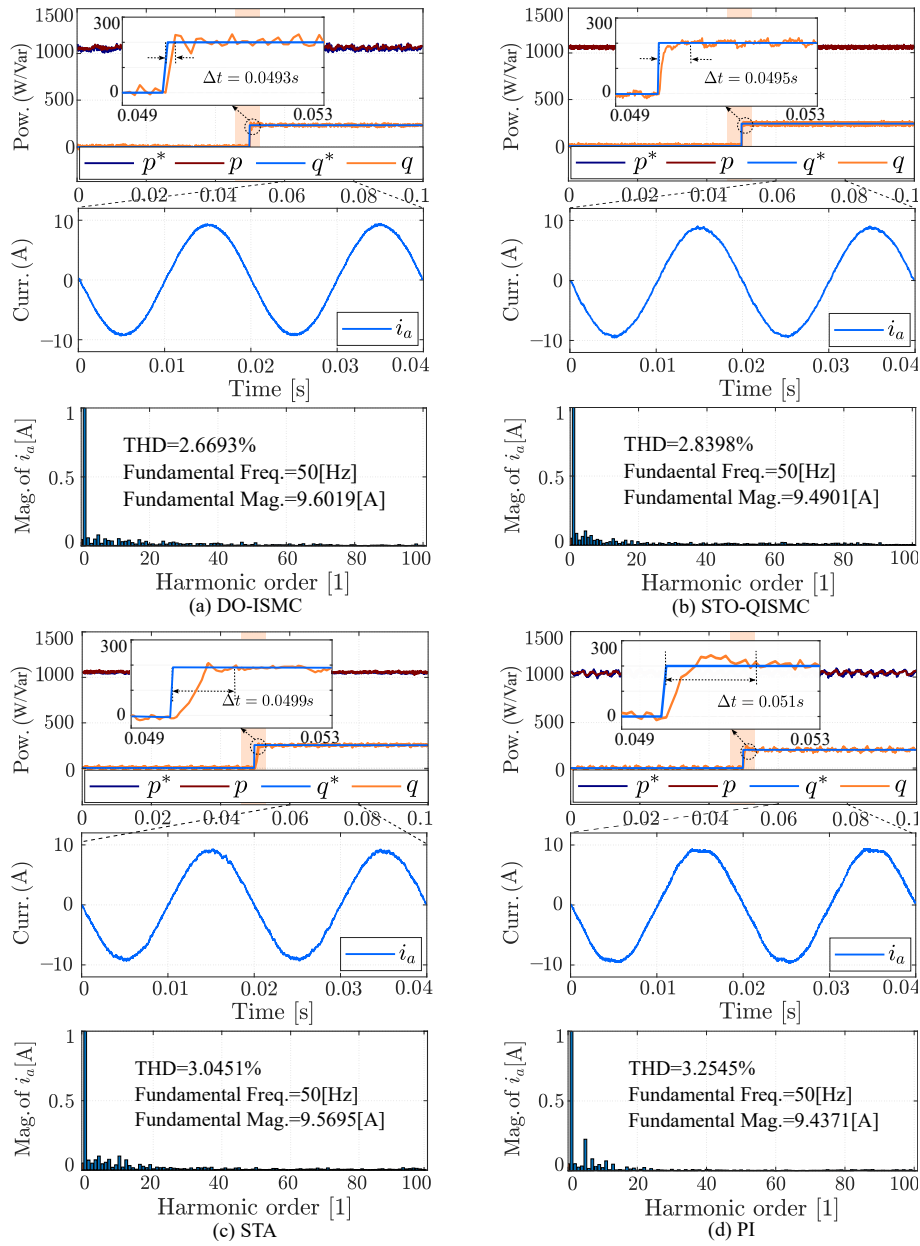


Figure 4.7: Comparisons are made for power tracking, varying q^* from 0Var to 200Var, with fixed values of $v_{dc}^* = 160V$ and $R_L = 30\Omega$.

using SMO gain ξ_2' , ξ_2'' is the second, and ξ_2''' is the largest. When designing the observer gain ξ_2 using pole placement method, if the eigenvalues are located in the left half-plane and are farther away from the imaginary axis, the resulting gain used in the balancing loop of the system can ensure a smaller magnitude of the voltage difference. As the over-lager SMO gain may be sensitive to measurement noise and may lead to unstable operation, a better voltage balancing performance requires trial-and-error.

Table 4.4: Comparisons of AO-P, STO-ISM, STA and PI Methods

Voltage Regulation Index (Figure 4.5: 0.1s-0.11s)		AO-P Method	STO-ISM Method	STA Method	PI Method
Static Index v_{dc} (V)	OS ¹	3.5	7.5	6.8	8.2
	OST ²	0.08	0.15	0.16	0.26
	ME ³	0.0224	0.0232	0.0428	0.0737
	MSE ⁴	0.3049	0.3101	0.3218	0.3408
	VE ⁵	0.0808	0.0916	0.0828	0.0884
Static Index v_{dc} (V)	AO-P vs STA	AO-P vs STO-ISM		AO-P vs PI	
	Enhancement (%)				
OS ¹	53.3%	48.5%		57.3%	
OST ²	46.7%	50%		69.2%	
ME ³	3.45%	47.7%		69.6%	
MSE ⁴	1.68%	5.25%		10.5%	
VE ⁵	11.8%	2.42%		8.60%	

1: Over-shooting (V); 2: Over-settling Time (S); 3: Mean Error; 4: Mean Square Error; 5: Variance Error.

Table 4.5: Comparisons of DO-ISM, STO-ISM and PI Methods

Powing Tracking Index (Figure 4.7: 0.06s-0.08s)		DO-ISM Method	STO-ISM Method	STA Method	PI Method
Static Index q (Var)	OST ²	0.0493	0.0495	0.0499	0.051
	ME ³	0.0662	0.3709	0.4202	0.2717
	MSE ⁴	10.9050	11.6411	15.3726	13.5668
	VE ⁵	120.1044	128.9034	146.9148	185.8247
	THD	2.6693%	2.8398%	3.0451%	3.2545%
Static Index q (Var)	DO-ISM vs STO-ISM	DO-ISM vs STA		DO-ISM vs PI	
	Enhancement (%)				
OST ²	0.4%	1.2%		3.3%	
ME ³	82.2%	84.2%		75.6%	
MSE ⁴	6.32%	29.1%		19.6%	
VE ⁵	6.83%	18.2%		35.4%	
THD	6.0%	12.3%		18%	

2: Over-settling Time (S); 3: Mean Error; 4: Mean Square Error; 5: Variance Error.

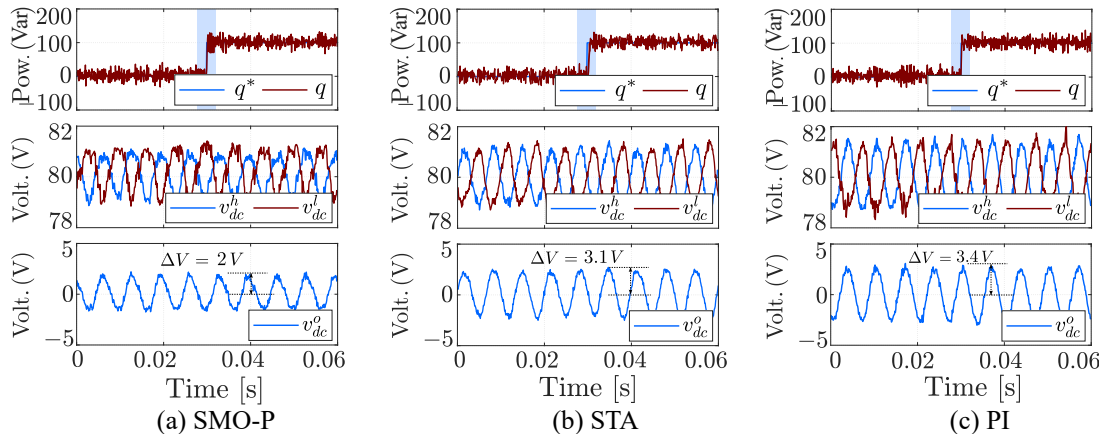


Figure 4.8: Comparisons of voltage balancing with q^* (0 Var $\rightarrow$ 100Var), $v_{dc}^* = 160\text{V}$, and $R_L = 30\Omega$.

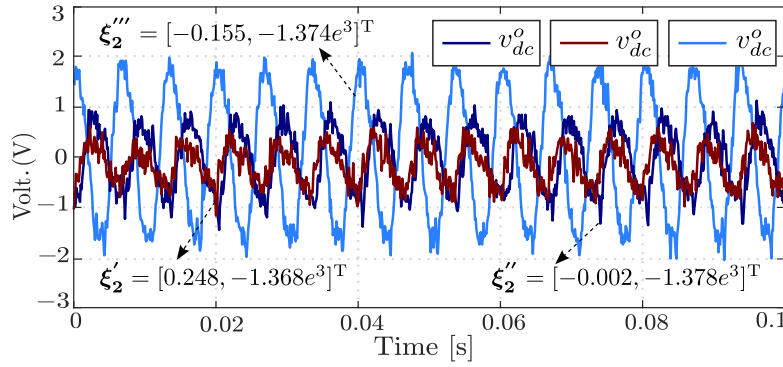


Figure 4.9: Dynamic response of v_{dc}^o with different ξ_2 for SMO gains ξ .

4.7.5 Robustness Evaluation

To assess the robustness of proposed control scheme (AO-P, DO-ISMC, SMO-P) for the 3L-NPC converter, experimental tests are divided into two groups: 1) evaluating external disturbance capacity, applied to the voltage regulation loop (with power and voltage balance loops employing DO-ISMC and SMO-P control methods, respectively), and 2) working conditions with inductance mismatches, applied to the power loop (with voltage regulation and voltage balance loops utilizing AO-P and SMO-P control methods, respectively).

4.7.5.1 External Disturbance DC-load Variations

A test bench is carried out to validate the robustness against external disturbance variations of the proposed control scheme compared with the STO-ISMC, STA and PI methods as displayed in Figure 4.10. It is observed that the output voltage, active power, and grid-side current have less voltage over/undershoot, faster dynamic response than the STO-ISMC, STA, and PI methods, respectively. This means the proposed control scheme is more robust than the STO-ISMC, STA, and PI methods against dc-link load variations.

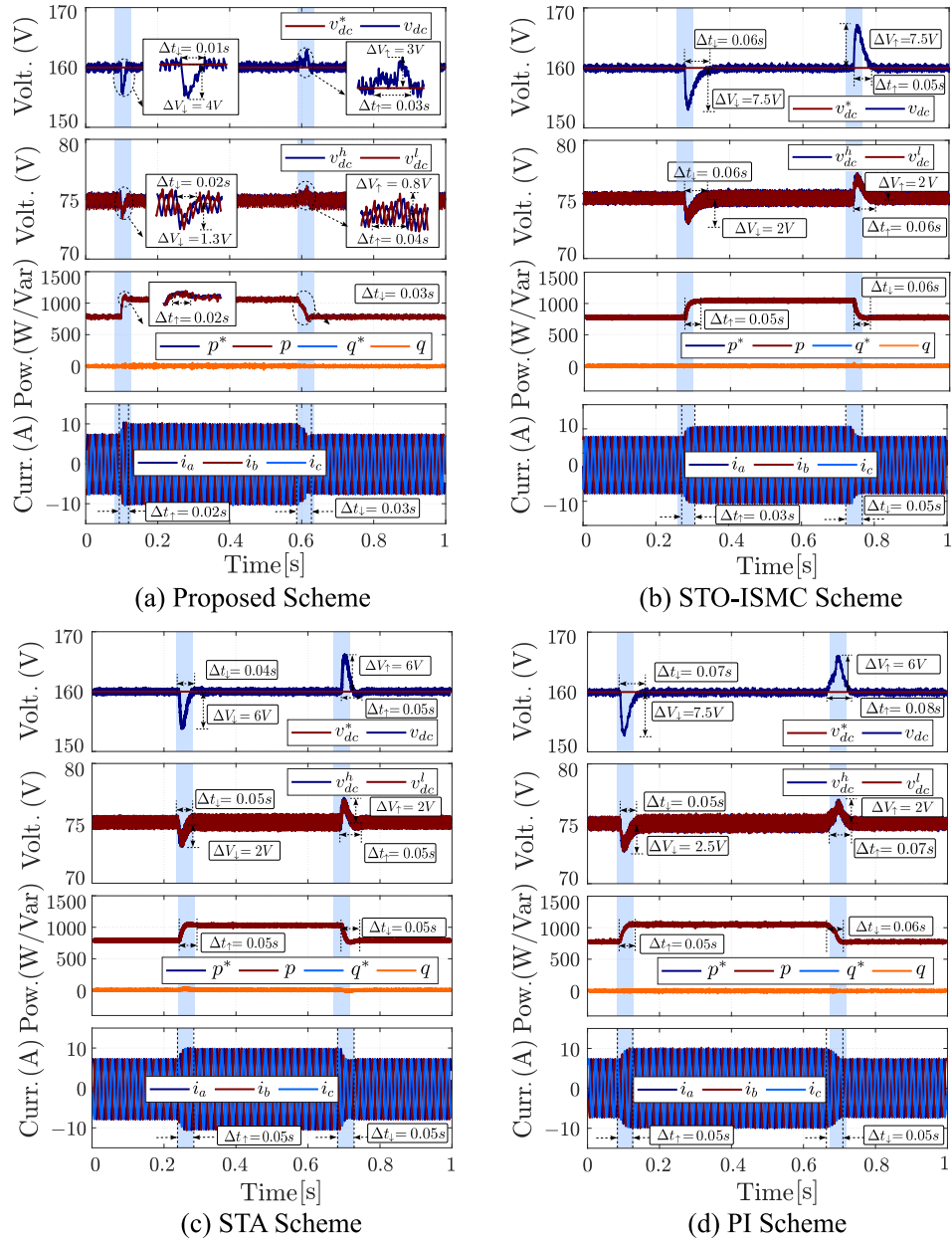


Figure 4.10: Comparisons of steady-state voltage regulation performance: R_L varies from 40Ω to 30Ω , and again to 40Ω with $v_{dc}^* = 160 \text{ V}$ and $q^* = 0 \text{ Var}$.

4.7.5.2 Performance Comparison of Inductance Mismatches

To further validate the impact of the proposed, STO-ISMV, STA, and PI control schemes, on the robustness of 3L-NPC power converter in mismatched inductor parameters, the inductance L is varied as is varied as $100\%L$, $50\%L$, and $150\%L$ of its nominal value. The comparison is illustrated in Figure 4.11. The control strategies are performed on a real-time hardware-in-the-loop due to mechanical limitations of the real plant. When the grid-side filter inductance is set to its nominal value, all the control schemes perform well and operate normally, as depicted in Figs. 4.11 (b), (e), (h), and (i). When the inductance value L is set to $150\%L$ of the nominal inductance,

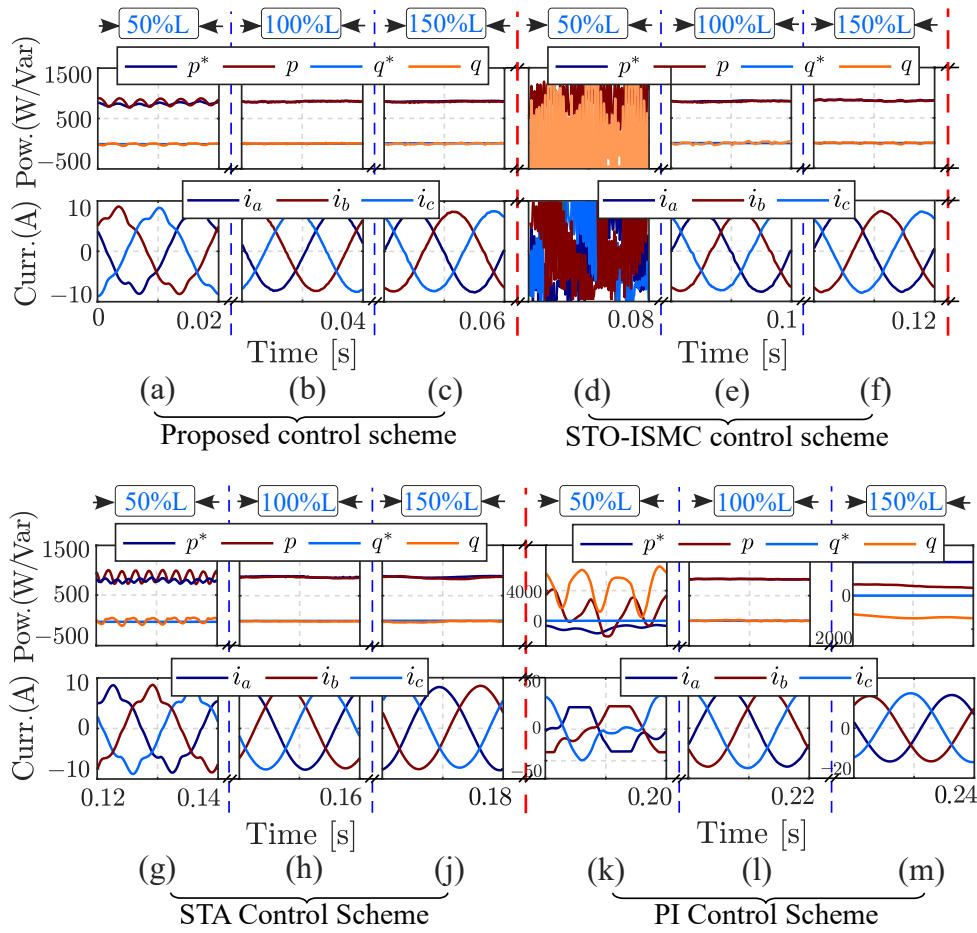


Figure 4.11: The filter inductance parameter mismatches: (a) work properly, (b) work properly, (c) work properly, (d) work improperly, (e) work properly, (f) work properly, (g) work grudgingly, (h) work properly, (j) work properly, (k) work improperly, (l) work properly, and (m) work improperly.

the proposed, STO-ISM, STA control schemes still function properly, as shown in Figs. 4.11 (c), (f), and (j). When the inductance value L is set to $50\%L$ of the nominal inductance, the proposed methods work properly, as illustrated in Figs. 4.11 (a), the STO-ISM work improperly, STA methods work grudgingly, as illustrated in Figs. 4.11 (d), and (g). Contrastingly, the PI control scheme operates ineffectively at both $150\%L$ and $50\%L$, as depicted in Figs. 4.11 (m) and (k). In summary, the proposed control scheme demonstrates superior robustness compared to the STO-ISM, STA and PI control schemes in the presence of inductance parameter mismatch.

4.8 Summary

In this chapter, a new effective robust adaptive sliding mode control scheme, including three composite controllers, was developed to improve the control performance of the 3L-NPC power converter in microgrids against multi-disturbances. The AO-P controller was applied to regulate the dc-link voltage in the presence of the unknown dc-link equivalent loads. The DO-ISM was

Table 4.6: Comparisons of the SMO-P, STA and PI Methods

Voltage Balancing Index (Figure 4.8: 0.04s-0.06s)		SMO-P Method	STA Method	PI Method
Static Index v_{dc}^o (V)	CVI ¹	2	3.1	3.4
	ME ²	0.0787	0.1368	0.2338
	MSE ³	0.9991	1.3883	1.9109
	VE ⁴	1.0020	1.9321	3.6237
Static Index v_{dc}^o (V)	SMO-P vs STA		SMO-P vs PI	
	Enhancement (%)			
CVI ¹		35.5%		41.2%
ME ²		42.4%		66.3%
MES ³		28.0%		47.7%
VE ⁴		48.1%		72.3%

1: Capacitor Voltage Imbalance; 2: Mean Square Error; 3: Variance Error; 4: Variance Error.

implemented to track the power references in the face of lumped disturbances caused by system parameter mismatches, unmodeled dynamics, measurement noise, and bias caused by current path changes and switch mode noise. The SMO-P was designed to balance (*two*) capacitors, avoiding ac-voltage low-frequency oscillation. Experimental results confirmed the proposed control scheme outperformed the STO-ISM, STA and PI control schemes in terms of dynamic response, THDs, and anti-multi-disturbances.

CHAPTER 5

Super-Twisting Current-Voltage Control for 3L-NPC Power Converters

3L-NPC power converters are necessary interfaces to form microgrids. Naturally, designing a suitable control scheme featuring superior dynamics, strong robustness, and simple structure is a promising solution to guarantee more efficient operation of the converter. The schematic diagram of the power converter in microgrids is depicted in Figure 5.1. This chapter proposes a robust high-quality current control strategy for the 3L-NPC power converter in the stationary $\alpha\beta$ frame. A LO coupled with an STA current controller is proposed to deal with the poor sinusoidal current tracking issue due to the existing inductance/grid frequency deviations and the disturbance of the sinusoidal dynamic nature. Additionally, an ESMDO-based P control is built to dramatically enhance voltage regulation performance in the case of capacitance deviations and unknown dc-loads. Experimental data confirm the effectiveness of the proposed solution outperforms the conventional PR and PI control in terms of accurate tracking current/voltage, anti-disturbance, and grid current total harmonic distortion.

5.1 Introduction

Broadly speaking, control objectives of the underlying 3L-NPC power converter in microgrids include (at least): (i) simple control structures; (ii) high-quality current tracking performance; (iii) fast and accurate voltage regulation performance; (iv) strong robustness against parameter deviations and disturbances. Grid current of power converter are usually controlled in the dq synchronously rotating reference frame based the phase locked loop (PLL). However, the performance of PLL presents problems facing of distorted and imbalanced grid voltages [151–153]. Aiming at issues, current control in the $\alpha\beta$ synchronous stationary coordinate system for the 3L-NPC power converter can avoid using PLL to simple control structure as well as eliminate coupling-current items. Additionally, when applying the cascade PR controller [154, 155] to cope with the currents in the $\alpha\beta$ synchronous stationary, a low current control performance is

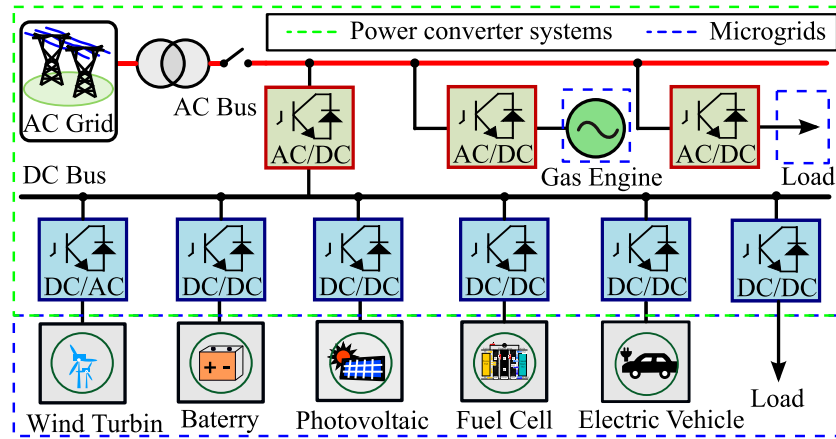


Figure 5.1: The schematic diagram of the power converter in MESs.

inevitably seen with disturbances/uncertainties, such as power interruption, voltage sag/swell and harmonic distortion, parameter deviations.

SMC is more effective in achieving the desired dynamic response and robustness for power converters, due to its fast dynamic response, disturbance rejection, and simple implementation [58–60]. A DSMC law is implemented to track the converter side current for grid-connected voltage source inverters with LCL output filters in [156]. In [157], a full-order terminal SMC is proposed to track errors of the currents to converge their equilibrium point within finite time. However, the above methods suffer from the chattering in the control input referred which is not preferred in practice. To improve the chattering problem, a novel active-disturbance-rejection-based sliding-mode current control scheme for PMSM drives is proposed in [158]. Moreover, the boundary layer approach is also a good solution to suppress the chattering. An adaptive fuzzy SMC based on the boundary layer approach is proposed for motor drive in [159]. However, the boundary layer approach results in a trade-off between control performance and chattering elimination. Hence, at least for now, HOSMC is effective way to mitigate the chattering by integrating to turn discontinuous control inputs into continuous ones. Like [120], a new nonlinear current control strategy is proposed based on back-stepping control and high-order sliding mode differentiator to actively compensate harmonics/inter-harmonics of local loads for distributed generation unit interfacing converters. Admittedly, HOSMC (generally equal or greater than three orders) is inevitably affected by the use of higher-order time derivatives of sliding variables, thus a better alternative is the second-order sliding mode method.

STA is one of the SOSMC methods introduced to suppress the chattering effects [73, 150]. Different from the conventional first-order sliding mode control, STA has a finite-time convergence with continuous control input as well as strong robustness under high uncertainties. In addition, STA can directly deal with a time-varying set-point well, even in the presence of disturbances. A very large disturbance may result in system instability or require higher STA controller gains that lead to system chattering. Hence, it is necessary to design an observer to estimate disturbance, as feedforward compensation in control input, to alleviate the system chattering.

On the other hand, it is crucial to ensure a constant dc-link voltage for the 3L-NPC power converter control system in industrial applications. In the ideal setting, the constant dc-link voltage can be directly controlled by PI controller. However, the dynamic performance is not

satisfactory in the presence of the unknown external disturbances, measurement noises, and non-constant power generated by the microgrids. To address these issues, numerous observer-based compensation approaches have been reported [160–163]. In [93], a robust controller based on ESO is proposed for a dc-dc interleaved boost converter, in which ESO is utilized to estimate the converter uncertainties and external disturbances. A finite-time disturbance observer, providing accurate estimations of the lumped disturbances, is introduced into the controller design to achieve better voltage tracking performance in [164]. An ESMDO is designed to estimate the total disturbances of the system to enhance the speed control performance of the permanent magnet synchronous motor in [165]. Different from the above observers, the ESMDO not merely has a good disturbance estimation ability but also has an inbuilt low-pass filter without introducing a phased delay in [102].

Motivated by the above observation, this work proposes a novel robust high-quality current control with fast convergence for the 3L-NPC power converter of microgrids in $\alpha\beta\gamma$ frame. Contributions of this work include:

- 1) The proposed current control scheme does not require any PLL, and can omit current decoupling processes. As a result, it has low computational complexity and is easy to implement.
- 2) A LO coupled with an STA current controller is proposed to enhance sinusoidal current tracking performance, resulting in robust properties against grid frequency/inductance deviations. Additionally, the LO is designed to offset the disturbance of the sinusoidal dynamic nature, alleviating the system chattering. The reason is that the STA switching gain is simply selected to be bigger than the bound of the sinusoidal disturbance estimation error and parameter deviations rather than that of the lumped of the disturbance and deviations.
- 3) The ESMDO based P control is used in the voltage regulator to enhance dramatically its convergence speed and anti-disturbance capability, achieving fast and accurate voltage tracking. The ESMDO is designed by introducing the sliding mode reaching law to estimate the system uncertainties and unknown dc-link load, in which the observer estimation is considered as a feed-forward compensation term into the voltage regulator.

5.2 System Description

In this section, the underlying 3L-NPC power converter system will be described and fully modelled.

5.2.1 Converter Description and Modeling

The topology structure of the 3L-NPC power converter is shown in Figure 5.2a, as one of the commonly used converters in power electronics. Certainly, the dc-link can easily be extended to other application scenarios, like photovoltaics, wind turbines and electric vehicles, etc. In this work, to simplify the following modeling and analysis, the dc-link is connected to a resistance load. The nomenclature list of the 3L-NPC power converter is presented in Table 5.1.

Based on the 3L-NPC power converter topology [19], it can be obtained in stationary $\alpha\beta\gamma$



Table 5.1: Nomenclatures

Symbol	Description
e_{abc}	$[e_a, e_b, e_c]^T$, grid voltage vector in abc frame.
i_{abc}	$[i_a, i_b, i_c]^T$, grid current vector in abc frame.
$v_{\alpha\beta}$	$[v_\alpha, v_\beta]^T$, grid voltage vector in $\alpha\beta$ -axis.
$i_{\alpha\beta}$	$[i_\alpha, i_\beta]^T$, grid current vector in $\alpha\beta$ -axis.
$u_{\alpha\beta}$	$[u_\alpha, u_\beta]^T$, average duty cycles in $\alpha\beta$ -axis.
VT_l^n	$n = \{a, b, c\}, l = \{1, 2, 3, 4\}$, insulated-gate bipolar transistors.
VD_l^n	$n = \{a, b, c\}, l = \{1, 2\}$, clamping diodes.
v_{dc}	$v_{dc}^h + v_{dc}^l$, the sum of dc-link capacitor voltages.
v_{dc}^o	$v_{dc}^h - v_{dc}^l$, the difference of dc-link capacitor voltages.
R_L, i_L	dc-load resistance, dc-link load current.
L, C, ω	linear inductor, dc-link capacitor, grid voltage frequency.

coordinates,

$$\dot{i}_{\alpha\beta} = \frac{v_{\alpha\beta}}{L} - \frac{1}{2L}v_{dc}u_{\alpha\beta} + \frac{Y}{L}, \quad (5.1)$$

$$\dot{v}_{dc} = \frac{u_{\alpha\beta}^T i_{\alpha\beta}}{C} - \frac{2v_{dc}}{CR_L}, \quad (5.2)$$

$$\dot{v}_{dc}^o = \frac{2}{\sqrt{6}C}Z + \frac{1}{\sqrt{6}C}u_{\alpha\beta}^T i_{\alpha\beta} u_\gamma, \quad (5.3)$$

where $Y = \begin{bmatrix} \frac{1}{2\sqrt{6}}(u_\beta^2 - u_\alpha^2) - \frac{1}{\sqrt{3}}u_\alpha u_\gamma \\ \frac{u_\alpha u_\beta}{\sqrt{6}} - \frac{u_\beta u_\gamma}{\sqrt{3}} \end{bmatrix} v_{dc}^o$, $Z = \begin{bmatrix} u_\alpha^2 - u_\beta^2, & -u_\alpha u_\beta \end{bmatrix} i_{\alpha\beta}$, and $u_{\alpha\beta}, u_\gamma$ are control signals.

In this research, three assumptions have been imposed on the 3L-NPC power converter system (5.1)-(5.3) to operate correctly:

- a1: the source voltages are assumed to be symmetrical and balanced.
- a2: the averaged duty cycles $u_{\alpha\beta}$ are defined as values in $[-1, 1]$ range so mentioned Y of (5.1) is two order of magnitude lower than the second term of (5.1).
- a3: v_{dc}^o of equation (5.3) is lower by more than v_{dc} one order of magnitude, due to v_{dc}^o must be near 0.

5.3 Control Strategy Design

In this section, a cascaded control scheme of the 3L-NPC power converter is shown in Figure 5.2b.

5.3.1 Control Objectives

The control objectives for the 3L-NPC power converter system are to achieve current tracking, voltage regulation, and capacitor voltage balancing.

1. In the voltage regulation loop, the dc-link voltage is regulated to track its reference value (i.e., $v_{dc} \rightarrow v_{dc}^*$), forced by the active power reference p^* , which serves as the input for the current tracking loop (see Fig. 5.2-b.①).
2. In the current tracking loop, currents $i_{\alpha\beta}$ are utilized to track their references (i.e., $i_{\alpha\beta} \rightarrow i_{\alpha\beta}^*$), where $i_{\alpha\beta}^*$ is derived from the voltage regulation loop. The signals $u_{\alpha\beta}$ is this loop outputs, while q^* is given to adjust the power factor (see Fig. 5.2-b.②).
3. In the voltage balancing loop, the objective is to minimize the voltage difference across (two) capacitors (i.e., $v_{dc}^o \rightarrow 0$). The outputs from this loop and the current tracking loop make the final average duty cycle, governing the power converter (Fig. 5.2-b.③).

5.3.2 Current Tracking Loop

In this section, a STA combined with a LO current control is proposed to drive $i_{\alpha\beta} \rightarrow i_{\alpha\beta}^*$, enhancing dramatically the dynamic performance and anti-disturbance capability of current tracking in the case of parameter deviations and uncertainties.

With the consideration of (5.1), the inductance current dynamics can be approximated as follows,

$$\dot{i}_{\alpha\beta} = \frac{v_{\alpha\beta}}{L} - \frac{v_{dc}}{2L} u_{\alpha\beta}, \quad (5.4)$$

Define the current tracking error,

$$\tilde{i}_{\alpha\beta} = i_{\alpha\beta} - i_{\alpha\beta}^*. \quad (5.5)$$

Take the derivative in (5.5) as follows,

$$\dot{\tilde{i}}_{\alpha\beta} = \dot{i}_{\alpha\beta} - \dot{i}_{\alpha\beta}^*. \quad (5.6)$$

Substituting (5.4) into (5.6), it can be obtained,

$$\dot{\tilde{i}}_{\alpha\beta} = -\frac{v_{dc}}{2L} u_{\alpha\beta} + \frac{v_{\alpha\beta}}{L} + \bar{d}_{\alpha\beta} - \dot{i}_{\alpha\beta}^*, \quad (5.7)$$

where $\bar{d}_{\alpha\beta}$ represents parameter deviations and measurement noises of the system (5.7), and $\dot{i}_{\alpha\beta}^*$ is a sinusoidal waveform treated as an external disturbance. Hence, a LO is design to offset it.

Considering the current nature $i_{\alpha\beta}^*$, one can get,

$$\begin{aligned} i_{\alpha\beta}^* &= m_{\alpha\beta} \sin(\omega t + \theta), \\ \dot{i}_{\alpha\beta}^* &= \omega m_{\alpha\beta} \cos(\omega t + \theta), \\ \ddot{i}_{\alpha\beta}^* &= -\omega^2 m_{\alpha\beta} \sin(\omega t + \theta), \end{aligned} \quad (5.8)$$

where $\mathbf{m}_{\alpha\beta} = [m_\alpha, m_\beta]^T$ is current amplitude of $i_{\alpha\beta}^*$, ω is the grid angular frequency, and θ is initial phase angles of the current.

The dynamic of (5.8) can be obtained as

$$\begin{cases} \dot{\xi}_{\alpha\beta} = A\xi_{\alpha\beta}, \\ \mathbf{y}_{\alpha\beta} = C\xi_{\alpha\beta}, \end{cases} \quad (5.9)$$

where $\xi_{\alpha\beta} = [i_{\alpha\beta}^*, \dot{i}_{\alpha\beta}^*]^T$, $A = \begin{bmatrix} 0 & 1 \\ -\omega^2 & 0 \end{bmatrix}$, $C = [1, 0]$ and $\mathbf{y}_{\alpha\beta}$ is control output.

5.3.2.1 Luenberger Observer (LO) Design

According to (5.9), the expressions of the LO dynamic can be gotten as,

$$\begin{cases} \dot{\hat{\xi}}_{\alpha\beta} = A\hat{\xi}_{\alpha\beta} + \mathcal{L}(\mathbf{y}_{\alpha\beta} - \hat{\mathbf{y}}_{\alpha\beta}), \\ \hat{\mathbf{y}}_{\alpha\beta} = C\hat{\xi}_{\alpha\beta}, \end{cases} \quad (5.10)$$

where $\hat{\xi}_{\alpha\beta} = [\hat{i}_{\alpha\beta}^*, \hat{\dot{i}}_{\alpha\beta}^*]^T$, and $\hat{\mathbf{y}}_{\alpha\beta}$ is control output of the LO.

Define the LO error is $\tilde{\xi}_{\alpha\beta} = \xi_{\alpha\beta} - \hat{\xi}_{\alpha\beta}$, then error dynamic of the LO is as follows:

$$\dot{\tilde{\xi}}_{\alpha\beta} = \dot{\xi}_{\alpha\beta} - \dot{\hat{\xi}}_{\alpha\beta} = (A - \mathcal{L}C)\tilde{\xi}_{\alpha\beta}, \quad (5.11)$$

where $\mathcal{L}$ is the LO observer gain. The stability of the LO can be guaranteed that $(A - \mathcal{L}C)$ is Hurwitz matrix, i.e., every eigenvalue has strictly negative real part. Therefore, it can be concluded that $\tilde{\xi}_{\alpha\beta} \rightarrow \xi_{\alpha\beta}$, i.e., $\dot{i}_{\alpha\beta}^*$ can be estimated as $\hat{\dot{i}}_{\alpha\beta}$.

5.3.2.2 LO-based Super-Twisting Algorithm (LO-STA) Control Design

Based on (5.7), the LO-STA control is designed as:

$$\mathbf{u}_{\alpha\beta} = \frac{2L}{v_{dc}}(\delta_{\alpha\beta} + \frac{1}{L}\mathbf{v}_{\alpha\beta} - \hat{\mathbf{d}}_{\alpha\beta}), \quad (5.12)$$

with

$$\delta_{\alpha\beta} = \lambda_{\alpha\beta}|\tilde{i}_{\alpha\beta}|^{\frac{1}{2}}\text{sign}(\tilde{i}_{\alpha\beta}) + \varphi_{\alpha\beta} \int_{t_0}^t \text{sign}(\tilde{i}_{\alpha\beta})ds, \quad (5.13)$$

where $\delta_{\alpha\beta} = [\delta_\alpha, \delta_\beta]^T$ is considered as control signal of the LO-STA, $\lambda_{\alpha\beta} = [\lambda_\alpha, \lambda_\beta]^T > 0$, $\varphi_{\alpha\beta} = [\varphi_\alpha, \varphi_\beta]^T > 0$.

Substituting (5.12) in (5.7), it can be obtained that

$$\dot{\tilde{i}}_{\alpha\beta} = -\delta_{\alpha\beta} + \bar{\mathbf{d}}_{\alpha\beta} + \hat{\mathbf{d}}_{\alpha\beta} - \dot{i}_{\alpha\beta}^*. \quad (5.14)$$

Define observer error $\tilde{D}_{\alpha\beta} = \hat{\mathbf{d}}_{\alpha\beta} - \dot{i}_{\alpha\beta}^*$, and (5.14) can be rewritten,

$$\dot{\tilde{i}}_{\alpha\beta} = -\delta_{\alpha\beta} + \bar{\mathbf{d}}_{\alpha\beta} + \tilde{D}_{\alpha\beta}. \quad (5.15)$$

Define $\tilde{\mathbf{d}}_{\alpha\beta} = \bar{\mathbf{d}}_{\alpha\beta} + \tilde{\mathbf{D}}_{\alpha\beta}$, then (5.15) can be reexpressed as

$$\dot{\tilde{\mathbf{i}}}_{\alpha\beta} = -\delta_{\alpha\beta} + \tilde{\mathbf{d}}_{\alpha\beta}. \quad (5.16)$$

Using (5.13), the error dynamic (5.16) can be rewritten as:

$$\begin{aligned} \dot{\tilde{\mathbf{i}}}_{\alpha\beta} &= -\lambda_{\alpha\beta} |\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}} \text{sign}(\tilde{\mathbf{i}}_{\alpha\beta}) + \phi_{\alpha\beta}, \\ \dot{\phi}_{\alpha\beta} &= -\varphi_{\alpha\beta} \text{sign}(\tilde{\mathbf{i}}_{\alpha\beta}) + \tilde{\mathbf{d}}_{\alpha\beta}. \end{aligned} \quad (5.17)$$

Theorem 5.3.1 *Considering the current systems (5.17), supposed that $\dot{\tilde{\mathbf{d}}}_{\alpha\beta}$ is uniformly bounded perturbations, i.e., $|\dot{\tilde{\mathbf{d}}}_{\alpha\beta}| \leq \bar{\mathbf{D}}_{\alpha\beta}$, if there exists*

$$\begin{cases} \lambda_{\alpha\beta} > 2, \\ \varphi_{\alpha\beta} > \frac{\lambda_{\alpha\beta}^3 + \bar{\mathbf{D}}_{\alpha\beta}^2(4\lambda_{\alpha\beta} - 8)}{\lambda_{\alpha\beta}(4\lambda_{\alpha\beta} - 8)}, \end{cases} \quad (5.18)$$

then (5.17) achieves stability in finite time with

$$t \geq T = \frac{2\lambda_{\max}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})}{\lambda_{\min}(\mathbf{Q}_{\alpha\beta})} V_{\alpha\beta}^{\frac{1}{2}}(\tilde{\mathbf{i}}_{\alpha\beta}(0), \phi_{\alpha\beta}(0)), \quad (5.19)$$

where

$$\begin{aligned} \mathbf{P}_{\alpha\beta} &= \begin{bmatrix} 2\varphi_{\alpha\beta} + \frac{1}{2}\lambda_{\alpha\beta}^2 & -\frac{1}{2}\lambda_{\alpha\beta} \\ -\frac{1}{2}\lambda_{\alpha\beta} & 1 \end{bmatrix}, \\ \mathbf{Q}_{\alpha\beta} &= \begin{bmatrix} \lambda_{\alpha\beta}(\varphi_{\alpha\beta} - \frac{\lambda_{\alpha\beta}}{4} + \frac{\lambda_{\alpha\beta}^2}{2}) - \bar{\mathbf{D}}_{\alpha\beta}^2 & \frac{\lambda_{\alpha\beta}}{2}(1 - \frac{\lambda_{\alpha\beta}}{2}) \\ \frac{\lambda_{\alpha\beta}}{2}(1 - \frac{\lambda_{\alpha\beta}}{2}) & \frac{\lambda_{\alpha\beta}}{2} - 1 \end{bmatrix}, \end{aligned}$$

$V_{\alpha\beta}$ is the Lyapunov function for the system (5.17), $\lambda_{\max}(\cdot)$ and $\lambda_{\min}(\cdot)$ denote the maximum and minimum eigenvalue, respectively, and $\tilde{\mathbf{i}}_{\alpha\beta}(0)$ and $\phi_{\alpha\beta}(0)$ are initial value.

Proof Considering the current dynamic system (5.17), a Lyapunov function is selected as,

$$V_{\alpha\beta} = \varsigma_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta}, \quad (5.20)$$

where $\varsigma_{\alpha\beta} = [\varsigma_{\alpha\beta 1}, \varsigma_{\alpha\beta 2}]^T = \left[|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}} \text{sign}(\tilde{\mathbf{i}}_{\alpha\beta}), \phi_{\alpha\beta} \right]^T$, and $\mathbf{P}_{\alpha\beta} = \begin{bmatrix} 2\varphi_{\alpha\beta} + \frac{1}{2}\lambda_{\alpha\beta}^2 & -\frac{1}{2}\lambda_{\alpha\beta} \\ -\frac{1}{2}\lambda_{\alpha\beta} & 1 \end{bmatrix}$

is a given positive definite symmetric matrix.

The time-derivative of $\varsigma_{\alpha\beta}$ can be written by $|\dot{\tilde{\mathbf{i}}}_{\alpha\beta}| = \dot{\tilde{\mathbf{i}}}_{\alpha\beta} \text{sign}(\tilde{\mathbf{i}}_{\alpha\beta})$,

$$\begin{aligned} \dot{\varsigma}_{\alpha\beta} &= \frac{1}{|\varsigma_{\alpha\beta 1}|} (\mathbf{A}_{\alpha\beta} \varsigma_{\alpha\beta} + \mathbf{B}_{\alpha\beta} \tau_{\alpha\beta}) \\ &= \frac{1}{|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}}} (\mathbf{A}_{\alpha\beta} \varsigma_{\alpha\beta} + \mathbf{B}_{\alpha\beta} \tau_{\alpha\beta}), \end{aligned} \quad (5.21)$$

where $\mathbf{A}_{\alpha\beta} = \begin{bmatrix} -\frac{\lambda_{\alpha\beta}}{2} & \frac{1}{2} \\ -\varphi_{\alpha\beta} & 0 \end{bmatrix}$, $\mathbf{B}_{\alpha\beta} = [0, 1]^T$, and $\tau_{\alpha\beta} = |\varsigma_{\alpha\beta 1}| \dot{\tilde{\mathbf{d}}}_{\alpha\beta}$.

The derivative of (5.20),

$$\begin{aligned} \dot{V}_{\alpha\beta} &= 2\dot{\varsigma}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta} \\ &= \frac{1}{|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}}} (2\varsigma_{\alpha\beta}^T \mathbf{A}_{\alpha\beta}^T + 2\tau_{\alpha\beta} \mathbf{B}_{\alpha\beta}^T) \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta} \\ &\leq \frac{1}{|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}}} (2\varsigma_{\alpha\beta}^T \mathbf{A}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta} + 2\tau_{\alpha\beta} \mathbf{B}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta} \\ &\quad + \bar{D}_{\alpha\beta}^2 |\tilde{\mathbf{i}}_{\alpha\beta}| - \tau_{\alpha\beta}^2) \\ &\leq \frac{1}{|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}}} \varsigma_{\alpha\beta}^T (\mathbf{A}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} + \mathbf{P}_{\alpha\beta} \mathbf{A}_{\alpha\beta} + \bar{D}_{\alpha\beta}^2 \mathbf{\Gamma}_{\alpha\beta} \\ &\quad + \mathbf{P}_{\alpha\beta} \mathbf{B}_{\alpha\beta} \mathbf{B}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta}) \varsigma_{\alpha\beta} \end{aligned} \quad (5.22)$$

where $\mathbf{\Gamma}_{\alpha\beta} = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix}$.

The equation (5.22) can be rewritten ,

$$\dot{V}_{\alpha\beta} \leq -\frac{1}{|\tilde{\mathbf{i}}_{\alpha\beta}|^{\frac{1}{2}}} \varsigma_{\alpha\beta}^T \mathbf{Q}_{\alpha\beta} \varsigma_{\alpha\beta}, \quad (5.23)$$

where $\mathbf{Q}_{\alpha\beta} = -(\mathbf{A}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} + \mathbf{P}_{\alpha\beta} \mathbf{A}_{\alpha\beta} + \bar{D}_{\alpha\beta}^2 \mathbf{\Gamma}_{\alpha\beta} + \mathbf{P}_{\alpha\beta} \mathbf{B}_{\alpha\beta} \mathbf{B}_{\alpha\beta}^T \mathbf{P}_{\alpha\beta})$.

Put $\mathbf{A}_{\alpha\beta}$, $\mathbf{P}_{\alpha\beta}$, $\mathbf{B}_{\alpha\beta}$, and $\mathbf{\Gamma}_{\alpha\beta}$ in $\mathbf{Q}_{\alpha\beta}$, then

$$\mathbf{Q}_{\alpha\beta} = \begin{bmatrix} \lambda_{\alpha\beta}(\varphi_{\alpha\beta} - \frac{\lambda_{\alpha\beta}}{4} + \frac{\lambda_{\alpha\beta}^2}{2}) - \bar{D}_{\alpha\beta}^2 & \frac{\lambda_{\alpha\beta}}{2}(1 - \frac{\lambda_{\alpha\beta}}{2}) \\ \frac{\lambda_{\alpha\beta}}{2}(1 - \frac{\lambda_{\alpha\beta}}{2}) & \frac{\lambda_{\alpha\beta}}{2} - 1 \end{bmatrix}.$$

Based on the Schur complement Lemma for satisfying $\mathbf{Q}_{\alpha\beta} > 0$ to guarantee $\dot{V}_{\alpha\beta} < 0$, one can select,

$$\begin{aligned} \lambda_{\alpha\beta} &> 2, \\ \varphi_{\alpha\beta} &> \frac{\lambda_{\alpha\beta}^3 + \bar{D}_{\alpha\beta}^2(4\lambda_{\alpha\beta} - 8)}{\lambda_{\alpha\beta}(4\lambda_{\alpha\beta} - 8)}, \end{aligned} \quad (5.24)$$

Based on (5.20), it can be obtained,

$$\lambda_{\min}(\mathbf{P}_{\alpha\beta}) \|\varsigma_{\alpha\beta}\|_2^2 \leq \varsigma_{\alpha\beta}^T \mathbf{P}_{\alpha\beta} \varsigma_{\alpha\beta} \leq \lambda_{\max}(\mathbf{P}_{\alpha\beta}) \|\varsigma_{\alpha\beta}\|_2^2, \quad (5.25)$$

where $\|\varsigma_{\alpha\beta}\|_2^2 = |\tilde{\mathbf{i}}_{\alpha\beta}| + \phi_{\alpha\beta}^2$.

According to (5.25),

$$\frac{V_{\alpha\beta}^{\frac{1}{2}}}{\lambda_{\max}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})} \leq \|\varsigma_{\alpha\beta}\|_2, \quad (5.26)$$

Further step, it can be obtained that,

$$|\varsigma_{\alpha\beta 1}| = \left| \tilde{i}_{\alpha\beta} \right|^{\frac{1}{2}} \leq \|\varsigma_{\alpha\beta}\|_2 \leq \frac{V_{\alpha\beta}^{\frac{1}{2}}}{\lambda_{\min}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})}. \quad (5.27)$$

Combine (5.23) and (5.27),

$$\begin{aligned} \dot{V}_{\alpha\beta} &\leq -\frac{1}{|\varsigma_{\alpha\beta 1}|} \lambda_{\min}(\mathbf{Q}_{\alpha\beta}) \|\varsigma_{\alpha\beta}\|_2^2 \\ &= -\frac{\|\varsigma_{\alpha\beta}\|_2}{|\varsigma_{\alpha\beta 1}|} \lambda_{\min}(\mathbf{Q}_{\alpha\beta}) \|\varsigma_{\alpha\beta}\|_2 \\ &\leq -\lambda_{\min}(\mathbf{Q}_{\alpha\beta}) \|\varsigma_{\alpha\beta}\|_2. \end{aligned} \quad (5.28)$$

According to (5.26) and (5.28), one can get that,

$$\dot{V}_{\alpha\beta} \leq -\frac{\lambda_{\min}(\mathbf{Q}_{\alpha\beta})}{\lambda_{\max}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})} V_{\alpha\beta}^{\frac{1}{2}} \quad (5.29)$$

and (5.29) is given by

$$V_{\alpha\beta}^{\frac{1}{2}}(\tilde{i}_{\alpha\beta}, \varphi_{\alpha\beta}) \leq V_{\alpha\beta}^{\frac{1}{2}}(\tilde{i}_{\alpha\beta}(0), \varphi_{\alpha\beta}(0)) - \frac{\lambda_{\min}(\mathbf{Q}_{\alpha\beta})}{2\lambda_{\max}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})} t, \quad (5.30)$$

This means at least for $t \geq T = 2\frac{\lambda_{\max}^{\frac{1}{2}}(\mathbf{P}_{\alpha\beta})}{\lambda_{\min}(\mathbf{Q}_{\alpha\beta})} V_{\alpha\beta}^{\frac{1}{2}}(\tilde{i}_{\alpha\beta}(0), \phi_{\alpha\beta}(0))$, it can be obtained the $\dot{V}_{\alpha\beta} \leq 0$, then (5.17) is asymptotic stability in finite-time T , that is, $\varsigma_{\alpha\beta} = 0$, then $\tilde{i}_{\alpha\beta} = 0$. ■

Based on the Theorem I, it can be concluded that $\tilde{i}_{\alpha\beta} \rightarrow i_{\alpha\beta}^*$.

5.3.3 Voltage Regulation Loop

In this subsection, an ESMDO-based P control is applied to the voltage regulator to dramatically enhance the dc-link voltage tracking performance against the lumped disturbances, including parameter uncertainties, measurement noises, and external unknown disturbances.

The voltage dynamic introduced in (5.2) can be rewritten as,

$$\begin{aligned} \dot{z} &= \frac{2}{C} p^* - \frac{4z}{C} \frac{1}{R_L} + (\Delta C p^* - 4z \Delta C) \\ &= \frac{2}{C} p^* - \frac{2}{C} g, \end{aligned} \quad (5.31)$$

where $z = \frac{1}{2} v_{dc}^2$, $p^* = \frac{v_{dc}}{2} u_{\alpha\beta}^T i_{\alpha\beta}^*$, g is considered as the lumped disturbances, including capacitance parameter deviation ΔC and unknown load R_L , and measurement noises.

Based on (5.31), the extended system is designed as,

$$\begin{cases} \dot{z} = \frac{2}{C} p^* - \frac{2}{C} g, \\ \dot{g} = \delta(t), \end{cases} \quad (5.32)$$

where $\delta(t)$ is the rate of lumped disturbances g .

5.3.3.1 Extended Sliding Mode Disturbance Observer (ESMDO) Design

According to (5.32), the ESMDO is designed,

$$\begin{cases} \dot{\hat{z}} = \frac{2}{C}p^* - \frac{2}{C}\hat{g} + u_{smf}, \\ \dot{\hat{g}} = \gamma u_{smf}, \end{cases} \quad (5.33)$$

where $\hat{g}$ is the estimate value of g , u_{smf} is the ESMDO function to be designed, and $\gamma < 0$ is the ESMDO gain.

Combined (5.32) with (5.33), the error system is obtained as follows,

$$\begin{cases} \dot{\tilde{z}} = -\frac{2}{C}\tilde{g} - u_{smf}, \\ \dot{\tilde{g}} = \delta(t) - \gamma u_{smf}, \end{cases} \quad (5.34)$$

where $\tilde{z} = z - \hat{z}$ and $\tilde{g} = g - \hat{g}$.

The time-derivative of $\tilde{z}$ can be designed as,

$$\dot{\tilde{z}} = -\varepsilon \text{sign}(\tilde{z}) - \zeta \tilde{z}, \quad (5.35)$$

where ε and ζ are positive values.

Substituted (5.34) into (5.35), one can be got

$$u_{smf} = \varepsilon \text{sign}(\tilde{z}) + \zeta \tilde{z} - \frac{2}{C}g, \quad (5.36)$$

where $\frac{2}{C}g$ is regarded as disturbance of u_{smf} .

The sliding mode variable is designed as

$$s = \tilde{z}. \quad (5.37)$$

Then to construct the following Lyapunov function for (5.37),

$$V = \frac{1}{2}s^2. \quad (5.38)$$

The derivative-time of (5.38) is presented,

$$\begin{aligned} \dot{V} &= \tilde{z}\dot{\tilde{z}} \\ &= \tilde{z}\left(-\frac{2}{C}\tilde{g} - u_{smf}\right) \\ &= \tilde{z}\left(-\frac{2}{C}\tilde{g} - \varepsilon \text{sign}(\tilde{z}) - \zeta \tilde{z}\right) \\ &\leq \|\tilde{z}\| \left(-\frac{2}{C}\tilde{g} - \varepsilon\right). \end{aligned} \quad (5.39)$$

It can be concluded that if $\varepsilon \geq \frac{2}{C}\|\tilde{g}\|$, the $\dot{V} \leq 0$, $t \rightarrow \infty$, i.e.,

$$\tilde{z} = \dot{\tilde{z}} = 0. \quad (5.40)$$

Using (5.34), it can be obtained that

$$\dot{\tilde{g}} + \gamma \frac{2}{C} \tilde{g} - \delta(t) = 0. \quad (5.41)$$

The solution of $\tilde{g}$ can be got as,

$$\tilde{g} = e^{\frac{2}{C}\gamma t} \left(\int \delta(t) e^{-\frac{2}{C}\gamma t} dt + G \right), \quad (5.42)$$

where G is a constant. Hence, g can be asymptotically estimated by $\hat{g}$ and the convergence speed is related to γ .

Define voltage regulation error $e_z = z - z^*$, in which z^* is desired value of z .

Based on (5.31), the error dynamic can be expressed as,

$$\dot{e}_z = \frac{2}{C} p^* - \frac{2}{C} g - \dot{z}^*. \quad (5.43)$$

5.3.3.2 ESMDO-based Proportional (ESMDO-P) Control Design

A P controller coupled with the ESMDO for (5.43) is proposed as

$$p^* = \frac{C}{2} (-k e_z + \frac{2}{C} \hat{g}), \quad (5.44)$$

where $k > 0$ is controller gain.

Moreover, put the composite controller (5.44), one can obtain the following closed system,

$$\dot{e}_z = -k e_z - \frac{2}{C} \tilde{g}. \quad (5.45)$$

It can concluded that e_z converges to zero when $t \rightarrow \infty$, then z^* can be tracked by z , i.e., the proposed method ESMDO-P can drive dc-link voltage v_{dc} to track its reference voltage v_{dc}^* .

5.3.4 Voltage Balancing Loop

In this section, dynamics of the dc-link capacitor voltage difference v_{dc}^o should be maintained zero to balance the capacitor voltages. Then (5.3) can reexpressed as,

$$\dot{v}_{dc}^o = \frac{2p^*}{C\sqrt{6}v_{dc}} u_\gamma + \frac{1}{C} \psi, \quad (5.46)$$

where u_γ is control input, and ψ is unknown disturbances.

5.3.4.1 Proportional-integral (PI) Control Design

A PI is implemented to balance effectively two capacitors voltage as follows,

$$u_\gamma = -k_{pb}(\tilde{v}_{dc}^o) + k_{ib} \int (\tilde{v}_{dc}^o) dt, \quad (5.47)$$

where k_{pb} and k_{ib} are control gains with positive values, $\tilde{v}_{dc}^o = v_{dc}^o - v_{dc}^{o*}$.

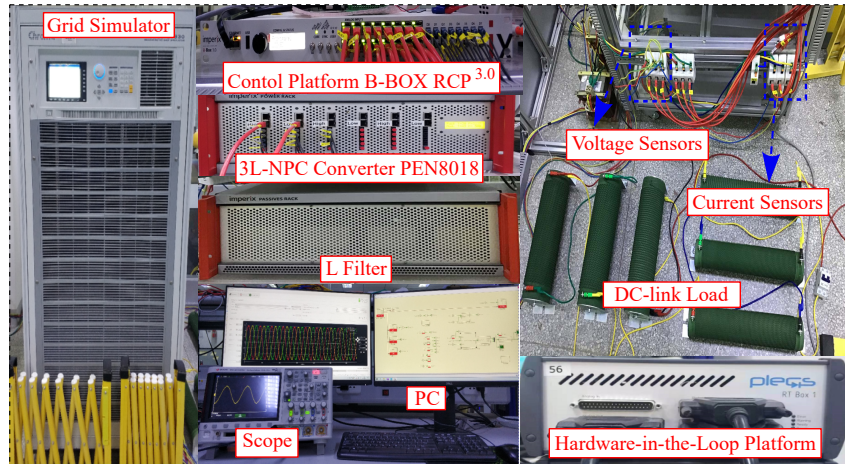


Figure 5.3: The lab-constructed 3L-NPC power converter prototype of Chap. 5.

5.4 Experimental Results

The experiments have been carried out on a lab-constructed 3L-NPC power converter prototype as shown in Figure 5.3. For the sake of providing a fair comparison, all control schemes are tuned to the best performance. The parameters of the experimental test bench are summed up in Table 5.2 and the parameters of the proposed controller gains and the conventional controller gains are found in Table 5.3.

Table 5.2: Parameters Used For Experiments

Description	Parameter	Value	Unites
Sampling frequency	f_s	10	kHz
Switching frequency	f_{sw}	10	kHz
Grid frequency	f	$50 \leftrightarrow 45 \leftrightarrow 55$	Hz
Grid phase voltage	v_{abc}	110	V
Phase inductor	L	5	mH
dc-link capacitor	C	1551	uF
Reference output voltage	v_{dc}^*	$400 \leftrightarrow 450$	V
Load resistance	R_L	$120 \leftrightarrow 80$	Ω

5.4.1 An Overall Evaluation Comparison

To assess the effectiveness and feasibility of the proposed control schemes (LO-STA, ESMDO-P), the overall comparative behavior between the proposed and the conventional control schemes (PR, PI) is collected in Figures 5.4 (a) and 5.4(b). For both, the dc-link load maintains 120Ω , v_{dc}^* is forced to step from 400V to 450V at 0.1s back to 400V at 0.6s, and the reactive power reference q^* steps from 0Var to 1000Var at 1.1s then to 0Var again at 1.6s during the whole process. In addition, the balancing voltage loop of both is adopted the same PI control. As it

Table 5.3: Controllers Parameters of Experiment

Control strategy	Current tracking loop	Voltage regulation loop	Voltage balancing loop
Proposed strategy (LO-STA, ESMDO-P)	$\lambda_\alpha = \lambda_\beta = 0.3,$ $\varphi_\alpha = \varphi_\beta = 180,$ $\mathcal{L} = \begin{bmatrix} 4e2 \\ -58696 \end{bmatrix},$	$k = 8e - 2,$ $\gamma = -3e - 2,$ $\varepsilon = 6e4,$ $\zeta = 5e3,$	$k_{pb} = 9e - 3,$ $k_{ib} = 6e - 4,$
Conventional strategy (PR,PI)	$k_{p\alpha} = k_{p\beta} = 0.2,$ $k_{r\alpha} = k_{r\beta} = 200,$ $\omega_c = 0.3,$	$k_p = 0.01,$ $k_i = 0.8,$	$k_{pb} = 9e - 3,$ $k_{ib} = 6e - 4,$

can be seen here, the overall performance of the proposed control scheme is improved, enabling accurate tracking of current/voltage in steady- and transient states with strong robustness.

5.4.2 Performance Comparisons of Voltage Regulation

To verify the effectiveness and superiority of the proposed ESMDO-P control method over the PI control method, two cases are discussed: 1) *the capacitor voltage v_{dc}^* changes*; 2) *the dc-load R_L changes*. The same test conditions are set up in the current tracking loop and voltage balancing loop, utilizing the PR controller and the PI controller, respectively. From Figures 5.5 and 5.6, the proposed ESMDO-P control method not only has the smaller startup overshoot/undershoot voltage and the shorter settling time, but also has the better static performance in comparison with PI control method. Contrast details are shown in Table 5.4. As a result, the voltage regulation performance by the proposed ESMDO-P control method has a great improvement.

5.4.3 Performance Comparison of Current Tracking

To verify the superior performance of the proposed LO-STA control method in terms of current tracking and grid current total harmonic distortion (THD) over the PR control method, this test is carried out. Figure 5.7 shows the reactive power reference q^* steps from 0Var to 2000Var with $v_{dc}^* = 400V$, $R_L = 120\Omega$, meanwhile, voltage regulation loop and voltage balance loop are consistently controlled by PI controller. It is obvious that the proposed LO-STA control method has some promotions compared with the PR control method. Check Table 5.5 for more comparison details.

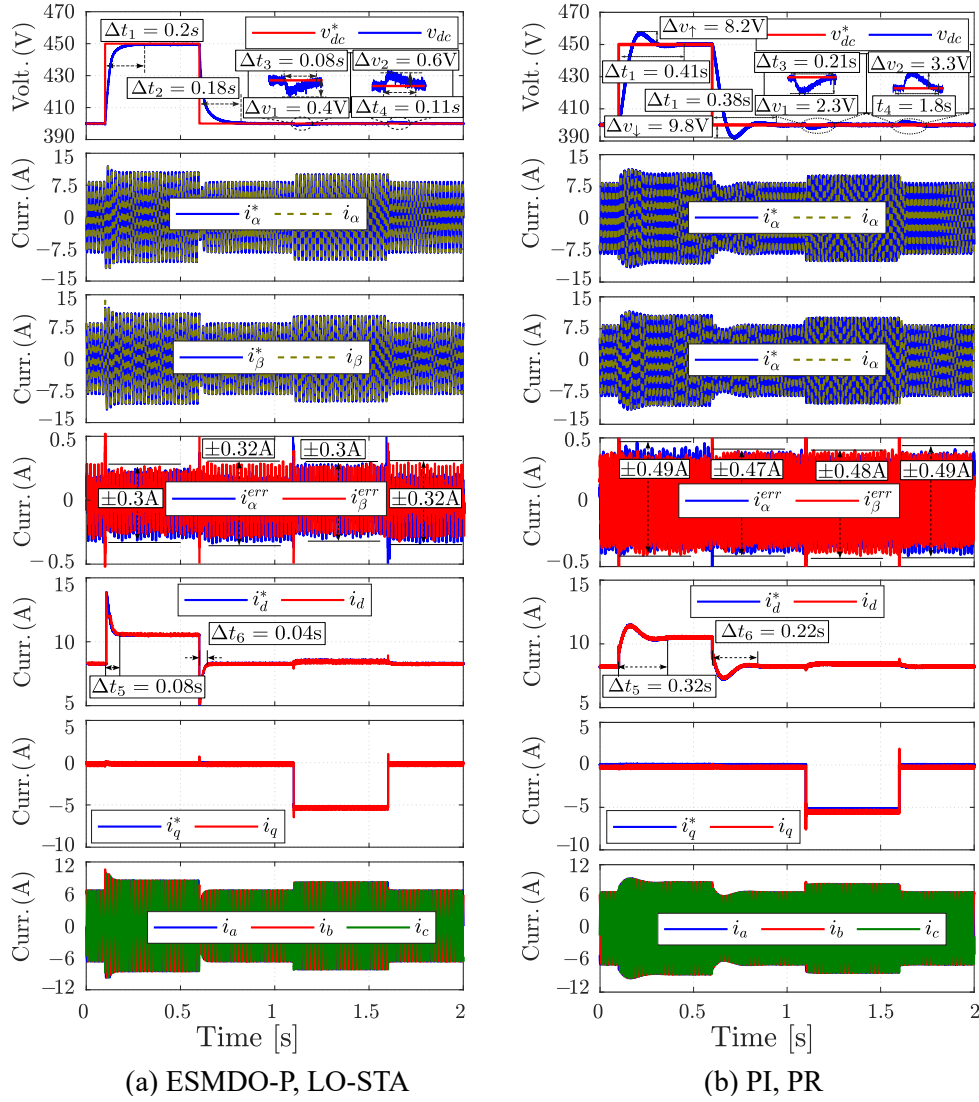


Figure 5.4: An overall evaluation comparison.

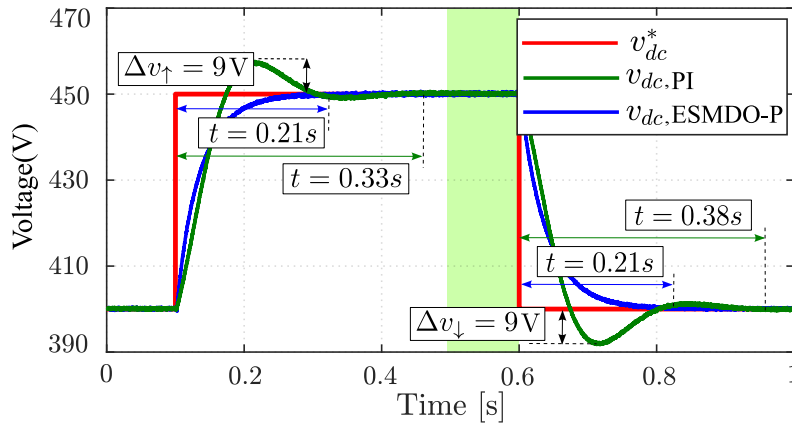
Figure 5.5: Performance comparison of voltage regulation when the capacitor voltage reference v_{dc}^* steps from 400V to 450V again back to 400V.

Table 5.4: Performance Comparison of the Proposed ESMDO-P VS PI

Dynamic Index (V)	Figure 5.5: ESMDO VS PI		Figure 5.6: ESMDO VS PI	
	Enhancement			
Overshoot voltage	no overshoot		67.7%	
Undershoot voltage	no undershoot		69.7%	
Over- settling time	36.4%		68.4%	
Under- settling time	44.7%		41.7%	
Static Index (V)	Figure 5.5 (0.5s-0.6s)		Figure 5.6 (0.9s-1s)	
	ESMDO	PI	ESMDO	PI
Mean error	0.0046	0.0301	0.0065	0.0671
Mean square error	0.1502	0.1549	0.1361	0.1522
Variance error	0.0362	0.0273	0.0185	0.0187

Table 5.5: Performance Comparisons of the Proposed LO-STA VS PR

Dynamic Index (A)		LO-STA	PR	Enhancement
Current undershoot	i_α	2	3.2	37.5%
	i_β	-	-	-
Settling time	i_α	1e-3	5e-3	80%
	i_β	0	2e-5	no settling time
Static Index (A)		LO-STA	PR	Enhancement
Mean square error	i_α	0.0954	0.1945	51.0%
	i_β	0.0928	0.1949	52.4%
Variance error	i_α	0.0091	0.0379	76.0%
	i_β	0.0086	0.038	77.4%
Max current error	i_α^{err}	0.32	0.48	33.3%
Max current error	i_β^{err}	0.3	0.48	37.5%
THD	i_a	2.8582%	3.5384%	19.2%

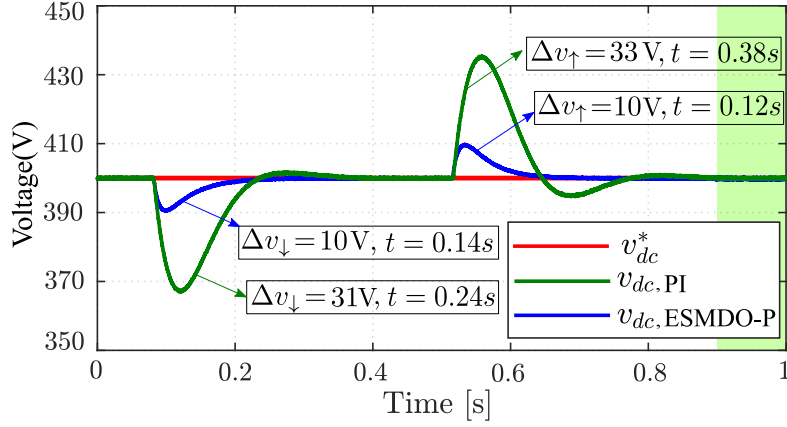


Figure 5.6: Performance comparison of voltage regulation when the dc-link load steps from 120Ω to 80Ω again back to 120Ω .

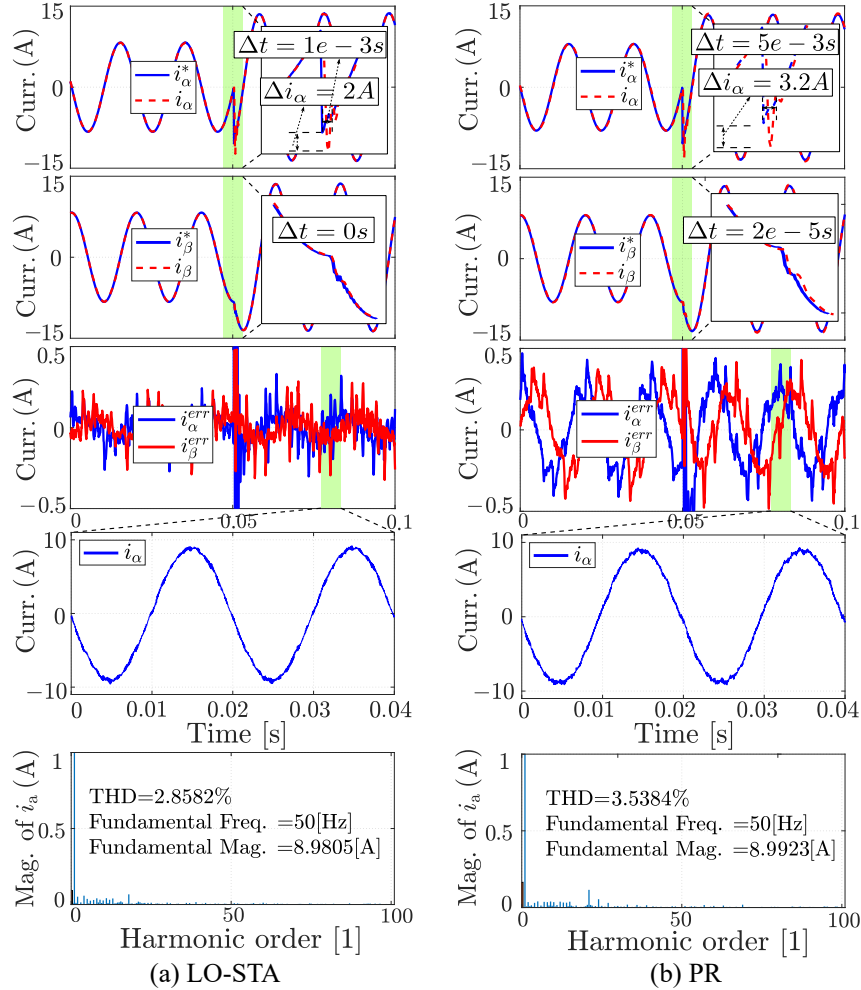


Figure 5.7: Performance comparison of current tracking i_α and i_β when reactive power reference steps from $q^*=0\text{Var}$ to $q^*=2000\text{Var}$.

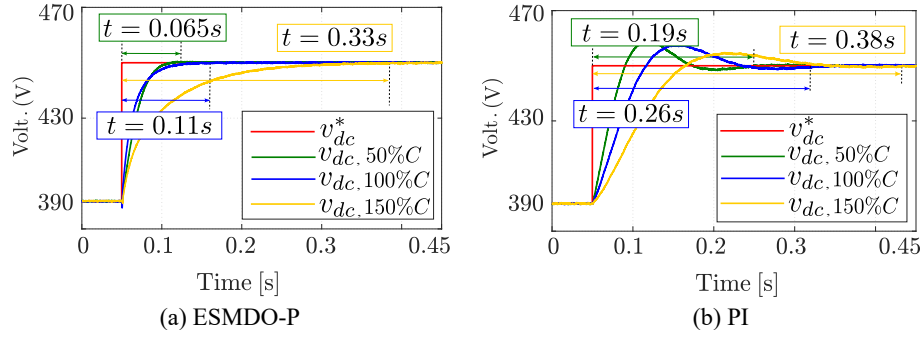


Figure 5.8: Capacitance parametric robustness comparison when the capacitor voltage reference v_{dc}^* steps from 400V to 450V.

5.4.4 Performance Evaluation with Parametric Robustness

5.4.4.1 Capacitance Parametric Robustness Comparison

To assess the dc-link capacitance C parametric robustness both the proposed ESMDO-P method and the PI method, the C is set as 100% C , 50% C and 150% C under v_{dc}^* steps from 400V to 450V with $R_L = 120\Omega$. All the above are implemented in current tracking loop using the PR controller and the voltage balancing loop using PI controller. From the experimental results shown in Figure 5.8, it is evident that the proposed ESMDO-P method has better robustness against capacitance parametric variation than PI method.

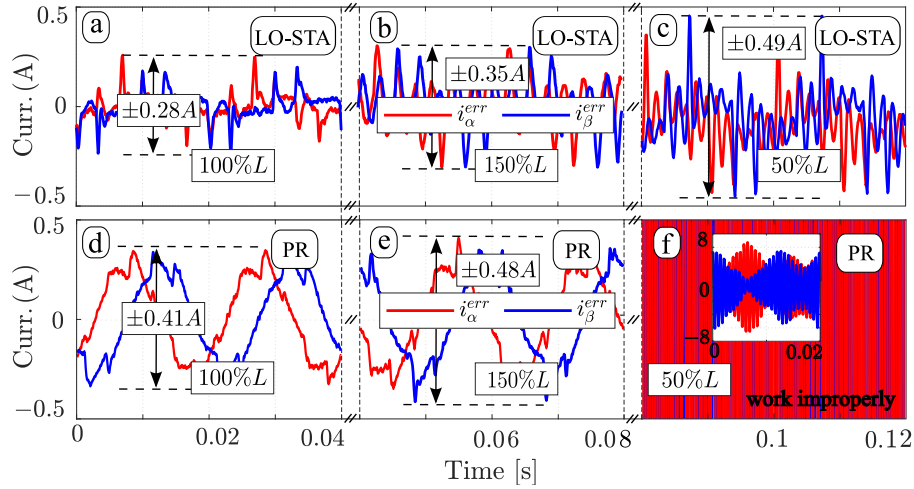


Figure 5.9: Current errors performance comparison when exist inductance deviation: 1) LO-STA (max current error): (a) $i_{\alpha}^{err}/i_{\beta}^{err} = 0.28A$, (b) $i_{\alpha}^{err}/i_{\beta}^{err} = 0.35A$, and (c) $i_{\alpha}^{err}/i_{\beta}^{err} = 0.49A$; 2) PR (max current error): (d) $i_{\alpha}^{err}/i_{\beta}^{err} = 0.41A$, (e) $i_{\alpha}^{err}/i_{\beta}^{err} = 0.48A$, and (f) work improperly.

5.4.4.2 Performance Comparison of Inductance Deviation

To further investigate robustness of the grid-side filter inductance with parametric uncertainty, the proposed LO-STA and PR methods are verified by setting 50%L, 100%L, and 150%L. The both control strategies are performed on a real-time Hardware-in-the-Loop due to mechanical

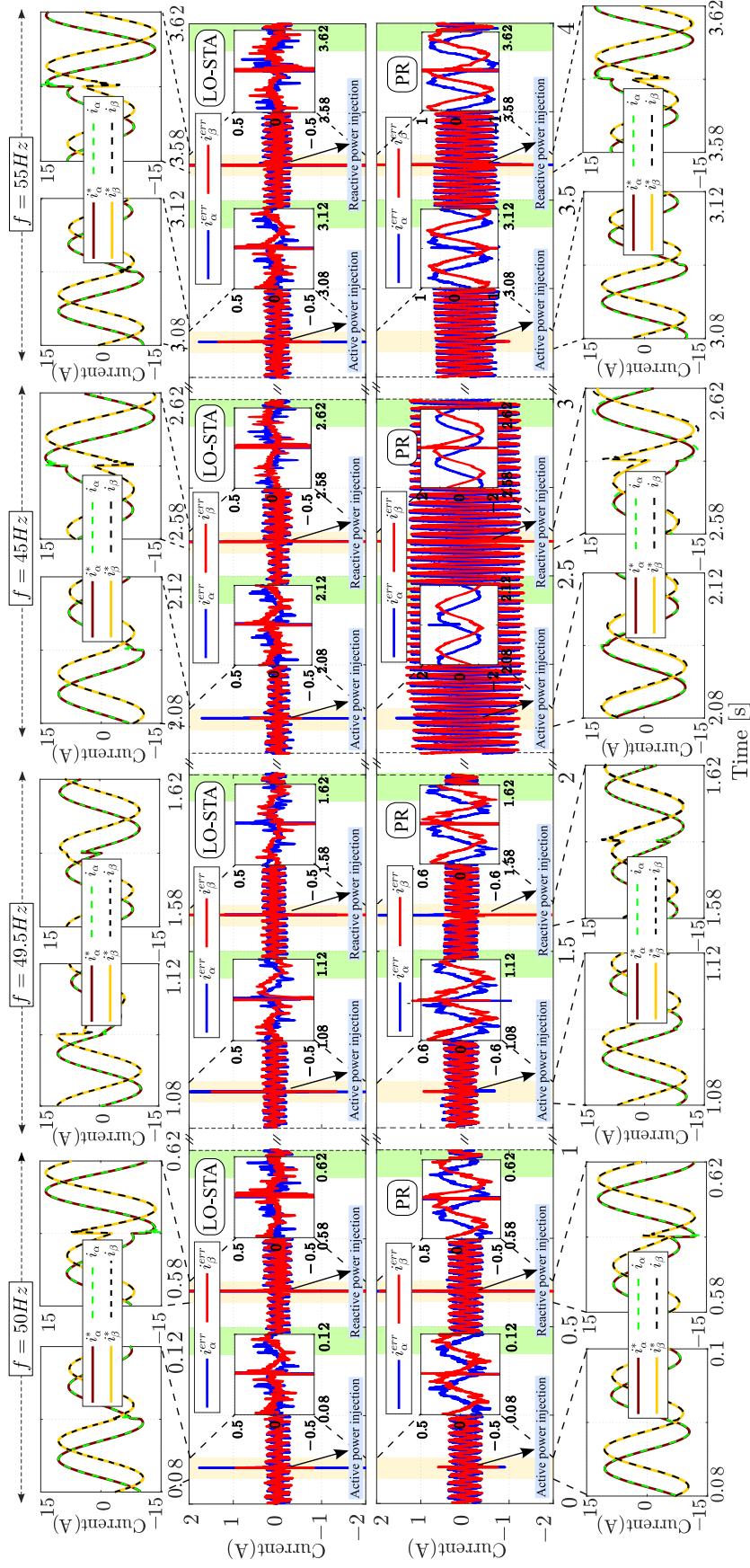


Figure 5.10: Performance comparison of grid frequency variation of current error i_a^{err} and i_b^{err} with combined active power and reactive power injection under v_{dc}^* steps from 0 Var to 2000Var at 0.6s, dc-load $R_L = 120\Omega$.

Table 5.6: The Current Static Performance Comparisons of both the Proposed LO-STA and the PR Control from Figure 5.10

Control Method	Grid Frequency	50 Hz				45 Hz				55 Hz			
		Active Power		Reactive Power		Active Power		Reactive Power		Active Power		Reactive Power	
		(0.4s-0.5s)		(0.9s-1s)		(1.4s-1.5s)		(1.9s-2s)		(3.4s-3.5s)		(3.9s-4s)	
LO-STA	Index(A)	MSE ¹	Var ²	MSE	Var	MSE	Var	MSE	Var	MSE	Var	MSE	Var
	i_α	0.0784	0.0062	0.0917	0.0096	0.0736	0.0054	0.0984	0.0097	0.1996	0.0400	0.0999	0.0100
	i_β	0.0775	0.0060	0.0930	0.0087	0.0685	0.0047	0.0929	0.0087	0.1997	0.0401	0.1000	0.0100
PR	i_α	0.1996	0.0400	0.1899	0.0362	0.9011	0.8160	0.8535	0.7320	0.4293	0.1846	0.4056	0.1646
	i_β	0.1997	0.0401	0.1898	0.0362	0.8114	0.6455	0.7776	0.5941	0.4547	0.2048	0.4225	0.1771

1: Mean Square Error (A). 2: Variance Error (A).

limitations of the real plant. When the inductance is 50%L, 100%L, 150%L, the details of the current errors i_{α}^{err} and i_{β}^{err} are shown in Figure 5.9. Specifically, compared with the PR method, the proposed method LO-STA mitigates the current tracking error (in the case of i_{α}^{err}) by 39.0%, 23.4% corresponding to 50%L, 100%L, respectively. Furthermore, the current tracking error of proposed method stays within a reasonable less than 0.5A when the inductance is 150%L, but the PR method does not work properly as shown in Figure 5.9(f). In summary, it can be concluded that the proposed LO-STA current control is more robust for current tracking in dealing with inductance parameter uncertainty.

5.4.5 Performance Comparison of Grid Frequency Variation

To demonstrate the proposed LO-STA current control is superior to the PR control method in terms of grid frequency variation, this case is verified with grid frequencies 45Hz, 49.5Hz, 50Hz, and 55Hz, considering the combined active and reactive power injection. The grid frequency 45Hz and 55Hz are selected to make the comparison of test results more obvious. Taking into account the same test condition, both the voltage regulation loop and the voltage balancing loop are controlled by the PI controller. From Figure 5.10, it can be observed that the current tracking errors of i_{α}^{err} and i_{β}^{err} (max current error) are almost identical (less than 0.35A) by utilizing the proposed LO-STA control method. However, for the PR control method, when grid-frequency becomes from 50Hz to 49.5Hz, then to 45Hz and finally to 55Hz, while obvious changes of the current tracking error increase in the above situation displayed from 0.47A to 0.58A, then to 1.8A and finally to 0.9A, respectively. Hence, it can be concluded that the proposed LO-STA control is more robust against the grid frequency variation than the PR control. Additionally, the static performance comparisons between the proposed LO-STA control and the PR control are shown in Table 5.6, in which proposed LO-STA current control exhibits better static performance. Based on the above, it can be concluded that the proposed LO-STA current control outperforms the PR in terms of dealing with the robustness of grid frequency variation.

5.5 Summary

3L-NPC Power converters play an important role in microgrids, as a necessary interface to microgrids. A reasonable control scheme is a key to achieve high efficiency operation of the converter. However, some drawbacks associated with dynamic-/steady-state performance and anti-disturbance capability are present in the type of existing control strategy. This chapter proposes the robust high-quality current control scheme to address the above problems. Specifically, the STA coupled with the LO current controller is adopted to deal with the sinusoidal current tracking with inductance/grid frequency deviations and sinusoidal disturbance, ensuring the current tracking performance. Additionally, the P controller coupled with the ESMDO is proposed to regulate voltage, achieving fast and accurate tracking voltage. From all experimental results, it can be concluded that the proposed control scheme can attain exceptional voltage regulation and current tracking performances such as more stable output voltage, less current tracking error and lower THD than the conventional control method under various parameter deviations/disturbances.

CHAPTER 6

Data-driven Model-Free Adaptive Control for 3L-NPC Power Converters

Model-based control is widely used in converters due to its ease, but it poses challenges when dealing with inaccurate models, critical loads, and multidisturbances. To fix them, this work proposes a data-driven MFAC for a 3L-NPC power converter system to realize robust, high-quality current and voltage control without any system parameters. Specifically, an MFAC is a potential method for estimating the system dynamics by designing adaptive laws derived intuitively from Lyapunov theory to regulate current and voltage, guaranteeing self-adaptability to unmodeled dynamics, model variations, parameter mismatches, and disturbances. Striving for easier access to adaptive law gains, a particle swarm optimization (PSO) algorithm is embedded in MFAC to automatically determine them utilizing the fitness function defined by the voltage and current. Experimental data confirm that the proposal outperforms artificial neural networks-based PI and super-twisting algorithm-based model control schemes under multi-scenarios in terms of transient-/ steady-state, grid current harmonic distortions, and robustness.

6.1 Introduction

Power converter control has always aimed at achieving high-quality grid current, accurate voltage, and (or) current tracking and strong robustness, particularly in critical loads, distorted models, and parameter mismatches, remaining cost-effective and easy-to-design. Traditionally, PI control has dominated in various converter domains [166], primarily recognized for its low cost and simplicity. However, PI control is limited in its adaptation to complex systems, trade-offs between stability and performance, and concerns regarding robustness. To deal with these issues, [167] introduces a cascaded proportional control with algebraic estimation for ac/dc converters, aiming to regulate the average dc-bus energy and instantaneous input power while addressing system nonlinearity, load power variations, and uncertainties. Certainly, nonlinear control methods (e.g., multi-agent control [168], sliding mode control [155], neural network control [141], and

event-triggered control [169] etc.) are also widely developed, enabling providing stability support during power/current and voltage fluctuations facing disturbances and uncertainties.

An AC method is proposed in [170] for regulating the dc-voltage output of a three-phase power converter, focusing on estimating the external dc-load disturbances. To successfully use AC methods, it is usually essential to identify time-varying model parameters or design adaptive laws; this method cannot be applied when these key parameters are difficult to identify or estimate online. To circumvent this drawback, an STA control is proposed in [171] to improve the voltage regulation, addressing the external dc-load disturbances. However, this method needs to improve on the drawbacks of chattering and poor tracking performance. Then an H_∞ control strategy is devised to achieve optimal, robust, and multi-variable control, encompassing grid synchronization, active power, and voltage regulation, specifically tailored for power converters in [172]. Despite this, this control design approach tends to be overly conservative, prioritizing robustness at the expense of transient performance, as it conservatively handles worst-case model uncertainties. The above methods effectively tackle disturbance rejection via FBC [173]. However, FBC adjustment may exhibit a relatively slow response when attenuating strong disturbances. In contrast, FFC [174] represents a direct and actively prevents disturbances technique. With the recognition of the benefits of FFC as well as the desire to overcome the limitations of FBC, disturbance observer-based controls have been highly sought after, such as [175, 176].

A robust control that integrated a DSTO with QISMCMC is presented in [21], tailored for a 3L-NPC power converter system, markedly improving power-voltage regulation performance and anti-disturbance capability. In addition, an observer-based optimal voltage control is proposed for three-phase uninterruptible power supply systems of [177], utilizing an optimal control term to minimize system errors and a compensating observer to address uncertainties. These two ways effectively address multi-disturbance and uncertainty issues, but the models still restrict them. These methods are often associated with additional assumptions and consider factors such as model uncertainty, making it difficult to quantify their limitations through physical or recognition modeling. As a result of these issues, the computation complexity can be increased, and in some cases, even the performance of the control can be impaired.

MFC, renowned for its simplicity and robustness, is increasingly in demand. It offers the features: (i) precise control of system dynamics and stability performance without requiring knowledge of the plant model; (ii) ensuring consistent operation even in the face of model variations. Inspired by this, the idea in [129] is improved current control performance, introducing a model-free predictive current control (MFPCC) based on an ultra-local model (ULM) with an ESO for PMSM drives that do not need motor details. Moreover, the bandwidth limitation in the ESO will reduce the control performance. As a result, [178] proposes the MFPCC based on ULM using an adaptive sliding mode observer to improve system parameter robustness without bandwidth limitations relative to [129]. These two methods effectively enhance robustness; however, ULMs, while inherently addressing local control issues, present complex control challenges when faced with systems that exhibit significant variability or high non-linearity. Accordingly, some studies of MFC based on data-driven methods have been conducted for the power converter system. [84] combines data-driven with dual droop control for ac/dc interlink converter to achieve effective power sharing. Despite reducing reliance on accurate models, this control method may face problems due to input coupling and operational constraints in complex systems. The robust predictive control employs data-driven neural predictors to improve the

issue, as presented in [179]. This method leverages real-time and historical data to estimate unknown system dynamics, thereby boosting control robustness and flexibility. It may still be necessary to carefully tune the architecture and learning parameters of the neural network.

Inspired by the issues above, this work proposes a novel data-driven model-free adaptive control (MFAC) approach for a 3L-NPC power converter system in MDs, in which the data-driven MFAC dynamically adjusts real-time input data based on updated system outputs, following pre-defined adaptive current/voltage laws, while optimizing all MFAC coefficients using the particle swarm optimization (PSO) algorithm.

Contributions of this work can be summarized as:

1. A data-driven MFAC for the 3L-NPC converter system solely relies on the input and output data of the system without any of its parameters, achieving accurate current/voltage tracking while effectively mitigating tracking errors, dynamic over-/under-shoot, and grid current harmonics, thereby enhancing the system stability, robustness, and flexibility.
2. Towards alleviating the challenge of selecting adaptive coefficients for the data-driven MFAC, a PSO algorithm is employed to autonomously determine them based on a fitness function defined by the dc-voltage and dq-current, streamlining manual effort and enhancing efficiency.
3. A rigorous stability analysis of the data-driven MFAC dynamics of closed-loop systems provides a theoretical basis for the power converter system, guaranteeing self-adaptability to unmodeled dynamics, model variations, parameter mismatches, and disturbances.

6.2 Converter Description and Modeling

The topology of the 3L-NPC power converter is depicted in Figure 6.1B, accompanied by the parameter list provided in Table 6.1. The ac side connects to the three-phase ac power supply with inductors, while the dc side interfaces with various applications such as photovoltaics, wind turbines, electric vehicles, etc. Note that dc-link load change can cause abrupt variations in voltage (active power). The dc-link is connected to an unknown equivalent resistive load to simplify the following modeling and analysis.

This work adopts the mathematical model of the converter in dq -synchronous reference frame, as follows [180]:

$$\dot{i}_d = \omega i_q + \frac{1}{L} v_d - \frac{v_{dc}}{2L} u_d, \quad (6.1)$$

$$\dot{i}_q = -\omega i_d + \frac{1}{L} v_q - \frac{v_{dc}}{2L} u_q, \quad (6.2)$$

$$\dot{v}_{dc} = \frac{1}{C} u_d i_d + \frac{1}{C} u_q i_q - \frac{2}{C} i_L, \quad (6.3)$$

$$\dot{v}_{dc}^o = \frac{2p^*}{C\sqrt{6}v_{dc}} u_0 + \frac{1}{C} \psi, \quad (6.4)$$

where grid voltages e_{abc} , currents i_{abc} , and dc-voltages v_{dc}^h , v_{dc}^l are measured. Based on the Clarke-Park transformation, the voltage v_{dq} and currents i_{dq} can be represented. p^* is conducted

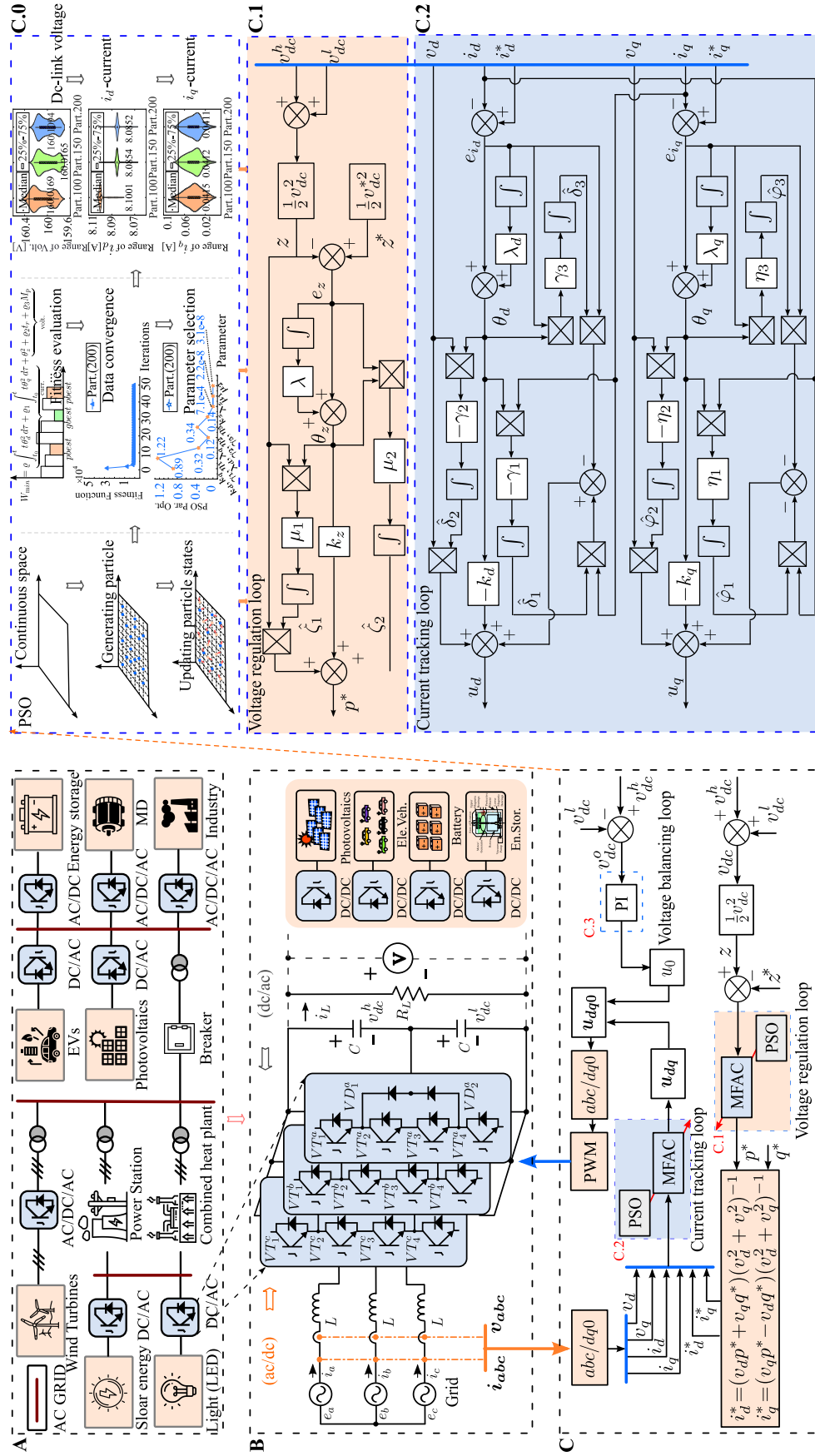


Figure 6.1: Grid-connected 3L-NPC power converter with L filter.

Table 6.1: List of Nomenclatures

Symbol	Description
e_{abc}	$[e_a, e_b, e_c]^T$, grid voltage vector in abc frame
i_{abc}	$[i_a, i_b, i_c]^T$, grid current vector in abc frame
v_{dq}	$[v_d, v_q]^T$, grid voltage vector in $dq - axis$
i_{dq}	$[i_d, i_q]^T$, grid current vector in $dq - axis$
u_{dq0}	$[u_d, u_q, u_0]^T$, average duty cycles in $dq - axis$
VT_m^n	$n = \{a, b, c\}$, $m = \{1, 2, 3, 4\}$, insulated-gate bipolar transistors
VD_l^n	$n = \{a, b, c\}$, $l = \{1, 2\}$, clamping diodes
v_{dc}	$v_{dc}^h + v_{dc}^l$, the sum of dc-link capacitor voltages
C	dc-link capacitor
L	filter inductor
R_L	total dc load resistance
i_L	dc-link load current
ω	Angular frequency of grid voltage

by voltage regulation. ψ is regarded as the sinusoidal disturbance. u_d , u_q , and u_0 make the final average duty cycle, governing the power converter.

6.3 Control Scheme Design

In this subsection, the approximate model for the 3L-NPC converter system, derived from dynamic analysis, is presented in Figure 6.1-C. The cascade control system incorporates current tracking, voltage regulation, and voltage balancing.

6.3.1 Control Objectives

The control structure for the 3L-NPC power converter system is shown in Fig. 6.1-C. The proposed data-driven MFAC control is employed to achieve three goals: current tracking, voltage regulation, and capacitor voltage balancing.

1. In the voltage regulation loop, the dc-link voltage is regulated to track its reference value (i.e., $v_{dc} \rightarrow v_{dc}^*$), forced by the active power reference p^* , which serves as the input for the current tracking loop (see Fig. 6.1-C.1).
2. In the current tracking loop, active and reactive currents are utilized to track their references using control signals u_d and u_q (i.e., $i_d \rightarrow i_d^*$ and $i_q \rightarrow i_q^*$), where i_d^* and i_q^* are derived from the voltage regulation loop. The signals u_d and u_q are this loop outputs, while q^* is given to adjust the power factor (see Fig. 6.1-C.2).

3. In the voltage balancing loop, the goal is to minimize the voltage difference across (two) capacitors (i.e., $v_{dc}^o \rightarrow 0$). The outputs from this loop and the current tracking loop make the final average duty cycle, governing the power converter (see Fig. 6.1-C.3).

6.3.2 Current Tracking Loop

An MFAC is proposed to drive the dq -current $i_d \rightarrow i_d^*$, and $i_q \rightarrow i_q^*$ in the current tracking loop, effectively dealing with issues caused by unmodeled dynamics, inductance parameter mismatches, and disturbances (see Figure 6.1-C.2).

Building upon (6.1) and (6.2), the dq -axis current errors e_{i_d} and e_{i_q} are defined as follows,

$$e_{i_d} = i_d^* - i_d, \quad (6.5)$$

$$e_{i_q} = i_q^* - i_q, \quad (6.6)$$

where i_d^* and i_q^* are current reference of i_d and i_q , respectively.

By integrating integral terms into (6.5) and (6.6), θ_d and θ_q can be reexpressed as follows:

$$\theta_d = e_{i_d} + \lambda_d \int_0^t e_{i_d} d\tau, \quad (6.7)$$

$$\theta_q = e_{i_q} + \lambda_q \int_0^t e_{i_q} d\tau, \quad (6.8)$$

where $\lambda_d > 0$ and $\lambda_q > 0$ are set.

Remark 6.3.1 To further eliminate the steady-state errors of (6.5) and (6.6), the integral items $\int_0^t e_{i_d} d\tau$ and $\int_0^t e_{i_q} d\tau$ are introduced in (6.7) and (6.8), which is similar to the design of coping with input saturation, thereby enhancing the system robustness against uncertainties and disturbances.

Considering dynamics of (6.7) and (6.8) with (6.1)-(6.2), ones get,

$$\dot{\theta}_d = -wi_q - \frac{1}{L}v_d + \frac{v_{dc}}{2L}u_d + \lambda_d e_{i_d}, \quad (6.9)$$

$$\dot{\theta}_q = wi_d - \frac{1}{L}v_q + \frac{v_{dc}}{2L}u_q + \lambda_q e_{i_q}. \quad (6.10)$$

Assumption 6.3.1 In the current tracking loop, the dc-voltage v_{dc} , dq -current references i_d^* and i_q^* are very slowly during a short sampling period and bounded, i.e., $\dot{v}_{dc} = 0$, $\dot{i}_d^* = 0$, $\dot{i}_q^* = 0$.

6.3.2.1 Model-free Adaptive Control (MFAC) Design

Theorem 6.3.1 Consider the current tracking error closed-loop systems (6.9) and (6.10) with the Assumption 6.3.1. Given parameters $k_d > 0$, $k_q > 0$, $\lambda_d > 0$, $\lambda_q > 0$, $\gamma_i > 0$, and $\varphi_i > 0$ ($i \in \{1, 2, 3\}$), if there exist Lyapunov functions $V_d(\theta_d, i_q, v_d, e_{i_d})$ and $V_q(\theta_q, i_d, v_q, e_{i_q})$ that are positive definite, and only the adaptive laws (6.16) and (6.19) are designed such that

the derivatives $\dot{V}_d(\theta_d, i_q, v_d, e_{i_d}) \leq 0$ and $\dot{V}_q(\theta_q, i_d, v_q, e_{i_q}) \leq 0$, then designed MFAC controllers (6.13) and (6.14) guarantee that the (6.9) and (6.10) stability and assure θ_d and θ_q converge to zero as $t \rightarrow \infty$ in the presence of uncertainties and interactions.

Proof Construct the dq -current Lyapunov functions,

$$V_d(\theta_d, i_q, v_d, e_{i_d}) = \frac{1}{2}(\theta_d^2 + \frac{v_{dc}}{2L} \sum_{i=1}^3 \frac{1}{\gamma_i} \tilde{\delta}_i^2), \quad (6.11)$$

$$V_q(\theta_q, i_d, v_q, e_{i_q}) = \frac{1}{2}(\theta_q^2 + \frac{v_{dc}}{2L} \sum_{i=1}^3 \frac{1}{\eta_i} \tilde{\varphi}_i^2), \quad (6.12)$$

where $\gamma_i > 0$ and $\eta_i > 0$ are adaptive law gains with $i \in \{1, 2, 3\}$. $\tilde{\delta}_i(t)$ and $\tilde{\varphi}_i(t)$ are: $\tilde{\delta}_1 = \frac{2Lw}{v_{dc}} - \hat{\delta}_1$, $\tilde{\delta}_2 = \frac{2}{v_{dc}} - \hat{\delta}_2$, and $\tilde{\delta}_3 = \frac{2\lambda_d L}{v_{dc}} - \hat{\delta}_3$, where $\hat{\delta}_1$, $\hat{\delta}_2$, and $\hat{\delta}_3$ are the d -current adaptive laws; $\tilde{\varphi}_1 = \frac{2Lw}{v_{dc}} - \hat{\varphi}_1$, $\tilde{\varphi}_2 = \frac{2}{v_{dc}} - \hat{\varphi}_2$, and $\tilde{\varphi}_3 = \frac{2\lambda_q L}{v_{dc}} - \hat{\varphi}_3$, where $\hat{\varphi}_1$, $\hat{\varphi}_2$, and $\hat{\varphi}_3$ are the q -current adaptive laws.

The model-free adaptive controllers are designed as,

$$u_d = -k_d \theta_d + \hat{\delta}_1 i_q + \hat{\delta}_2 v_d - \hat{\delta}_3 e_{i_d}, \quad (6.13)$$

$$u_q = -k_q \theta_q - \hat{\varphi}_1 i_d + \hat{\varphi}_2 v_q - \hat{\varphi}_3 e_{i_q}, \quad (6.14)$$

where k_d and k_q is positive constants, $\hat{\delta}_1$, $\hat{\delta}_2$, $\hat{\delta}_3$, $\hat{\varphi}_1$, $\hat{\varphi}_2$, and $\hat{\varphi}_3$ are adaptive law to be design as follows.

Based on (6.13), the time-derivative along the system (6.11),

$$\begin{aligned} \dot{V}_d(\theta_d, i_q, v_d, e_{i_d}) &= \theta_d \dot{\theta}_d + \frac{v_{dc}}{2L} \sum_{i=1}^3 \frac{1}{\gamma_i} \tilde{\delta}_i \dot{\tilde{\delta}}_i \\ &= \theta_d \left(-w i_q - \frac{1}{L} v_d + \frac{v_{dc}}{2L} u_d + \lambda_d e_{i_d} \right) \\ &\quad - \frac{v_{dc}}{2L} \sum_{i=1}^3 \frac{1}{\gamma_i} \tilde{\delta}_i \dot{\hat{\delta}}_i \\ &= \frac{v_{dc}}{2L} \theta_d \left(-k_d \theta_d + \hat{\delta}_1 i_q + \hat{\delta}_2 v_d - \hat{\delta}_3 e_{i_d} \right) \\ &\quad + \frac{v_{dc}}{2L} \left(-\frac{2Lw}{v_{dc}} i_q - \frac{2}{v_{dc}} v_d + \frac{2\lambda_d L}{v_{dc}} e_{i_d} \right) \\ &\quad - \frac{v_{dc}}{2L} \sum_{i=1}^3 \frac{1}{\gamma_i} \tilde{\delta}_i \dot{\hat{\delta}}_i \end{aligned}$$

$$\begin{aligned}
&= -\frac{v_{dc}}{2L}k_d\theta_d^2 + \theta_d\frac{v_{dc}}{2L}\left[-(-\hat{\delta}_1 + \frac{2Lw}{v_{dc}})i_q\right. \\
&\quad \left.-(-\hat{\delta}_2 + \frac{2}{v_{dc}})v_d + (-\hat{\delta}_3 + \frac{2\lambda_d L}{v_{dc}})e_{i_d}\right] \\
&\quad -\frac{v_{dc}}{2L}\sum_{i=1}^3\frac{1}{\gamma_i}\tilde{\delta}_i\dot{\hat{\delta}}_i \\
&= -\frac{v_{dc}}{2L}\left(k_d\theta_d^2 + \tilde{\delta}_1 i_q \theta_d + \tilde{\delta}_2 v_d \theta_d\right. \\
&\quad \left.-\tilde{\delta}_3 e_{i_d} \theta_d\right) - \frac{v_{dc}}{2L}\sum_{i=1}^3\frac{1}{\gamma_i}\tilde{\delta}_i\dot{\hat{\delta}}_i.
\end{aligned} \tag{6.15}$$

To attain $\dot{V}_d(\theta_d, i_q, v_d, e_{i_d}) \leq 0$ of (6.15), then the adaptive law of the d -current can be designed as,

$$\dot{\hat{\delta}}_1 = -\gamma_1 i_q \theta_d, \dot{\hat{\delta}}_2 = -\gamma_2 v_d \theta_d, \dot{\hat{\delta}}_3 = \gamma_3 e_{i_d} \theta_d, \tag{6.16}$$

where $\gamma_1 > 0$, $\gamma_2 > 0$ and $\gamma_3 > 0$ are chosen as the gains of adaptive laws.

Substituting (6.16) into (6.15), the (6.15) can be rewritten,

$$\dot{V}_d(\theta_d, i_q, v_d, e_{i_d}) = -\frac{v_{dc}}{2L}k_d\theta_d^2, \tag{6.17}$$

when $k_d > 0$ is selected, then the $\dot{V}_d(i_q, v_d, e_{i_d}) \leq 0$.

Similar to previously discussed (6.15)-(6.17), the time-derivative of (6.12) can be gotten as,

$$\begin{aligned}
\dot{V}_q(\theta_q, i_d, v_q, i_{e_q}) &= -\frac{v_{dc}}{2L}(k_q\theta_q^2 - \tilde{\varphi}_1 i_d \theta_q + \tilde{\varphi}_2 v_q \theta_q - \tilde{\varphi}_3 e_{i_q} \theta_q) \\
&\quad -\frac{v_{dc}}{2L}\sum_{i=1}^3\frac{1}{\eta_i}\tilde{\varphi}_i\dot{\hat{\varphi}}_i.
\end{aligned} \tag{6.18}$$

The adaptive laws of q -current can be designed as

$$\dot{\hat{\varphi}}_1 = \eta_1 i_d \theta_q, \dot{\hat{\varphi}}_2 = -\eta_2 v_q \theta_q, \dot{\hat{\varphi}}_3 = \eta_3 e_{i_q} \theta_q. \tag{6.19}$$

Putting (6.19) into (6.18), the (6.18) can be represented,

$$\dot{V}_q(\theta_q, i_d, v_q, i_{e_q}) = -\frac{v_{dc}}{2L}k_q\theta_q^2, \tag{6.20}$$

when $k_q > 0$ is selected, then the $\dot{V}_q(\theta_q, i_d, v_q, i_{e_q}) \leq 0$.

Integrating both sides of (6.17), one gets,

$$\int_0^\infty \dot{V}_d(\theta_d, i_q, v_d, e_{i_d})d\tau = -\frac{v_{dc}}{2L}k_d \int_0^\infty \theta_d^2 d\tau \leq 0, \tag{6.21}$$

that is,

$$\frac{v_{dc}}{2L}k_d \int_0^\infty \theta_d^2 d\tau \leq V_d(0) - V_d(\infty). \tag{6.22}$$

Since $V_d(t) \geq 0$ and $\dot{V}_d(t) \leq 0$, it can be obtained that,

$$\frac{v_{dc}}{2L} k_d \int_0^\infty \theta_d^2 d\tau \leq V_d(0) - V_d(\infty) \leq V_d(0), \quad (6.23)$$

due to $\frac{v_{dc}}{2L} k_d > 0$, one can obtain the following inequality,

$$\int_0^\infty \theta_d^2 d\tau \leq \infty, \quad (6.24)$$

meanwhile, similarly (6.21)-(6.24), it can be also obtained that

$$\int_0^\infty \theta_q^2 d\tau \leq \infty, \quad (6.25)$$

That implies that $\theta_d \in \mathcal{L}_2$ and $\theta_q \in \mathcal{L}_2$, i.e., θ_d and θ_q are bounded. On the basis of (6.16) and (6.19), the adaptive laws $\hat{\delta}_1, \hat{\delta}_2, \hat{\delta}_3, \hat{\varphi}_1, \hat{\varphi}_2$, and $\hat{\varphi}_3$ are also bounded.

One can demonstrate the uniform continuity of $\dot{V}_d(\theta_d, i_q, v_d, e_{i_d})$ by analyzing its derivative to determine if it is bounded. Then $\ddot{V}_d(\theta_d, i_q, v_d, e_{i_d})$ is computed,

$$\begin{aligned} \ddot{V}_d(\theta_d, i_q, v_d, e_{i_d}) &= -\frac{v_{dc}}{L} k_d \theta_d \dot{\theta}_d \\ &= -\frac{v_{dc}}{L} k_d \theta_d \left(-w i_q - \frac{1}{L} v_d \right. \\ &\quad \left. + \frac{v_{dc}}{2L} u_d + \lambda_d e_{i_d} \right) \\ &= -\frac{v_{dc}}{L} k_d \theta_d \left(-w i_q - \frac{1}{L} v_d \right. \\ &\quad \left. + \frac{v_{dc}}{2L} [-k_d \theta_d + \hat{\delta}_1 i_q \right. \\ &\quad \left. + \hat{\delta}_2 v_d - \hat{\delta}_3 e_{i_d}] + \lambda_d e_{i_d} \right). \end{aligned} \quad (6.26)$$

Since $\theta_d, i_q, v_d, e_{i_d}, \hat{\delta}_1, \hat{\delta}_2$ and $\hat{\delta}_3$ are bounded by the virtue that $\dot{V}(\theta_d, i_q, v_d, e_{i_d}) \leq 0$, therefore, $\ddot{V}_d(\theta_d, i_q, v_d, e_{i_d})$ is bounded. Thus, $\dot{V}_d(\theta_d, i_q, v_d, e_{i_d})$ is uniformly continuous. It follows from the Barbalat's lemma that $\dot{V}_d(\theta_d, i_q, v_d, e_{i_d}) \rightarrow 0$, as a result, $\theta_d \rightarrow 0$ as $t \rightarrow \infty$. Likewise, one can also get $\dot{V}_q(\theta_q, i_d, v_q, e_{i_q}) \rightarrow 0$, hence, $\theta_q \rightarrow 0$ as $t \rightarrow \infty$. ■

Based on *Theorem 6.3.1*, θ_d and θ_q can asymptotically converge to zero, that is, $i_d \rightarrow i_d^*$ and $i_q \rightarrow i_q^*$. As a result, MFAC is robust and has a fast convergence rate against disturbances.

Remark 6.3.2 *In contrast to model-based control frameworks, the proposed MFAC operates independently of system models (6.1)-(6.2), thereby eliminating issues caused by model distortion, unmodeled dynamics, model variations, and parameter mismatches. In this way, the MFAC control is considered safer and more reliable for converter applications.*

6.3.3 Voltage Regulation Loop

An MFAC method is tasked with accurate regulating the dc-voltage v_{dc} to track its reference value in presence of model distortion, unmodeled dynamic, parameter mismatches, and unknown external disturbance (see Figure 6.1-C.1).

Assumption 6.3.2 *The dq-current tracking dynamics of (6.1)-(6.2) are much faster than the dc-link voltage dynamic of (6.3).*

According the Assumption 6.3.2, the dynamic of dc-link voltage of (6.3) can be rewritten by multiplying v_{dc} . This yields,

$$v_{dc}\dot{v}_{dc} = \frac{v_{dc}}{C}(u_d i_d + u_q i_q - 2v_{dc} i_L). \quad (6.27)$$

Considering that $(\frac{v_{dc}^2}{2})' = v_{dc}\dot{v}_{dc}$, one can write (6.27) as,

$$v_{dc}\dot{v}_{dc} = \frac{v_{dc}}{C}(\mathbf{u}_{dq}^T \mathbf{i}_{dq} - 2v_{dc} i_L). \quad (6.28)$$

Taking into account that the voltage generated in the converter terminals is defined as $\mathbf{v}_{VSI} = \frac{1}{2}v_{dc}\mathbf{u}_{dq}$, and that definition of the power consumed in the dc-link is $p_L = v_{dc}i_L$, then, (6.28) can be rewritten as,

$$v_{dc}\dot{v}_{dc} = \frac{2}{C}(\mathbf{v}_{VSI}^T \mathbf{i}_{dq} - p_L). \quad (6.29)$$

Considering the Instantaneous Power Theory, one can define the instantaneous active power going into the power converter as $p_{VSI} = \mathbf{v}_{VSI}^T \mathbf{i}_{dq}$, then, (6.29) can be rewritten as,

$$v_{dc}\dot{v}_{dc} = \frac{2}{C}(p_{VSI} - p_L), \quad (6.30)$$

Note that in the dc-link both capacitors are in series, so one can define the equivalent capacitance of the dc-link as:

$$C_{eq} = \frac{C}{2}, \quad (6.31)$$

where C_{eq} is considered as an unknown parameter. With yields the final expression for the energy balance as:

$$v_{dc}\dot{v}_{dc} = \frac{1}{C_{eq}}(p_{VSI} - p_L). \quad (6.32)$$

One can define the new variable $z \triangleq \frac{v_{dc}^2}{2}$, that allows one to write the power consumed in the load as $p_L = \frac{2z}{R_L}$. R_L is considered as an unknown external disturbance.

Next, considering that the instantaneous active power going into the power converter comes from the grid, the instantaneous active power reference must be equal to p_{VSI} to keep the system balanced. So the dynamic of the dc-link capacitor is represented by:

$$\dot{z} = \frac{p^*}{C_{eq}} - \frac{2z}{R_L C_{eq}}, \quad (6.33)$$

p^* is viewed as the desired active power. Then according to (6.33), one can get

$$\dot{z} = \frac{2}{C}p^* - \frac{4}{R_L C}z, \quad (6.34)$$

where C is the capacitance, considered an unknown parameter. The dc-link voltage regulation error is defined as,

$$e_z = z^* - z, \quad (6.35)$$

To reduce the dc-voltage steady-state error, the integral item of (6.35) is introduced as,

$$\theta_z = e_z + \lambda \int_{t_0}^t e_z d\tau, \quad (6.36)$$

where $\lambda > 0$ is set. The time-derivative of (6.36) is

$$\dot{\theta}_z = -\frac{2}{C}p^* + \frac{4}{R_L C}z + \lambda e_z. \quad (6.37)$$

Similar to *Theorem 6.3.1*, the voltage model-free adaptive control of *Theorem 6.3.2* can be obtained as follows.

6.3.3.1 MFAC Design

Theorem 6.3.2 Consider the voltage error closed-loop system (6.37) with Assumption ???. Given parameters $k > 0$, $\lambda > 0$, $\mu_j > 0$ ($j \in \{1, 2\}$), if there exist a Lyapunov function

$$V_z(\theta_z, z, e_z) = \frac{1}{2}(\theta_z^2 + \frac{1}{2C} \sum_{j=1}^2 \frac{1}{\mu_j} \tilde{\zeta}_j^2), \quad (6.38)$$

that are positive definite, and only the adaptive laws $\dot{\hat{\zeta}}_1 = \mu_1 z \theta_z$, $\dot{\hat{\zeta}}_2 = \mu_2 e_z \theta_z$ are designed such that the derivatives $\dot{V}_z(\theta_z, z, e_z) \leq 0$, then the designed MFAC controllers

$$p^* = k_z \theta_z + \hat{\zeta}_1 z + \hat{\zeta}_2 e_z, \quad (6.39)$$

guarantee that the (6.37) stability and assure the voltage regulation θ_z converge to zero as $t \rightarrow \infty$ in the presence of uncertainties and disturbances.

The stability proof of *Theorem 6.3.2* is similar to the *Theorem 6.3.1*. Omitted here for space. Consequently, it can be inferred that the dc-voltage regulation error $e_z \rightarrow 0$ as $t \rightarrow \infty$, i.e., $z \rightarrow z^*$. Hence, the proposed MFAC method can effectively regulate the dc-voltage v_{dc} to follow its reference v_{dc}^* .

6.3.4 Voltage Balancing Loop

Due to the emphasis placed on the current-voltage control, a PI controller is adopted for balancing (two) capacitor voltages, maintaining the voltage difference v_{dc}^o at zero.

6.3.4.1 PI Design

A PI control method is adopted on the basis of (6.4), that is,

$$u_o = k_{bp}(\tilde{v}_{dc}^o) + k_{bi} \int (\tilde{v}_{dc}^o) dt. \quad (6.40)$$

Here, u_o represents the control signal. $k_{bp} > 0$ and $k_{bi} > 0$ are control gains. $\tilde{v}_{dc}^o = v_{dc}^o - v_{dc}^{o*}$, where the reference value v_{dc}^{o*} is set to zero.

6.3.5 Brief Review on Particle swarm optimization (PSO)-based-MFAC Optimization Process

Optimizing coefficients (gains) for the MFAC method poses a challenge due to the inherent complexity of tuning process ¹. Here, the PSO algorithm [181] is employed as an effective solution to search and optimize the current MFAC gains: $k_d, k_q, \lambda_d, \lambda_q, \delta_i, \psi_i$ ($i \in \{1, 2, 3\}$), and voltage MFAC gains: k_z, λ, ζ_j ($j \in \{1, 2\}$), thereby enhancing current/voltage control performance and robustness (see Figure 6.1-C.0).

PSO is a swarm-based stochastic method based on the way schooling fish and birds search for targets or food. Each swarm member is called a particle with the solution to the problem. At the outset, the algorithm initializes particles with random speed and position. The fitness of each member is assessed for voltage and current tracking using the fitness function,

$$W_{\min} = \underbrace{\varrho \int_{t_0}^t t \theta_d^2 d\tau + \varrho_1 \int_{t_0}^t t \theta_q^2 d\tau}_{\text{current}} + \underbrace{\theta_z^2 + \varrho_2 t_r + \varrho_3 M_p}_{\text{voltage}}, \quad (6.41)$$

where ϱ and ϱ_1 are weighting factors of current errors θ_d and θ_q , respectively. t_r is the rise time of dc-link voltage and ϱ_2 is the weighting factor of t_r , and M_p is the overshoot of the voltage and ϱ_3 is the weighting factor of M_p .

Remark 6.3.3 *In the optimization process, the parameters t_r and M_p , denoting the rise time and overshoot of the voltage outer loop's step response, respectively, are nonlinearly mapped to the adaptive rate parameters. This process formulates their relationship as a black-box optimization problem, enabling parameter adjustment based solely on system response data independent of any internal system model. By analyzing the system's responses to variations in control parameters, we define a cost function (6.41) to iteratively optimize the adaptive rate ratios, thereby deriving a set of adaptive rules that meet predefined performance criteria.*

Based on the minimum fitness value $W_{\min}$, the global best particle ($gbest$) is chosen. The movement of particle is influenced by three factors: Experience of $gbest$; Experience of $pbest$; Inertia. The particles in the PSO algorithm updated their velocity and position in the following formulas:

$$\begin{aligned} v_{m,g}^{k+1} &= \bar{\omega} v_{m,g}^k + c_1 r_1 (pbest_{m,g}^k - x_{m,g}^k) + c_2 r_2 (gbest_g^k - x_{m,g}^k) \\ x_{m,g}^{k+1} &= x_{m,g}^k + v_{m,g}^{k+1} \end{aligned} \quad (6.42)$$

where $m = 1, 2, \dots, M$, $g = 1, 2, \dots, G$. To evaluate the PSO method in selecting and optimizing the MFAC parameters, the PSO parameters are presented in Table 6.2.

¹Based on *Theorems 6.3.1* and *6.3.2*, in the current and voltage MFAC systems, selecting control gains presents a significant challenge, where manual adjustment becomes laborious or ineffective, resulting in poor performance and limited adaptability. PSO possesses profits in optimized parameters, likewise: 1) exploring high-dimensional spaces; 2) good search capabilities; and 3) reducing labor costs. Thus, PSO is a highly effective tool in this study for optimizing control gains in these systems, ensuring robust performance and adaptability across various operating scenarios.

Table 6.2: PSO-MFAC Parameters

Parameters	Indexes	Values
M	Population of Particles	200
G	Dimensional space	9
K	Maximum of iterations, where $\{1 \cdots k \cdots K\}$	50
$\bar{\omega}$	Inertia weight factor	0.7
c_1, c_2	Acceleration constant	1
r_1, r_2	Random number	$[0, 1]$
v_{max}	Maximum particle velocity	2
v_{min}	Minimum particle velocity	-2
U	Particle Position Maximum	10
L	Particle Position Minimum	1e-10
$v_{m,g}^k$	Velocity of particle m at iteration k ,	-
$x_{m,g}^k$	Current position of particle m at iteration k	-
$pbest_{m,g}^k$	$pbest$ of particle m	-
$gbest_g^k$	$gbest$ of the group	-

Algorithm 1: PSO-MFAC Optimization Procedure**Input** : Bounds of feasible operating regions**Output** : Optimized $W_{\min}$ **Begin** :

- 1 Set PSO parameters (particle, run, iteration numbers, etc.)
- 2 **for** each particle m **do**
- 3 Initialize velocity v_m and position x_m for particle m ;
- 4 Evaluate particle m and set $pbest_m = x_m$;
- 5 $gbest = \min\{pbest_m\}$
- 6 **while** $k \leq K$ **do**
- 7 **for** $m=1$ to M **do**
- 8 Update the velocity and position of each particle m ;
- 9 Recalculate fitness function
- 10 **if** $fitness(x_m) < fitness(pbest_m)$ **then**
- 11 $pbest_m = x_m$;
- 12 **if** $fitness(pbest_m) < fitness(gbest)$ **then**
- 13 $gbest = pbest_m$;
- 14 Print $gbest$;

A pseudo-code searching procedure of the PSO-MFAC is provided as **Algorithm 1**. Here, the proposed PSO-optimized MFAC control with the aim of the $W_{\min}$ is evaluated. According to the principle of PSO algorithm, Fig. 6.2 illustrates the variation of $gbest$ concerning the maximum number of iterations 50, with varying population. It can be seen that the $gbest$ will be stabilized

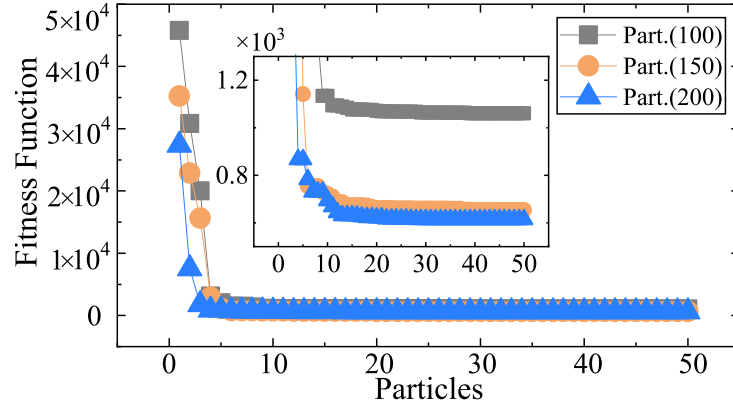


Figure 6.2: The fitness function $W_{\min}$ with the different number of particles (part.) under iteration number=50, when Voltage reference $v_{dc}^*=160\text{V}$, dc-load $R_L = 30\Omega$ and reactive power reference $q^* = 0$.

during the increase in the particles². Notably, increasing the number of particles improves the performance of the algorithm while increasing computation time, so it needs to make trade-offs; 200 particles are selected here. From Fig. 6.3, the PSO-MFAC can yield superior voltage and dq -current tracking for the 3L-NPC converter. when particles are 200 and iterations are 50.

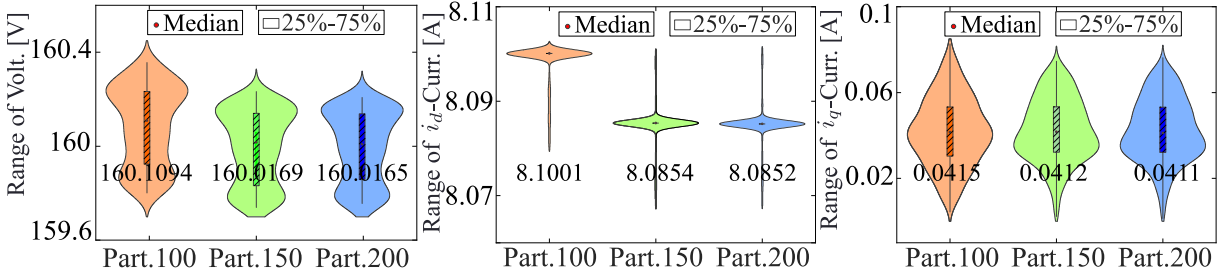


Figure 6.3: The range of dc-voltage and dq -current by particle (part.) of population = 100, 150, and 200 when iteration = 50 with $v_{dc}^*=160\text{V}$, $R_L = 30\Omega$, and $q^* = 0$.

6.4 Experimental Results

In this section, the effectiveness of the proposed PSO-MFAC control scheme is evaluated on a laboratory-built prototype of a 3L-NPC power converter, comparing it with the ANN-PI and STA control schemes, as depicted in Figure 6.4. The experiments are conducted fairly and objectively, quantitatively and qualitatively, to ensure the assessments are accurate, reliable, and practical. Parameters of the experimental platform are listed in Table 6.3, and the proposed PSO-MFAC, ANN-PI, and STA control parameters are shown in Table 6.4.

²The impact of particle population on PSO is discussed as: As particle swarms are multiplied, the search time for finding the global optimum is prolonged, while increasing the number of particle swarms is beneficial in locating the global optimum. Consequently, this study evaluates the performance and computational time of the particle swarm in the optimal parameter configuration for the MFAC, selecting 200 as the swarm population and capping the number of iterations to 50.

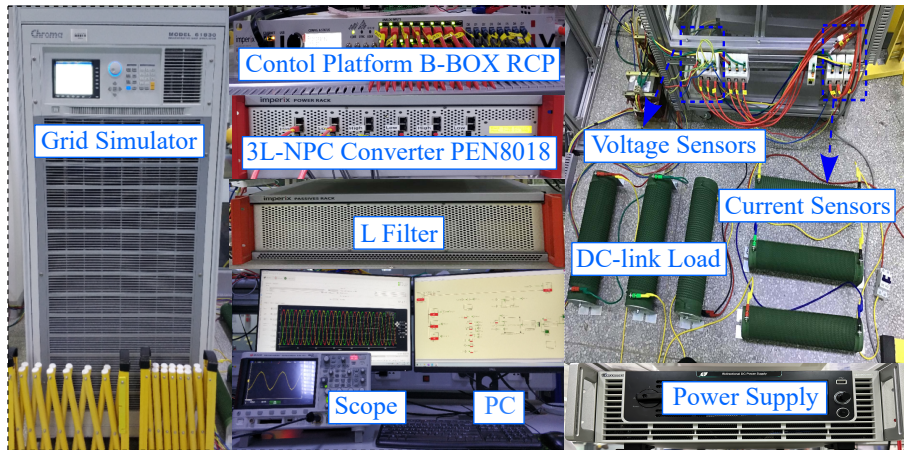


Figure 6.4: The lab-constructed 3L-NPC power converter prototype of Chap. 4.

Table 6.3: Parameters Used For Experiment

Description	Parameter	Value	Unites
Sampling frequency	f_s	10	kHz
Switching frequency	f_{sw}	10	kHz
Grid frequency	f	50	Hz
Grid phase voltage	v_{abc}	50	V
Phase inductor	L	5	mH
Dc-link capacitor	C	1551	uF
Load resistance	R_L	$30 \leftrightarrow 40$	Ω
Reference output voltage	v_{dc}^*	$140 \leftrightarrow 160$	V
Reactive power	q^*	$-300 \leftrightarrow 300$	Var

6.4.1 An Overall Evaluation Comparison

The overall assessment of the 3L-NPC power converter system is very important for validating the effectiveness of the proposed PSO-MFAC scheme. This test compares three control schemes, PSO-MFAC, ANN-PI, and STA, to determine their performance in regulating voltage and tracking current. The control performances for all the above methods are shown in Figure 6.5. It is evident that the proposed PSO-MFAC scheme outperforms the ANN-PI and STA control schemes regarding faster transient response speeds, better steady responses, less over-/under-shoot of the dc-link voltage when varying, and lower grid current THDs when changing reactive power.

6.4.2 Comparative Voltage Regulation Performance

The effectiveness and superiority of the proposed PSO-MFAC method are verified by comparing STA and ANN-PI control schemes through the dc-link voltage changes, as shown in Figure 6.6. Utilizing PI control for the current and the balancing loops ensures fairness in the voltage

Table 6.4: Controller Parameters

Control Scheme	Voltage Regulation Loop	Power Tracking Loop	Voltage Balancing Loop
Proposed PSO-MFAC	$k_z = 0.14$ $\lambda = 7.1e-4$ $\zeta_1 = 2.2e-8$ $\zeta_2 = 3.1e-8$	$k_d = k_q = 0.89$ $\lambda_d = \lambda_q = 1.22$ $\gamma_1 = \eta_1 = 0.32$ $\gamma_2 = \eta_2 = 0.12$ $\gamma_3 = \eta_3 = 0.34$	$k_{bp} = 1.5$ $k_{bi} = 0.15$
ANN-PI based [179]	$k_{vp} = 4e1$ $k_{vi} = 8e3$ $\alpha_{z1} = 5e-10$ $\alpha_{z2} = 4e-9$	$k_{dp} = k_{qp} = -5.5e-3$ $k_{di} = k_{qi} = -5e-5$ $\alpha_{d3} = \alpha_{q4} = 5e-4$ $\alpha_{d5} = \alpha_{q6} = 4e-2$ $\alpha_{d7} = \alpha_{q8} = 1e-5$	$k_{bp} = 1$ $k_{bi} = 0.1$
STA based [72]	$\iota_z = 1.3e3$ $\rho_z = 2.9e6$	$\iota_d = \iota_q = 7e3$ $\rho_d = \rho_q = 7e5$	$k_{bp} = 1.2$ $k_{bi} = 0.2$

Table 6.5: Comparisons of PSO-MFAC, ANN-PI, and STA Methods

Voltage Regulation Index (Figure 6.6: 0.25s-0.3s)		PSO-MFAC Method	ANN-PI Method	STA Method
Index v_{dc} (V)	UST ¹	0.05	0.058	0.06
	OST ²	0.047	0.058	0.06
	ME ³	0.0155	0.0172	0.0192
	MSE ⁴	0.2417	0.2742	0.2974
	VE ⁵	0.0583	0.0660	0.0750
Index v_{dc} (V)	PSO-MFAC vs ANN-PI		PSO-MFAC vs STA	
	Enhancement%			
UST ¹		13.79%		16.67%
OST ²		18.97%		21.67%
ME ³		9.88%		19.27%
MSE ⁴		11.85%		18.73%
VE ⁵		11.67%		22.27%

1: Under-settling Time (s) 2: Over-settling Time (s); 3: Mean Error; 4: Mean Square Error; 5: Variance Error.

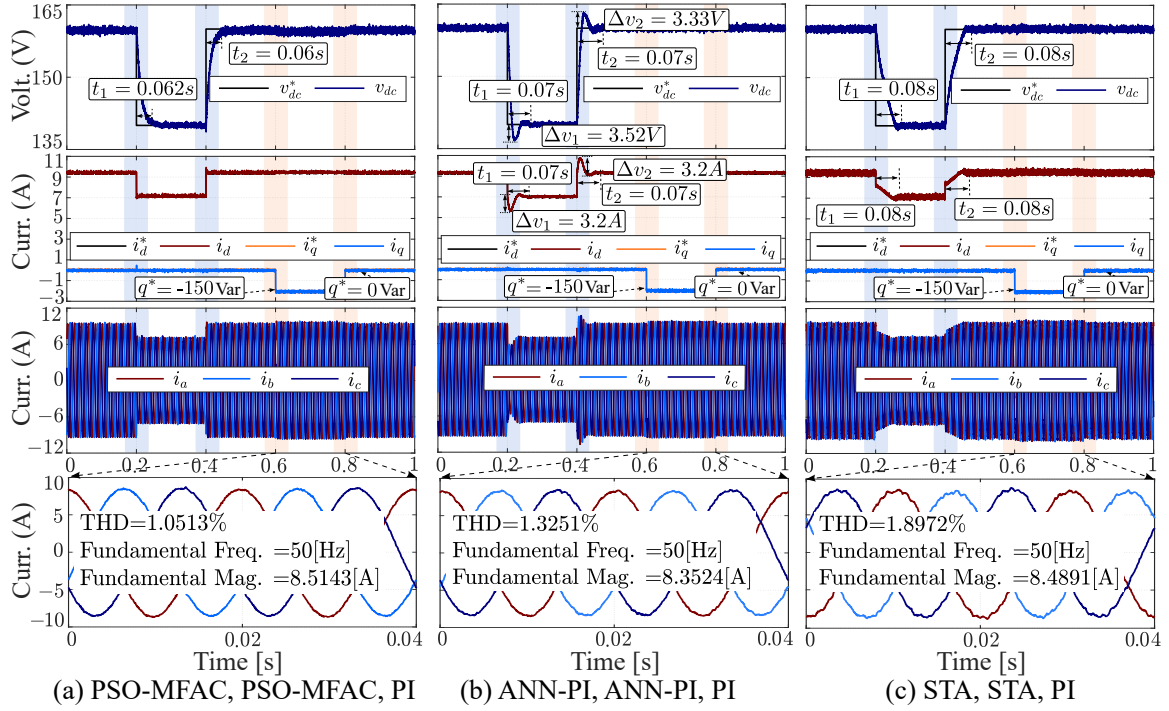


Figure 6.5: An overall evaluation comparison under v_{dc}^* is from 160V $\rightarrow$ 140V $\rightarrow$ 160V, q^* is from 0 $\rightarrow$ 150Var $\rightarrow$ 0 with $R_L = 30\Omega$.

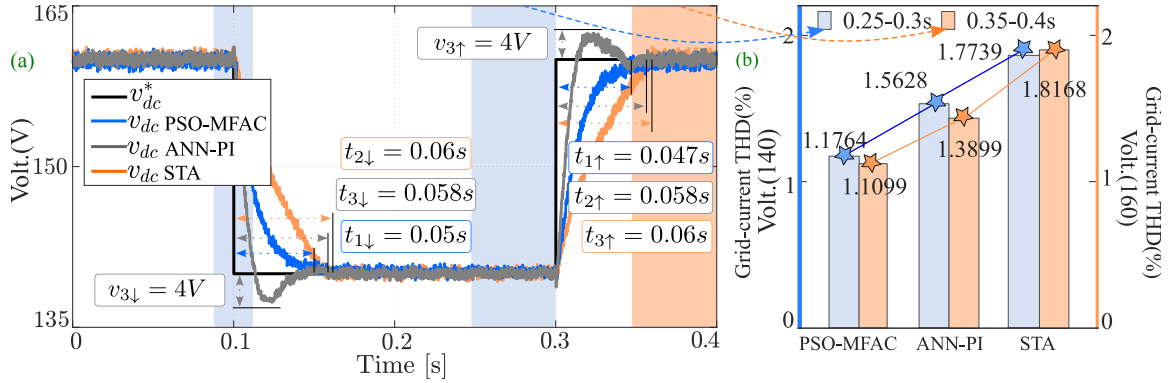


Figure 6.6: The voltage performance comparison: (a) (Transient-state) v_{dc}^* from 160V to 140V back to 160V set at 30Ω with $q^* = 0$; (b) (Steady-state) grid current THD under 160V and 140V, respectively.

regulation comparisons, as shown in Figures 6.6, it has been demonstrated that the proposed control PSO-MFAC method can produce better transient and steady-state voltage regulation performance. A detailed analysis of the performance of the proposed method shows that it exhibits smaller grid current THD, over-/under-shoot voltage, and faster convergence properties than STA and ANN-PI methods. Once again, the above point of view is borne out by the comparison presented in Table 6.5.

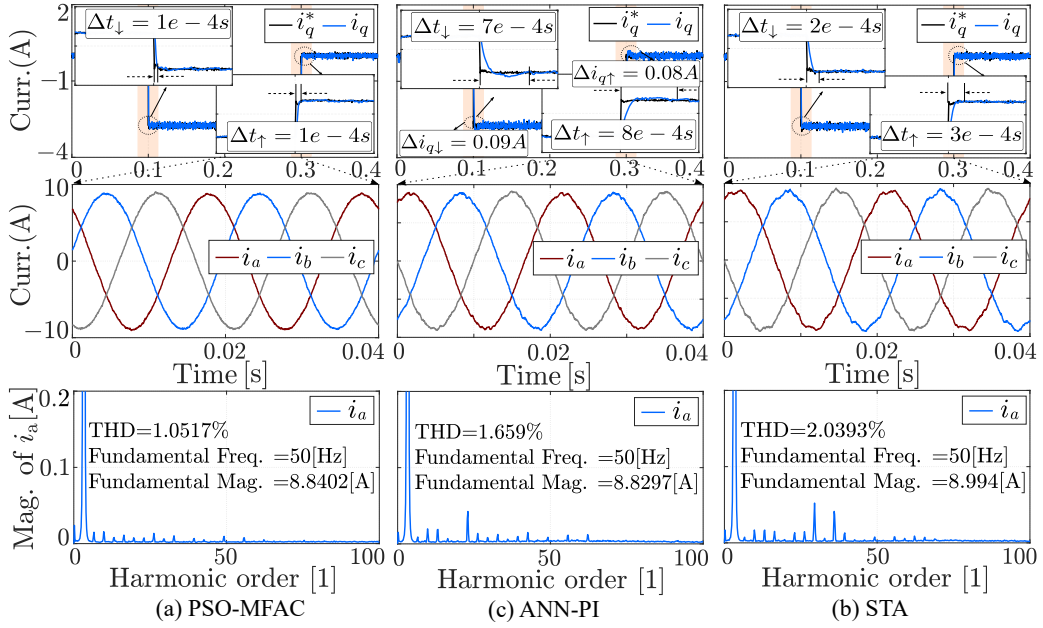


Figure 6.7: The reactive current performance comparison as q^* is from 0 to 200Var back to 0, and the grid-current THD comparison when the $q^*=200\text{Var}$ (steady-state), under the condition of $R_L=30\Omega$ and $v_{dc}^* = 160\text{V}$.

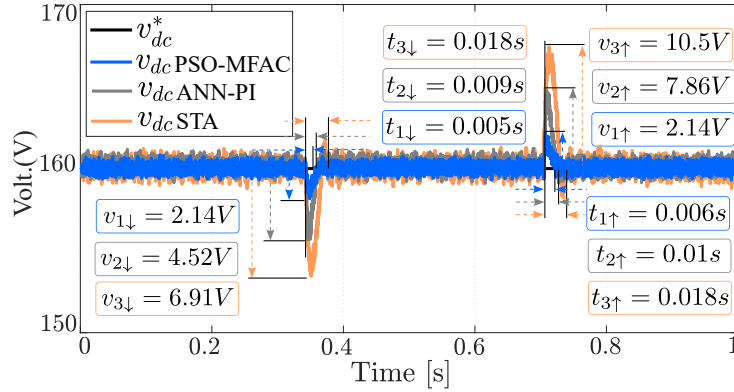


Figure 6.8: Comparative voltage regulation performance under external dc-load variation: R_L from 30Ω to 40Ω back to 30Ω , $v_{dc}^* = 160\text{V}$, and $q^* = 0$.

6.4.3 Comparative Current Tracking Performance

A comparison is made with ANN-PI and STA control methods to evaluate the performance of the PSO-MFAC method in current tracking and grid current quality. The reactive power reference is obtained by adjusting the q -current reference i_q^* . The operating condition is as follows: the PI control method is utilized for both the voltage regulation and balancing voltage loops. The control above methods are utilized to track the i_q^* , as depicted in Figure 6.7. From Figure 6.7, the proposed PSO-MFAC method achieves superior reactive current (i_q) control over STA and ANN-PI methods. In particular, PSO-MFAC exhibits a faster convergence speed for i_q without overshooting or undershooting under a 200Var reactive power load while maintaining high-quality sinusoidal waves in grid-side currents. From the perspective of grid-side current

Table 6.6: Comparisons of PSO-MFAC, ANN-PI, and STA Methods

Current Tracking Index (Figure 6.7: 0.25s-0.3s)		PSO-MFAC Method	ANN-PI Method	STA Method
Indix i_q (A)	UST ¹	1e-4	7e-4	2e-4
	OST ²	1e-4	8e-4	3e-4
	THD	1.0517%	1.6590%	2.0393%
	ME ³	0.0006	0.0019	0.0019
	MSE ⁴	0.0808	0.0950	0.0907
	VE ⁵	0.0065	0.0090	0.0083
Indix i_q (A)	PSO-MFAC vs ANN-PI		PSO-MFAC vs STA	
	Enhancement%			
UST ¹	85.71%		50%	
OST ²	87.5%		66.67%	
THD	36.61%		48.43%	
ME ³	68.42%		68.42%	
MSE ⁴	14.95%		10.92%	
VE ⁵	27.78%		21.69%	

1: Under-settling Time (s) 2: Over-settling Time (s); 3: Mean Error; 4: Mean Square Error; 5: Variance Error.

quality, the THD of i_a obtained by the proposed PSO-MFAC (1.0517%) is lower than that of the ANN-PI (1.659%) and STA (2.0393%) methods. Performance indices for reactive power and grid-side current are presented in Table 6.6. The proposed PSO-MFAC method outperforms others regarding grid-side current quality and current tracking performance.

6.4.4 Robustness Evaluation with External Disturbance

As depicted in Figure 6.8, comparisons are conducted among the proposed PSO-MFAC, ANN-PI, and STA methods in order to validate the robustness of the above methods to external disturbances. The current tracking and voltage balancing loops utilized the PI method under the operational conditions. From Figure 6.8, it is evident that the proposed method outperforms the ANN-PI and STA methods in terms of voltage regulation, particularly when external disturbances fluctuate ($30\Omega \rightarrow 40\Omega$, then $40\Omega \rightarrow 30\Omega$), resulting in smaller voltage amplitude variations and faster convergence rates.

6.4.5 Robustness Evaluation under Parameter Mismatches

The robustness of the proposed PSO-MFAC is evaluated parameter mismatches by comparing it to the STA and ANN-PI methods, considering deviations from the nominal inductance (L) and capacitance (C). This evaluation considers three scenarios for L and C : 50% L/C , 100%

L/C , and $150\% L/C$. For inductance mismatches, the current tracking loop is controlled using PSO-MFAC, STA, and ANN-PI methods, respectively, and the voltage regulation and balancing loops are controlled using ANN-PI and PI methods. Under capacitance mismatches, the voltage tracking loop employs PSO-MFAC, ANN-PI, and STA control methods for tuning, respectively, while current regulation and voltage balancing loops are handled by ANN-PI and PI control methods.

Table 6.7: Comparisons of PSO-MFAC, STA and ANN-PI Methods

Voltage Regulation Index (Fig. 6.8: 0.5s-0.6s)		PSO-MFAC Method	ANN-PI Method	STA Method
Indix v_{dc} (V)	UST ¹	0.005	0.009	0.018
	UV ²	2.14	4.52	6.91
	OST ³	0.006	0.01	0.018
	OV ⁴	2.14	7.86	10.5
	ME ⁵	2.036e-4	3.8083e-4	4.5009e-4
	MSE ⁶	0.0838	0.0889	0.0944
	VE ⁷	0.0070	0.0079	0.0089
Indix v_{dc} (V)	PSO-MFAC vs ANN-PI		PSO-MFAC vs STA	
	Enhancement(%)			
UST ¹	44.44%		72.22%	
UV ²	52.65%		69.03%	
OST ³	40%		66.67%	
OV ⁴	72.77%		79.62%	
ME ⁵	46.54%		54.76%	
MSE ⁶	5.74%		11.23%	
VE ⁷	11.39%		21.35%	

1 Under-settling Time(s); 2 Under-voltage(V); 3 Over-settling Time(s); 4 Over-voltage(V); 5 Mean Error; 6 Mean Square Error; 7 Variance Error.

The experimental results of robustness evaluation under capacitance and inductance mismatches are shown in Figures 6.9 and 6.10. The proposed PSO-MFAC method achieves lower grid-side current THD value (between 1.1746% and 1.7555%), outperforming both the ANN-PI (1.2186% $\rightarrow$ 2.4829%) and STA (1.4993% $\rightarrow$ 2.9324%) methods. These outcomes indicate that the PSO-MFAC method provides good robustness under conditions of inductance and capacitance parameter mismatches, as depicted in Figure 6.9. The voltage root mean square (RMS) value of the proposed PSO-MFAC method is within the range 0.1186 $\rightarrow$ 0.2672, which indicates that the reference voltage is more closely aligned with actual voltage than those of the ANN-PI (RMS value: 0.2634 $\rightarrow$ 0.2921) and STA (RMS value: 0.2315 $\rightarrow$ 0.3939) methods, thus providing better parameter robustness against parameter mismatches (inductances and capacitance) as illustrated in Figure 6.10 (a), (b), and (c). To assess the reactive current tracking process against the inductance mismatch capacity, the proposed PSO-MFAC, ANN-PI, and STA control methods are evaluated experimentally, as shown in Figures 6.10 (d), (e), and (f). Accordingly, the proposed PSO-MFAC is strong robust against parameter mismatches due to a smaller current RMS value (range: 0.0245 $\rightarrow$ 0.1173), while the range power RMS values for

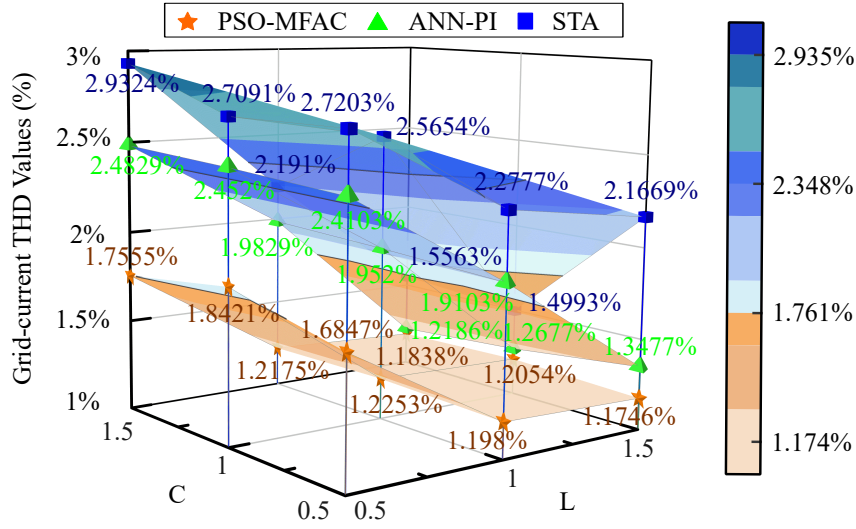


Figure 6.9: THD of grid-side current under inductance and capacitance mismatches under v_{dc}^* is 160V, R_L is 30Ω , and $q^* = 100\text{Var}$.

ANN-PI and STA are $(0.0515 \rightarrow 0.1652)$ and $(0.0311 \rightarrow 0.1705)$, respectively. Consequently, the proposed PSO-MFAC provides an anti-disturbance capability against capacitance and inductance parameter mismatches.

6.4.6 Versatile Multi-scenario Performance Assessment

To assess the versatility and adaptability of the proposed PSO-MFAC method in multi-scenario, this test introduces a new experimental configuration. In this setup, the dc-loads are replaced with a simulated dc-source, while a 3L-NPC converter performs the role of an interlinking converter, enabling ac-dc and dc-ac bidirectional power conversion.

In the multi-scenario performance evaluation of the converter, the proposed PSO-MFAC method is validated using current tracking and grid-side current THD as metrics, and its performance is compared with that of ANN-PI and STA methods. The test setup involves: 1) ac-dc conversion mode: the reactive power injection q^* is from 0 to 300 Var and the active power injection p^* from 450 W to 580 W, as presented in Figure 6.11.a-c, and 2) ac-dc conversion mode: q^* is from 0 to 300 Var, and p^* is from -450 W to -580 W, as illustrated in Figure 6.11.d-f. Compared to ANN-PI and STA control methods, the PSO-MFAC approach exhibits faster current tracking during active and reactive power injection in both ac-dc and dc-ac modes, reducing the ripple in power current. The PSO-MFAC presents a lower grid current THD during reactive power injection. Consequently, the proposed PSO-MFAC achieved these excellent performance results despite various factors, such as parameter mismatch, noise disturbance, and modeling errors in the multi-scenario evaluation.

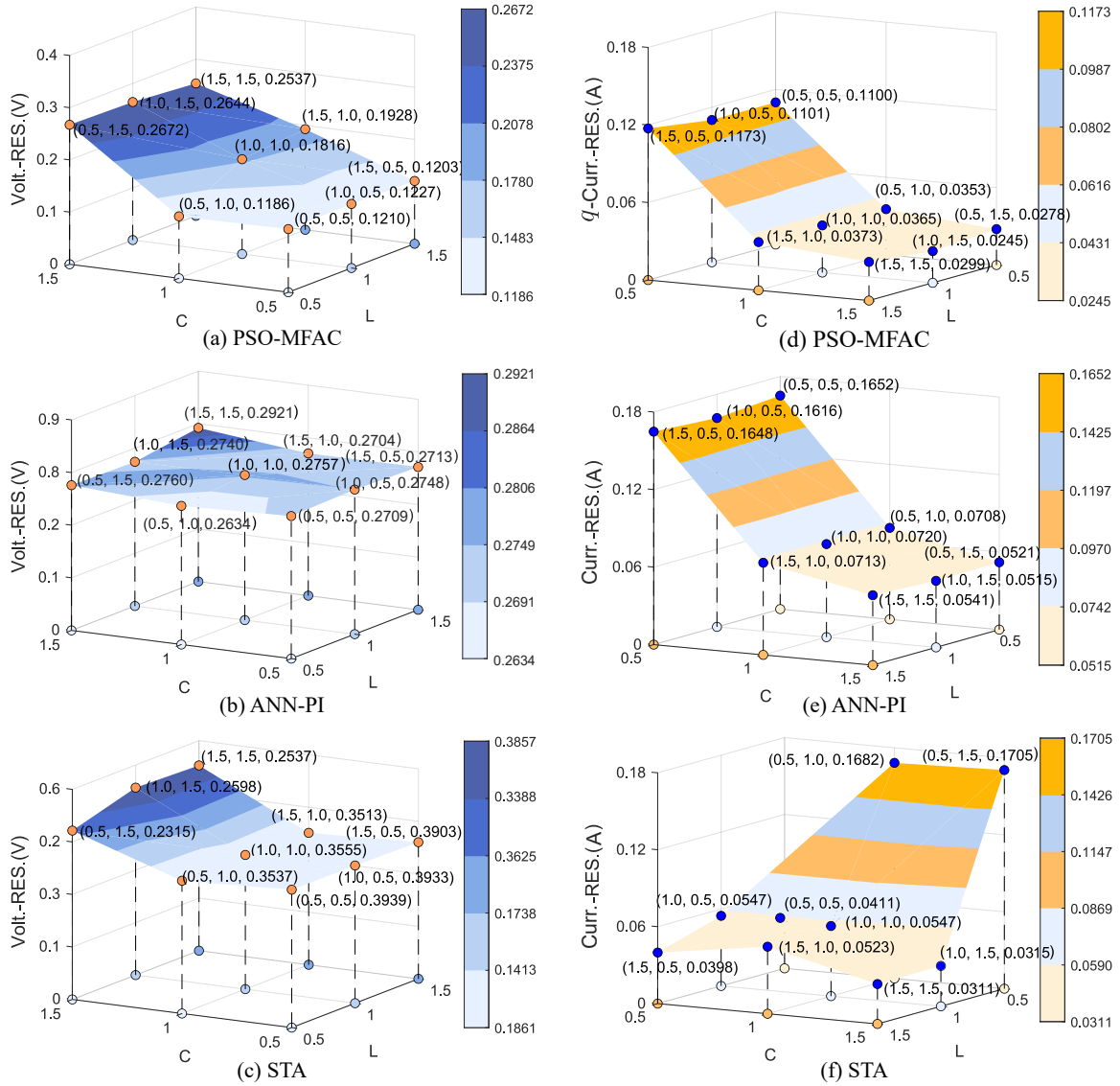


Figure 6.10: Parameter mismatches comparison (steady-state) under v_{dc}^* is 160V and R_L is 30Ω and $q^* = 100$ Var. RMS of the dc-link voltage under parameter mismatches (a) PSO-MFAC, (b) ANN-PI, and (c) STA; RMS of the q -current under parameter mismatches: (d) PSO-MFAC, (e) ANN-PI and (f) STA.

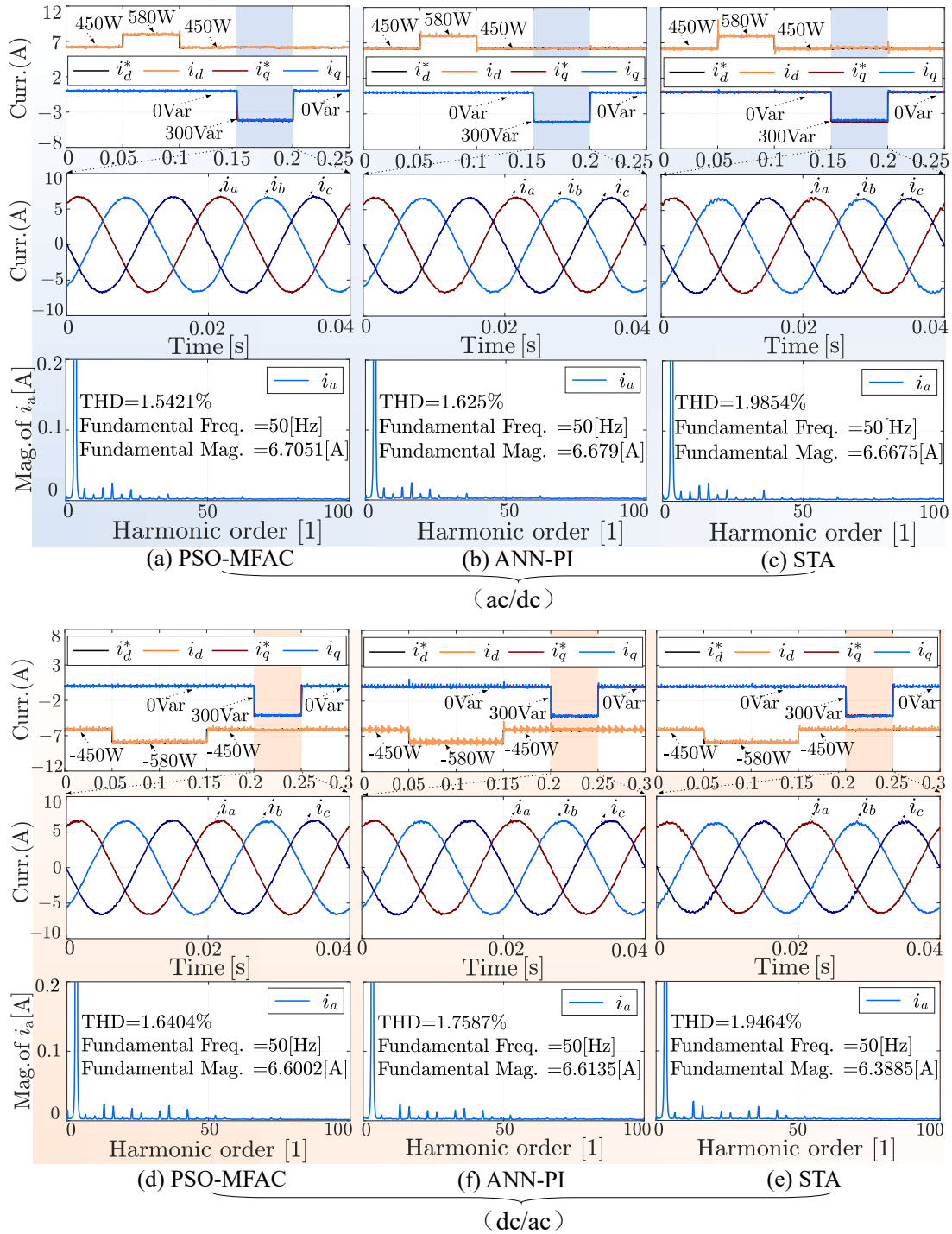


Figure 6.11: Comparative power tracking performance with q^* from 0 to 300Var back to 0, $v_{dc}^*=160\text{V}$, and $R_L=30\Omega$.

6.5 Summary

This chapter presented a data-driven MFAC for the 3L-NPC power converter system in micro-grids, effectively improving voltage and current quality without any system parameters while guaranteeing system stability, robustness, and flexibility. A data-driven MFAC was implemented to track voltage and current accurately by designing adaptive laws based on Lyapunov theory, which ensured self-adaptability to variations in dynamics, parameters, and disturbances, multi-scenarios. To facilitate access to adaptive law gains, the data-driven MFAC used a PSO algorithm, a metaheuristic optimization algorithm, to automatically determine gains, reducing manual effort. Finally, experimental results confirmed that the proposal outperformed both ANN-PI and STA control schemes in terms of dynamic response, THDs, and robustness. Specifically, overall experiment results show that the proposed control scheme:

- 1) eliminates over-/under-shoot voltage while improving over-/under-settling time by 11.43% ~ 14.29%, 22.25% ~ 25% respectively;
- 2) enhances ultimately current convergence speed (no over-/under-shoot in dq-current), and improves grid current THD by 20.66%, 44.59%, respectively; compared to ANN-PI and STA control schemes.

CHAPTER 7

Conclusions and Outlooks

This dissertation is carried out mainly to address the key issues under multi-disturbances, such as power/current quality, capacitor voltage imbalance, and dc-voltage regulation of the 3L-NPC power converter in microgrids, achieving high power conversion efficiency. Confronting these issues, this dissertation proposes control schemes for the converter, featuring efficiency, reliability, stability, and robustness, which can meet the requirements and applications for high power penetration in microgrids. Outlook is lastly shared.

7.1 Conclusions

Ch.1 outlined the background of a 3L-NPC power converter in microgrids, identifying the research motivations and control objectives. Then, the state-of-the-art control techniques were discussed. Moreover, an overview of the thesis organization was summarized.

Ch. 2 introduced the working principle, mathematical modeling, and pulse width modulation techniques of the 3L-NPC converter. Based on the above, Ch. 3 proposed a new disturbance super-twisting observer-based quasi-integral sliding mode control (DSTO-QISM) scheme for the 3L-NPC power converter in microgrids to address power ripple, voltage swings, and weak anti-disturbance, raising power/voltage regulation performance and anti-disturbance capability. A DSTO was used to offset multi-disturbances caused by parameter mismatches, unknown loads, current path changes, switch mode noise, and self-compensating power/voltage tracking biases in QISM. To further mitigate power/voltage steady-state error and boost system robustness, a new quasi-integral sliding mode surface was built, inherently improving power/voltage tracking performance. Rigorous stability analysis of the discrete-time dynamics of the corresponding power/voltage system provided a theoretical basis for the 3L-NPC power converter system. The experimental tests showed that the proposed control outperforms the discrete-time extended state observer-based quasi-integral sliding mode control (DESO-QISM), discrete-time super-twisting observer-based quasi-sliding mode control (DSTO-QSM), and discrete-time proportional-

integral (DPI) control under different scenarios.

To customize the designs and optimizations for each control structure outlined in Ch. 3, Ch. 4 was proposed. Additionally, Ch. 4 improved the capacitor voltage balancing issues. In detail, Ch. 4 proposed an effective robust adaptive sliding mode control scheme, including three composite controllers, was developed to improve the control performance of the 3L-NPC power converter in microgrids against multi-disturbances in Ch. 4. The adaptive observer-based proportional (AO-P) controller was applied to regulate the dc-link voltage in the presence of the unknown dc-link equivalent loads. The disturbance observer-based integral Sliding mode control (DO-ISMC) was implemented to track the power references in the face of lumped disturbances caused by system parameter mismatches, unmodeled dynamics, measurement noise, and bias caused by current path changes and switch mode noise. The sliding mode observer-based proportional (SMO-P) was designed to balance (*two*) capacitors, avoiding ac-voltage low-frequency oscillation. Experimental results confirmed that the proposed control scheme outperformed the super-twisting observer-based integral sliding mode control (STO-ISMC), super-twisting algorithm (STA), and proportional-integral (PI) control schemes in terms of dynamic response, THDs, and anti-multi disturbances.

On the other hand, designing a suitable high-quality current control scheme featuring superior dynamics, strong robustness, and simple structure is a promising solution to guarantee the more efficient operation of the converter. Ch. 5 proposed the robust high-quality current control (STA-based extended sliding mode disturbance observer (ESMDO)) scheme to address the above problems. Specifically, the STA coupled with the Luenberger observer (LO) current controller was adopted to deal with the sinusoidal current tracking with inductance/grid frequency deviations and sinusoidal disturbance, ensuring the current tracking performance. Additionally, the proportional (P) controller coupled with the ESMDO was proposed to regulate voltage, achieving fast and accurate tracking voltage. From all experimental results, the proposed control scheme attained exceptional voltage regulation and current tracking performances, such as more stable output voltage, less current tracking error, and lower THD than the proportional resonance (PR)/PI control method under various parameter deviations/disturbances.

However, the control schemes above for the converter are model-based, and specific challenges are present when confronted with inaccurate models, critical loads, and multiple disturbances. Therefore, Ch. 6 presented a data-driven model-free adaptive control (MFAC) for the 3L-NPC power converter system in microgrid systems effectively improving voltage and current quality without any system parameters while guaranteeing the system stability, robustness, and flexibility. An MFAC was employed to achieve accurate voltage and current tracking through the design of adaptive laws based on Lyapunov theory, which ensured self-adaptability in the face of unmodeled dynamics, model variations, parameter mismatches, and disturbances. To facilitate access to adaptive law gains, MFAC used a particle swarm optimization (PSO) algorithm to automatically determine gains, reducing manual effort. Finally, experimental results confirmed that the proposal outperformed the artificial neural networks-based PI (ANN-PI) and super-twisting algorithm (STA) control schemes regarding dynamic response, THDs, and robustness.

7.2 Outlooks

Considering the high penetration rate of power converters in microgrids, a scalable and flexible control scheme is urgently needed.

Distributed control strategy: As microgrid systems become more complex, the limitations of centralized control for multiple power converters become apparent. A distributed control strategy facilitates decentralization by allowing local information exchange, thereby improving the flexibility, and scalability of the converter system.

Intelligent perception performance: Integrating artificial intelligence technologies, such as neural networks, expert systems, fuzzy logic, and machine learning, into converter systems control for microgrid applications can enhance system adaptability and decision-making efficiency, facilitating autonomous reactive power compensation, voltage and frequency regulation, and promoting black start recovery.

In summary, developing distributed control strategies, and intelligent perception and control techniques are crucial for the future of microgrid converter control. These advancements will enhance the reliability, efficiency, and scalability of microgrid systems, making them more capable of satisfying the demands of modern power networks.

APPENDIX A

List of Publications

A.1 Books

- Y. Yin, **L. Liu**, Z. Hu, H. Lin, L. Wu, “Robust Control Strategies for Power Electronics in Smart Grid Applications”, *Springer Cham*, 2024, ISBN: 978-3-031-53188-0.

A.2 Journal Papers

- **L. Liu**, Z. Zhang, Y. Zhao, G. Chen, H. Xie, Y. Yin, S. Vazquez and R. Kennel, “Data-driven Model-Free Adaptive Control for Power Converter under Multi-scenarios in Microgrids”, **IEEE Transactions on Industrial Electronics**, *Early Access*, Jan. 2025. DOI: 10.1109/TIE.2024.3525142.
- **L. Liu**, Y. Yin, Z. Zhang, H. Chen, H. Xie, Y. Liu, Y. Zhao and R. Kennel, “Multidisturbances Compensation for Three-Level NPC Converters in Microgrids: A Robust Adaptive Sliding Mode Control Approach”, **IEEE Transactions on Industrial Informatics**, vol.20, no.6, pp.8917-8931, Jun. 2024. DOI: 10.1109/TII.2024.3370237.
- **L. Liu**, Z. Zhang, Y. Yin, S. Vazquez, Y. Zhao and R. Kennel, “An Efficient Robust Power–Voltage Control for Three-Level NPC Converters in Microgrids”, *IEEE Transactions on Industrial Informatics*, vol.20, no.4, pp.5849-5863, Apr. 2024. DOI: 10.1109/TII.2023.3331548.
- **L. Liu**, Z. Zhang, Y. Yin, Y. Li, H. Xie, M. Zhang, Y. Zhao and R. Kennel, “A Robust High-Quality Current Control With Fast Convergence for Three-Level NPC Converters in Microenergy Systems”, **IEEE Transactions on Industrial Informatics**, vol.19, no.11, pp.10716-10726, Nov. 2023. DOI: 10.1109/TII.2023.3241594.

- Y. Yin, **L. Liu**^{*}, S. Vazquez, R. Xu, Z. Dong, J. Liu, J. Leon, L. Wu and L. G. Franquelo, “Disturbance and Uncertainty Attenuation for Speed Regulation of PMSM Servo System Using Adaptive Optimal Control Strategy”, **IEEE Transactions on Transportation Electrification**, vol.9, no.2, pp.3410-3420, Jun. 2023. DOI: 10.1109/TTE.2022.3227070.
- H. Xie, **L. Liu**, Y. He, F. Xiang, J. Rodriguez and R. Kennel, “Gradient Descent-Based Objective Function Reformulation for Finite Control Set Model Predictive Current Control with Extended Horizon”, **IEEE Transactions on Industrial Electronics**, vol.69, no.9, pp.8667-8678, Sep. 2022. DOI: 10.1109/TIE.2021.3116567.

A.3 Conference Papers

- **L. Liu**, H. Xie, D. Kong, Y. Yin, Y. Zhao, Z. Zhang and R. Kennel, “A Disturbance Observer-Based Integral Sliding Mode Control for Two-Level Power Converter”, *2021 IEEE International Conference on Predictive Control of Electrical Drives and Power Electronics (PRECEDE)*, 2021. DOI: 10.1109/PRECEDE51386.2021.9680925.
- H. Xie, F. Wang, Y. He, **L. Liu**, C. Wu, R. Qi, G. Luo and R. Kennel, “Cascaded Predictive Current Control with a Piecewise Approach for Induction Machine Drives”, *2021 IEEE International Conference on Predictive Control of Electrical Drives and Power Electronics (PRECEDE)*, 2021. DOI: 10.1109/PRECEDE51386.2021.9680982.

APPENDIX B

Nomenclature

B.1 Abbreviarions

3L-NPC	Three-level power converters
DPC	Direct power control
DCC	Direct current control
PID	Proportional-integral-derivative
P	Proportional
PI	Proportional-integral
PD	Proportional-derivative
DPI	Discrete-time proportional-integral
PR	Proportional resonance
SMC	Sliding mode control
HOSMC	High-order sliding mode control
SOSMC	Second-order sliding mode control
ISMC	Integral Sliding mode control
CSMC	Continuous-time SMC
DSMC	Discrete-time SMC
QSMC	Quasi-sliding mode control
QISMC	Quasi-integral sliding mode control
ASMC	Adaptive sliding mode control
STA	Super-twisting Algorithm
AC	Adaptive control
MRAC	Model reference adaptive control
STC	Super-twisting control

MBAC	Model-based adaptive control
MFAC	Model-free adaptive control
MFPCC	Model-free predictive current control
MFC	Model-free control
FFC	Feedforward control
FBC	Feed-back control
ANN	Artificial neural network
SO	State observer
DO	Disturbance observer
LO	Luenberger observer
ROO	Reduced-order observer
SMO	Sliding mode observer
AO	Adaptive observer
STO	Super-twisting observer
DSTO	Disturbance super-twisting observer
PO	Perturbation observer
UIOs	Unknown input observers
I-EID based-estimator	Improved equivalent input disturbance based-estimator
ESMDO	Extended sliding mode disturbance observer
ESO	Extended state observer
AO-P	Adaptive observer-based proportional
DO-ISMC	Disturbance observer-based integral Sliding mode control
SMO-P	Sliding mode observer-based proportional
STO-ISMC	Super-twisting observer-based integral sliding mode control
ESMDO-P	Extended sliding mode disturbance observer-based proportional
LO-STA	Luenberger observer-based super-twisting algorithm
PSO	Particle swarm optimization
ULM	Ultra-local model
PSO-MFAC	Particle swarm optimization-based model-free adaptive control
ANN-PI	Artificial neural network-based proportional-integral
DSTO-QISM	Discrete-time super-twisting observer-based quasi-integral sliding mode control
DSTO-QSM	Discrete-time super-twisting observer-based quasi-sliding mode control
DESO-QISM	Discrete-time extended state observer-based quasi-integral sliding mode control
DESO-QSM	Discrete-time extended state observer-based quasi-sliding mode control
THD	Total Harmonic Distortion

APPENDIX C

Testbench

This work verifies the effectiveness of the proposed control scheme on the lab-constructed 3L-NPC power converter prototype, as shown in Figure C.1. It is essential to consider the required infrastructure for implementing the proposed research. Here are the key points for the infrastructure needed to implement the proposed method.

C.1 Flow Diagram of the Experiment Infrastructure

The core components of the experimental infrastructure for implementing the suggested method include the Chroma 61830 Grid Simulator, $\pm 50\text{A}$ current sensors, $\pm 800\text{V}$ voltage sensors, imperix passive filters box, 3L-NPC power converter PEN8018, B-Box RCP^{3.0} control platform, and RT BOX Hardware-in-the-Loop platform, as shown in Figure C.1. These devices and platforms offer a wide range of features and technical specifications, making them ideal for conducting power electronics experiments and validating the proposed control schemes. The following will provide a detailed description.

C.2 Grid Simulator (Chroma 61830)

The Chroma 61830 offers testing solutions required when connecting to the grid. It features full four-quadrant operation, energy regeneration, and voltage waveform editing capabilities (e.g., simulating voltage dips and voltage harmonic distortion) to meet the regulations above and testing standards. Most importantly, the Chroma series provides an energy-efficient solution as the energy generated during testing can be fed back into the grid using this instrument rather than dissipating as heat. In addition to distributed energy source testing applications, the Chroma 61830 regenerative grid simulator can also be used for testing other green energy-related products, such as vehicle-to-grid (V2G) applications and energy storage systems (ESS).

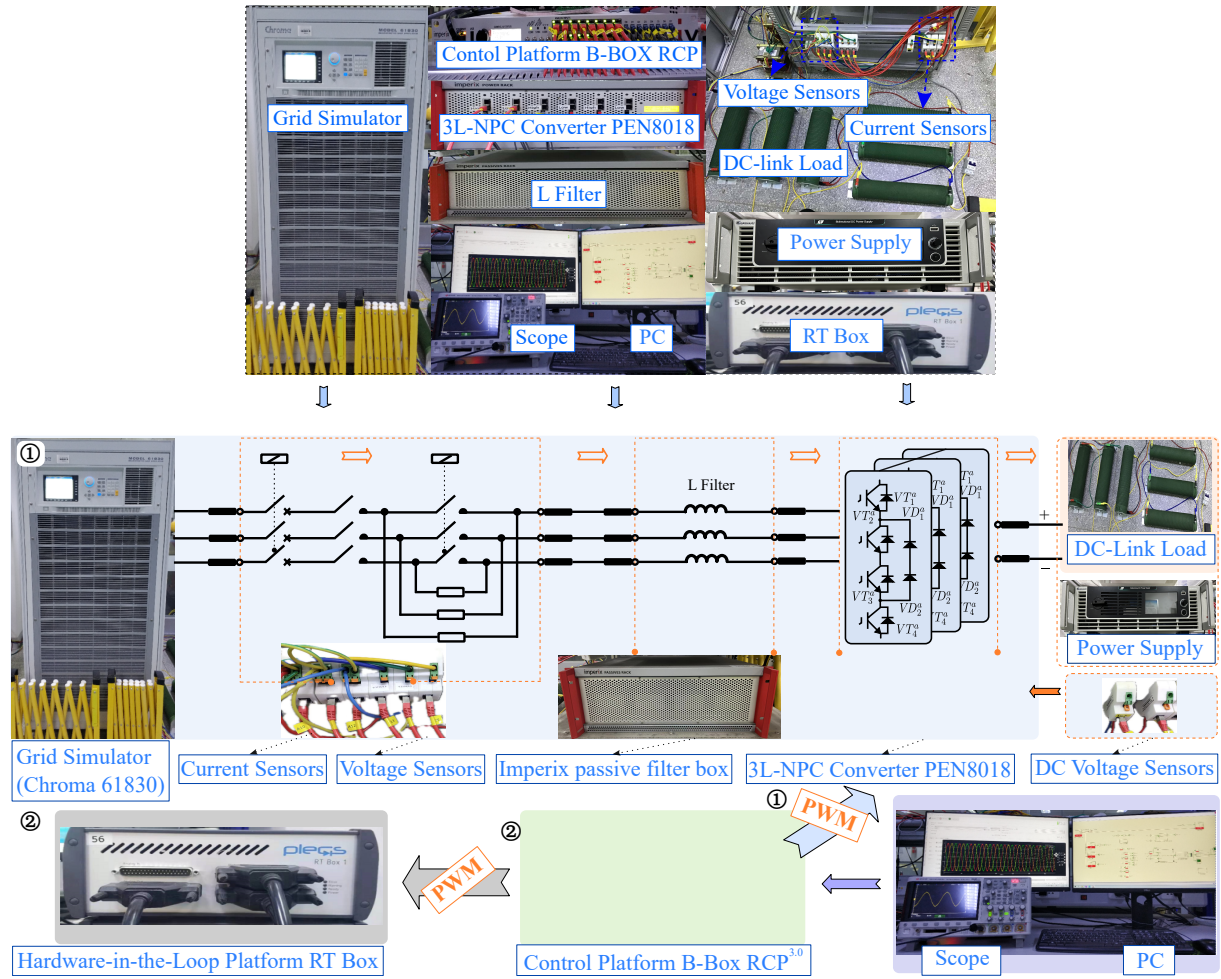


Figure C.1: The constructed 3L-NPC power converter prototype.

C.2.1 Feature

- (1) Voltage: $0 \sim 300$ V;
- (2) Frequency: DC, 30 Hz 100 Hz;
- (3) Energy regenerative function: 100% rated current recycling capability;
- (4) Conform with the test applications of PV inverter, Smart Grid, and EV associate products;
- (5) Selectable 1-phase/3-phase AC output;
- (6) Controllable voltage and frequency for change rate;
- (7) Output limit setting for voltage and current;
- (8) Voltage waveform setting for $0 \sim 360$ switching degree;

C.2.2 Specification

Table ?? lists the specifications of chroma 61830 grid simulator.

Table C.1: Key Features and Specifications of Current Sensors

No.	Key features and specifications
1	$\pm 50\text{A}$ measuring range
2	$0.2\text{-}6\text{mm}^2$ conductor cross-section (AWG 10-24)
3	Minimum 200kHz measurement bandwidth
4	Balanced full-differential signal output
5	Typical sensitivity of 99.0 mV/A
6	$1.0\text{ kV}_{\text{PK}}$ galvanic isolation (permanent)
7	$\pm 0.45\%$ typical precision
8	Compatible with standard $\pm 15\text{V}$ power supplies
9	Mountable on 35mm DIN rails

C.3 Current Sensors ($\pm 50\text{A}$ Din Rail-mountable Current Sensors)

The ModuLink $\pm 50\text{A}$ isolated current sensors are easy-to-use devices, tailored for power electronics applications. When employed with Imperix's BoomBox control platform, the sensors provide plug&play connectivity and can be directly powered by the control platform. The devices produce a balanced full-differential output signal, proportional to the current voltage. For best EMC performance, the output signal is typically carried by a shielded twisted pair embedded in RJ45-type cables. Modulink sensors can be easily clipped on 35mm DIN rails and are compatible with up to 6mm^2 wires. They guarantee the galvanic isolation of the sensing circuit up to 1 kV_{RMS} .

C.3.1 Key Features and Specifications of Current Sensors

Table. C.1 lists the key features and specifications of current sensors.

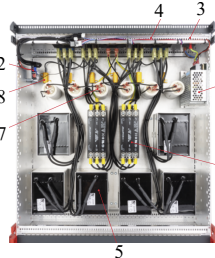
C.4 Voltage Sensors ($\pm 800\text{V}$ Din Rail-mountable Voltage Sensors)

The ModuLink $\pm 800\text{V}$ isolated voltage sensors are easy-to-use devices, tailored for power electronics applications. When employed with Imperix's BoomBox control platform, the sensors provide plug&play connectivity and can be directly powered by the control platform. The devices produce a balanced full-differential output signal proportional to the sensed voltage. For best EMC performance, the output signal is typically carried by a shielded twisted pair embedded in RJ45-type cables. Modulink sensors can be easily clipped on 35mm DIN rails and are compatible with up to 6mm^2 wires. They guarantee the galvanic isolation of the sensing circuit up to 1 kV_{RMS} . The voltage sensors are intended to enhance accuracy, with a minimum requirement for offset compensation.

Table C.2: Key Features and Specifications of Voltage Sensors

No.	Key features and specifications
1	$\pm 800\text{V}$ measuring range
2	Minimum 60 kHz measurement bandwidth
3	Balanced full-differential signal output
4	Typical sensitivity of 2.46 mV/V
5	1.2 kV _{PK} galvanic isolation (permanent)
6	$\pm 0.15\%$ typical precision
7	Compatible with standard $\pm 15\text{V}$ power supplies
9	Mountable on 35mm DIN rails

Table C.3: Components of Filters Box

No.	Designation	Manufacturer	Picture
1	Auxiliary power supply	Delta electronics	
2	Fuse	Bel fuse inc.	
3	Fan speed controller	CZB	
4	Cooling fan (3x)	Orion Fans	
5	Inductor (6x, 2.2mH, 32A)	Hahn	
6	EMC filter (2x)	Schaffner EMC inc.	
7	Film capacitor (3x, 10uF)	Arcotronics/KEMET	
8	Damping resistor (3x, 1 Ω)	Ohmite	

C.4.1 Key Features and Specifications of Voltage Sensors

The following Table C.2 lists the key features and specifications of voltage sensors.

C.5 Imperix Passive Filters Box

The passive filter box contains essential components for interfacing power converters with different loads. Individual sockets for 4mm laboratory plugs enable simple connection/disconnection from/to external components. For instance, Imperix type C enclosures offer similar connectivity for all types of power modules.

C.5.1 Internal Components

The system can be used for up to two inverters (three phase), or within back-to-back configurations. The passive filters box essentially implements:

- (1) 6x power inductors rated 2.2mH/32A.
- (2) 2x EMC filters with star-connected capacitors. The filters box embeds the components in Table C.3.

C.5.2 Main Electrical Schematic

The main power circuits are flexible and can be used in reconfigurable power converter systems. They consist of independent elements, including inductors, EMC filters, and associated capacitors and resistors.

C.5.3 Main Electrical Specifications

These specifications include operating AC voltage, current, inductance, output filter capacitance, and resistance.

C.6 3L-NPC Power Converter PEN8018

The 3L-NPC Power converter PEN8018 modules are designed for low-voltage power converters. The mechanical design is tailored for 19" rack integration with simple interconnections and direct connection to a BoomBox or any other control platform. Each module contains an NPC cell, corresponding to four power switches and two diodes. Direct access to the gating signals is offered using optical fiber inputs, while embedded measurement circuits provide direct analog outputs related to one of the half DC link voltages and the AC output current using galvanically-isolated sensors. Over-voltage, over-current and over-temperature protections are also integrated on the board for safer use in R&D applications. Besides, these protections are user programmable through a simple onboard micro-controller and a CPLD. Finally, the possibility to use alternative or customized mezzanine boards enables the compatibility with future developments and onboard data processing.

C.6.1 Typical Applications

The modules are ideally suited to build ambitious prototypes of low-voltage NPC power converters, ranging from conventional three-phase three-level inverters to more complex multilevel topologies. Typical power ratings are around 10 kW, depending on the nominal DC link voltage and the switching frequency.

C.6.2 Key Features and Specifications & Main Components

Table C.5 lists the key features, specifications, and main components of 3L-NPC Power Converter PEN8018.

C.7 Control Platform B-Box RCP^{3.0}

: The B-Box RCP (DSP+FPGA controller for rapid control prototyping applications) is a highly flexible power converter control system, tailored for rapid control prototyping applications in power electronics. Thanks to its high performance and flexibility, it facilitates the experimental validation of power converters control techniques in laboratory environments.

Table C.4: Key Features and Specification 3L-NPC Power Converter PEN8018

No.	Key features and specifications	Component	Devices
1	18A/800V maximum ratings, limited by losses	Power switches	1x Vincotech P924F33 module
2	600V/30A IGBTs	Capacitors	2x 517 uF
3	90A max pulsed current (limited by)		Panasonic EEU-EE2W470S
4	120W TDP envelope		(2 banks of 11x47uF each)
5	2 517uF/400V half DC buses	Drivers	4x Avago ACPL-P345
6	Up to 50 kHz switching frequency	Isolated DC/DC converters	4x Recom RK-0515S
7	4 optical inputs/1 optical output		
8	2 analog outputs	Current sensor	1x LEM HLSR 20-P/SP33
9	+5V and +12V power supplies	Voltage sensor	1x Resistive divider + Avago ACPL-C87B
10	Embedded voltage and current measurement (Upper or lower half-bus voltage measurement)		
11	Over voltage/current/temperature protection	Heatsinks	1x Dynatron G199
12	User-configurable CPLD	CPLD	1x Xilinx XC9536XL-10VQG44C
13	100 330mm Eurocard form factor		
14	Forward compatible with imperix RealSync technology and the related mezzanine board	Microcontrollers	2x Microchip PIC24F04KA101
			2x Microchip PIC24F04KA101

C.7.1 Hardware Components

- (1) Embedded Computing Unit: B-Box RCP^{3.0} is equipped with an embedded computing unit, typically utilizing a high-performance multi-core processor for executing control algorithms, data processing, and real-time operations.
- (2) Communication Interfaces: B-Box RCP^{3.0} supports a variety of communication interfaces, including Ethernet, CAN bus, UART, USB, and others, for connecting to external devices, sensors, and actuators to facilitate data transmission and control.
- (3) Expansion Slots: The platform typically includes expansion slots, allowing users to add additional hardware modules as needed, such as extra input/output modules or specific interface modules.

C.7.2 Software Platform

- (1) Embedded Operating System: B-Box RCP^{3.0} runs on an embedded operating system to ensure real-time operation and stability for the execution of control algorithms and applications.
- (2) Control Algorithm Development Tools: Developers typically use specialized software tools such as PLECS/MATLAB to design, test, and deploy control algorithms.
- (3) Communication Protocols: It supports various communication protocols to exchange data with other devices and systems, such as TCP/IP, Modbus, and others.

C.7.3 Data Storage and Management

- (1) Data Storage Device: B-Box RCP^{3.0} includes built-in storage or supports external storage media for storing data, logs, and configuration files.
- (2) Data Management Software: It often includes data management software for recording, retrieving, and analyzing the system's operational data, supporting remote monitoring and fault diagnosis.

C.7.4 Power Supply

Power Module: B-Box RCP^{3.0} is equipped with a power module to provide a stable power supply, ensuring the system's proper functioning.

C.7.5 Mechanical Structure and Casing

- (1) Mechanical Structure: B-Box RCP^{3.0} has specific physical dimensions and mechanical structure to accommodate various applications and installation requirements.
- (2) Casing: The platform typically includes a casing to protect internal components from environmental conditions.

C.7.6 Specification

Table C.5 lists the specification of B-Box RCP^{3.0}.

Table C.5: Key Specifications of B-Box RCP^{3.0}

Indexes		Channels
System on chip	Zynq XC7Z030-3	1
Processor	ARM cortex A9 1GHz	2
	1GB DDR3	
FPGAs	Kintex 7 125K (main)	1
	Artix 7 35T (auxiliary)	1
Anglog inputs	16bits @ 500ksps	16
Incremental decoder input	3-pins (A,B,Z)	4
	(shared with GPI inputs)	
PWM outputs	Optical	16
	Electrical (3.3V)	32
User I/Os	Electrical (3.3V)	36
high-speed	(bi-directional)	
General-purpose I/Os	IN, electrical (3.3V/5.0V)	16
	OUT, electrical (3.3V/5.0V)	16
Fault inputs	Electrical (3.3V)	16
	Electrical interlock	1
	Optical interlock	1
Communication	Optical interlock	1
	Ethernet 1 Gbps	1
	SFP+5Gbps (Real Sync)	3

C.8 Hardware-in-the-Loop Platform-RT BOX

The RT Box is a hardware-in-the-loop (HIL) simulator designed for engineers. It is most suitable for simulating power electronic circuits of moderate complexity and single-tasking control prototyping. Here are some key features and technical specifications.

C.8.1 Processor

The RT Box is equipped with a Xilinx Zynq Z-7030 system-on-chip that contains two CPU cores on an FPGA. One core is dedicated to real-time simulation, while the other runs an embedded Linux for communication and ancillary services. The FPGA fabric is used to control analog-to-digital converters (ADCs) and digital-to-analog converters (DACs) for analog channels, as well as for pulse-width modulation (PWM) generation and capture on digital channels.

Table C.6: Technical Specifications of RT-BOX

Indexes	Specifications		Indexes	Specifications	
Processor	Xilinx Zynq	Z-7030	Digital inputs	Channels	32
Analog input	CPU cores	2 x ARM Cortex-A9, 1 GHz		Logic levels	3.3 V (5 V tolerant)
	Channels	16		Connector	D-SUB 37 pin male
	Resolution	16 bit, simultaneous sampling		Channels	32
	Voltage ranges	-10...10V, -5...5V		Logic levels	3.3 V, 5 V
	Input type	Differential	Digital outputs	Connector	D-SUB 37 pin female
	Max. sample rate	2 Msps		Short-circuit	Permanent
Analog outputs	Input impedance	1 M Ω , 24pF		Overvoltage	-24 ... 24 V
	Connector	D-SUB 37 pin male	I/O protection	Gigabit Ethernet	1
	Channels	16		SFP+ interconnects	4
	Resolution	16 bit. simultaneous update		6.25 Gbps per lane	
	Voltage ranges	-10...10V, 0...10V -5...5V, 0...5V		CAN bus	2
	Max. update rate	2 Msps	Connectivity	USB A 2.0	1
	Output impedance	0 Ω		Firmware	SD card
	Max. output current	10 mA		Power supply (internal)	100 ... 240 Vac 50 ... 60 Hz
	Connector	D-SUB 37 pin female	Power supply (internal)		50 VA

C.8.2 Simulation Time Steps

The simulator can achieve simulation time steps as low as 1 us for small power electronic circuits like DC-DC converters.

C.8.3 Scalability

For larger systems such as multi-level inverters, multiple units of RT Box can be interconnected using high-speed serial communication links. The RT Box, in particular, has four high-speed bi-directional links, each supporting a data rate of 6.25 Gbit/s. Connected RT Boxes can operate synchronously or asynchronously.

C.8.4 Technical Specifications

The key technical specifications of the RT Box are listed as shown in Table C.6.

C.9 In summary

These devices and platforms together form a comprehensive set of experimental infrastructure, providing powerful tools for research and development in the field of power electronics. Researchers can choose the appropriate equipment based on their specific experimental requirements, ensuring high-precision data and control for their experiments and projects. The features and specifications of these devices will contribute to advancing the technology and applications of power electronics.

APPENDIX D

Engineering Guidelines

D.1 Engineering Issues in 3L-NPC Converter Systems

The 3L-NPC power converter system presents a highly complex and nonlinear dynamic behavior. It is subject to multiple challenges, including multi-source disturbances, weak grid conditions, harmonic distortions, voltage imbalances, and oscillatory transients as shown in Figure.D.1.

D.2 Multi-loop Control Architecture

The 3L-NPC converter system implements a multi-loop control architecture:

DC-voltage regulation loop : stabilizes the dc-link voltage.

Power/Current tracking loop : ensures accurate tracking of power commands or current references.

Capacitor voltage balancing loop : maintains the voltage balance between the DC-link capacitors, critical for stable 3L-NPC operation.

Each control loop addresses a specific dynamic aspect of the converter, contributing to comprehensive system regulation.

D.3 Control Method Classification

The 3L-NPC converter control methodologies are generally categorized as:

Direct power control (DPC) : controls instantaneous active (p) and reactive (q) power. It requires equilibrium points for stable operation and prioritizes fast dynamic response. DPC is well-suited for grid-connected systems and microgrids.

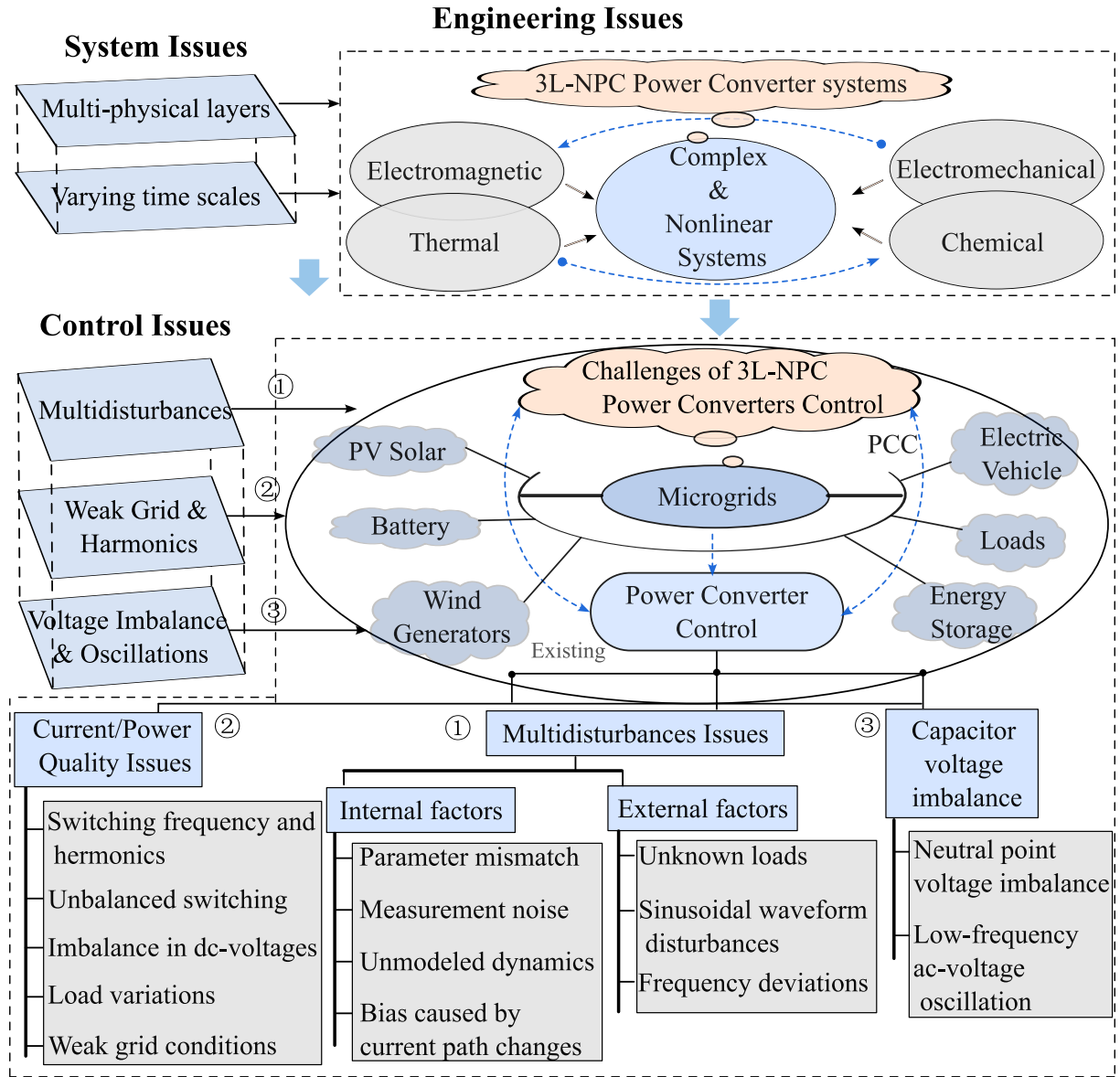


Figure D.1: Engineering issues in 3L-NPC converter systems

Direct current control (DCC) : regulates converter-side current components ($i_{\alpha\beta}$ or i_{dq}). DCC emphasizes steady-state tracking accuracy and does not rely on equilibrium points. It is commonly applied in power converters and motor drive systems.

The dissertation investigates both control approaches to comprehensively address the control needs of the 3L-NPC converter under various operational scenarios as shown in Table. D.1.

D.4 Experimental Validation Objectives

The proposed control strategies are validated on a laboratory-built 3L-NPC converter prototype. The experimental verification targets the following three primary objectives: dc-link voltage

Table D.1: Control Classification of Power Converter

Control Method	DPC	DCC
Control Variables	p, q	$i_{\alpha\beta}, i_{dq}$
Equilibrium Point Requirement	Yes (for power stability)	No (direct current control)
Performance Evaluation	Fast Dynamic response	Accurate Current tracking
Application Scenarios	Grid-connected, microgrids	Power converters, Motor drive

regulation; power or current tracking; capacitor voltage balancing.

These objectives are fundamental to ensuring the reliable and stable operation of 3L-NPC power converters in practical grid-connected applications.

D.5 Power-Voltage Control – Addressing Key Engineering Issues

Key challenges in power-voltage control include power ripple, voltage fluctuations, multi-disturbances, suboptimal dynamic response, harmonic injection, and capacitor voltage imbalance as shown in Figure.D.2.

D.5.1 Solutions

Quasi-Sliding Mode Control (QSMC) in Ch. 3: Effectively addresses power ripples, voltage swings, dynamic disturbances, and optimization issues. QSMC is suited for high-power converters operating in digital discrete-time environments.

Adaptive Sliding Mode Control (ASMC) in Ch. 4: Extends the capabilities of QSMC to also mitigate harmonic injection and capacitor imbalance, making it applicable for continuous-time systems in renewable energy storage.

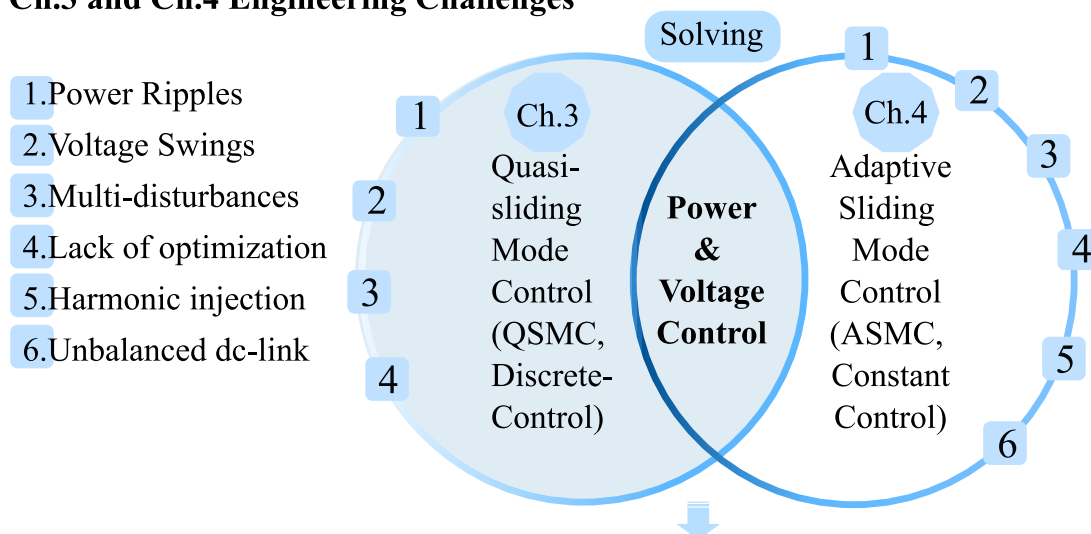
D.5.2 Core Solution Idea of Quasi-sliding Mode Control

The key to solving the core problem of Quasi-sliding mode control (QSMC) in Ch. 3 is illustrated in the Figure.D.3.

Discrete-time super-twisting observer (DSTO) : enables power/voltage bias compensation, chattering suppression, and multi-disturbance rejection (handling parameter mismatches, load uncertainty, topology changes, and sensor noise).

Quasi-integral sliding mode (QISM) : reduces steady-state power/voltage tracking errors.

Ch.3 and Ch.4 Engineering Challenges



Two Controls: Applications & Designs

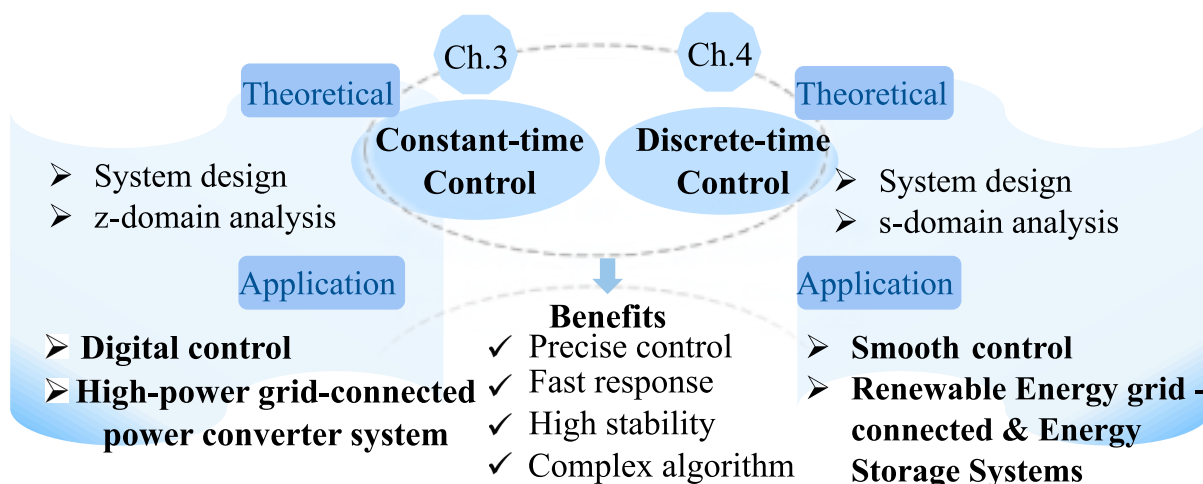


Figure D.2: Engineering issues of power and voltage control.

Quasi-integral sliding mode control (QISMIC) : ensures accurate power/voltage regulation, forming the theoretical foundation of the DSTO-QISMIC controller.

D.5.3 Core Solution Idea of Adaptive Sliding Mode Control (ASMC)

Despite effectiveness of QSMC, it cannot fully eliminate harmonic injection and capacitor voltage imbalance. To solve these:

Adaptive observer-based P (AO-P) control : estimates real-time disturbances (e.g., unknown loads, topology changes, sensor noise) to ensure fast voltage regulation.

Disturbance observer-based integral sliding mode control (DO-ISMC) : reconstructs multi-disturbances (e.g., inductance mismatch, current path changes, measurement noise),

Ch.3 Core solution Idea

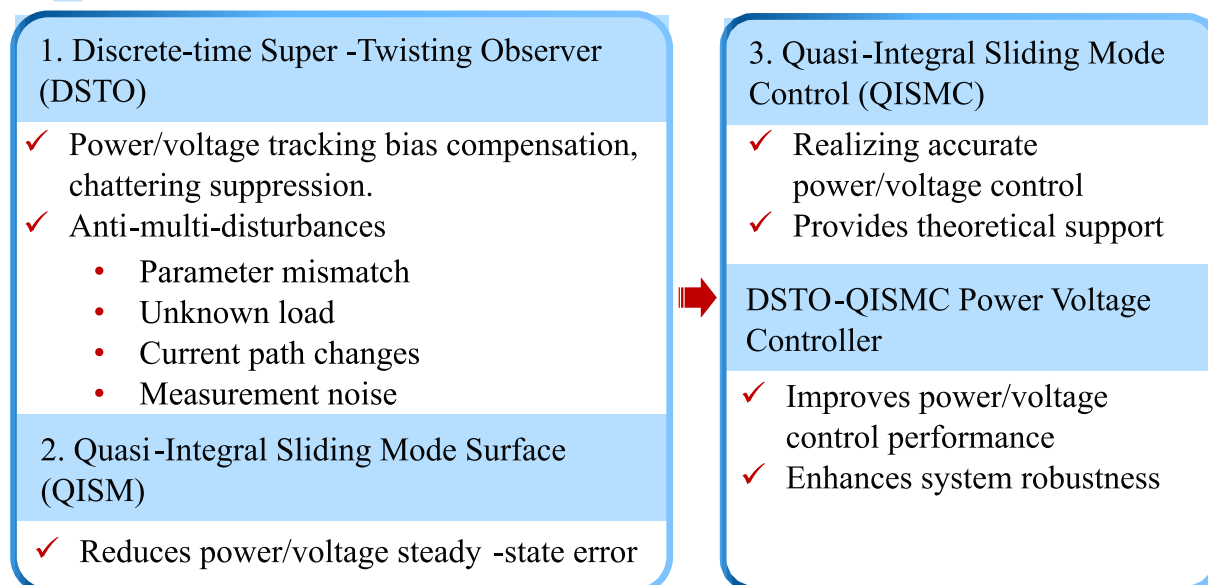


Figure D.3: Core solution idea of DSMC.

suppresses chattering, and improves power tracking.

Sliding mode observer-based P (SMO-P) control : focuses on suppressing AC voltage harmonics and balancing capacitor voltages.

Collectively, these methods constitute the ASMC Power/Voltage Controller, enhancing robustness and tracking accuracy under complex grid conditions. The Figure.D.4 shows the key to solving the core problem in Ch. 4.

D.6 Current and Voltage Control-Addressing Key Engineering Issues

Further challenges in grid-connected systems include redundant control layers, unstable grid frequencies, sinusoidal disturbances, multiple disturbances, model inaccuracies, and complex parameter tuning, as shown in Figure. D.5.

D.6.1 Core Solution Idea of Super-twisting Control

The key to solving the core problem in Ch. 5 is illustrated in Figure.D.6. Control Schemes:

Super-twisting control (STC) : solves dynamic instability, grid frequency deviation, and sinusoidal disturbance rejection without requiring a phase-locked loop (PLL), simplifying the decoupling process.

Ch.4 Core solution Idea

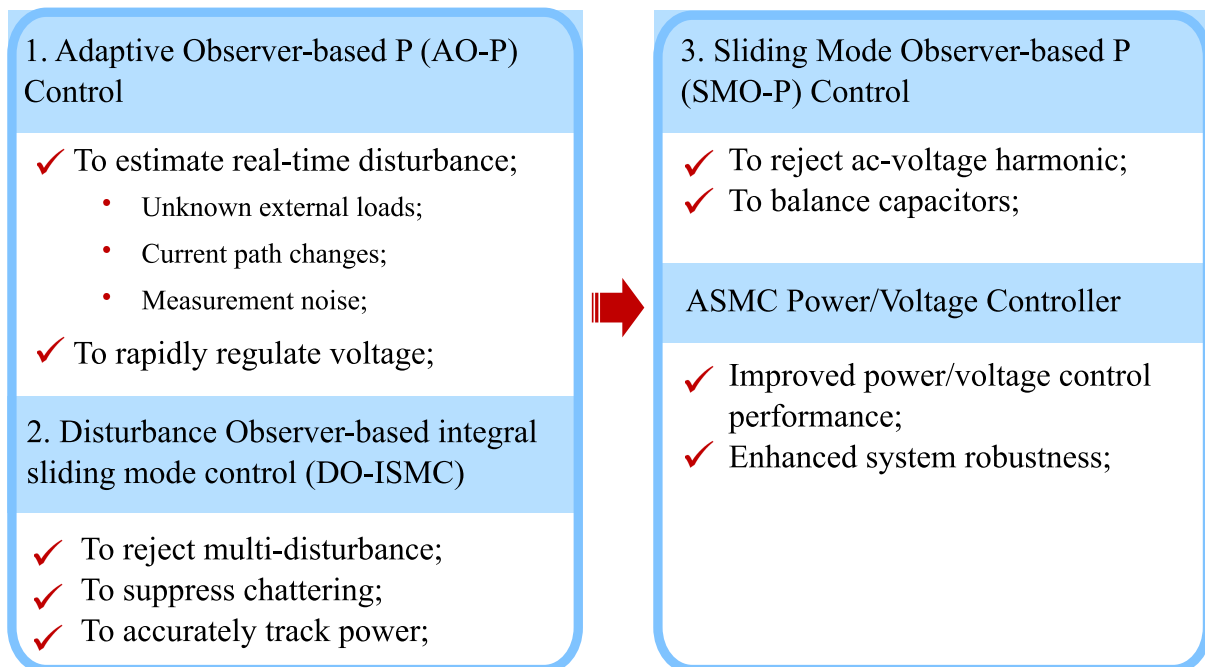


Figure D.4: Core solution idea of ASMC

Ch.5 and Ch.6 Engineering Challenges

- 1.Redundant control structures;
- 2.Unstable grid frequency;
- 3.Sinusoidal dynamic nature disturbance;
- 4.Multi-disturbances;
- 5.Inaccurate models;
- 6.Critical loads;
- 7.Complex coefficient tuning

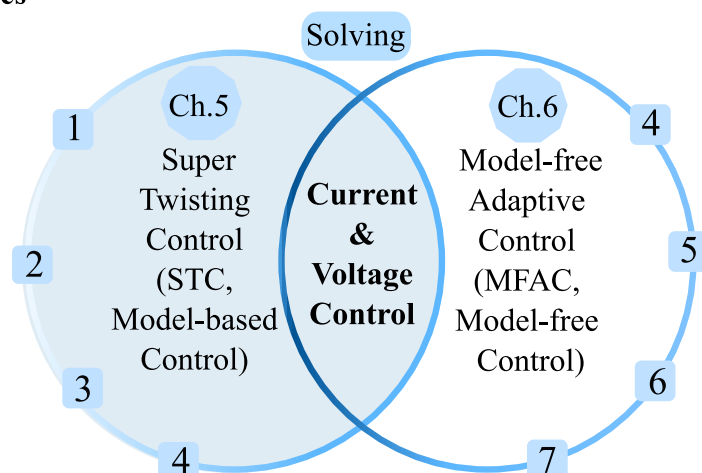


Figure D.5: Engineering issues of current and voltage control

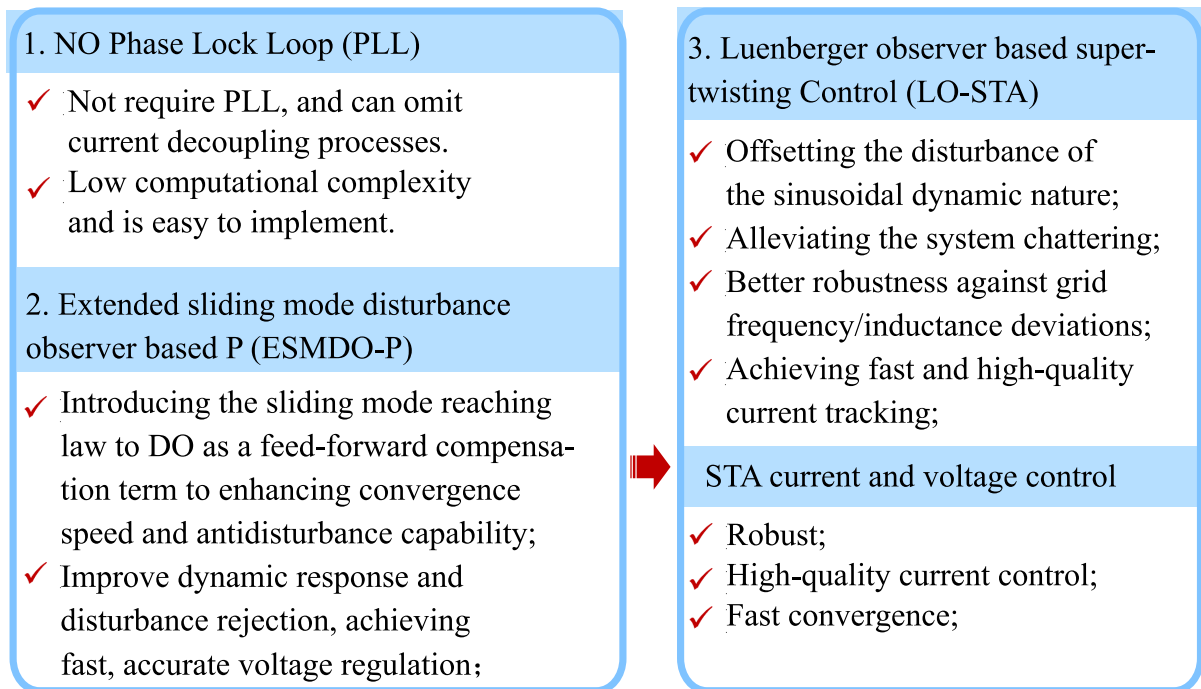
Ch.5 Core solution Idea

Figure D.6: Core solution idea of STC

Luenberger observer-based super-twisting algorithm (LO-STA) : improves current tracking under inductance variations and sinusoidal disturbances.

Extended sliding mode disturbance observer-based P (ESMDO-P) control : enhances dynamic response and disturbance rejection, achieving fast and accurate voltage regulation.

D.6.2 Core Solution Idea of Data-driven MFAC

While model-based control is popular due to its structured design, it struggles with system uncertainties and complex parameter tuning. To overcome this, a model-free adaptive control (MFAC) scheme is developed in Ch. 6. MFAC relies solely on input-output data, without requiring system parameters, ensuring robust and accurate current/voltage tracking under unknown system conditions.

To automate gain tuning, a particle swarm optimization (PSO) algorithm is integrated into the MFAC framework, enabling adaptive gain optimization and improving tracking efficiency and robustness. The core solution idea is shown in Figure. D.7.

Ch.6 Core solution Idea

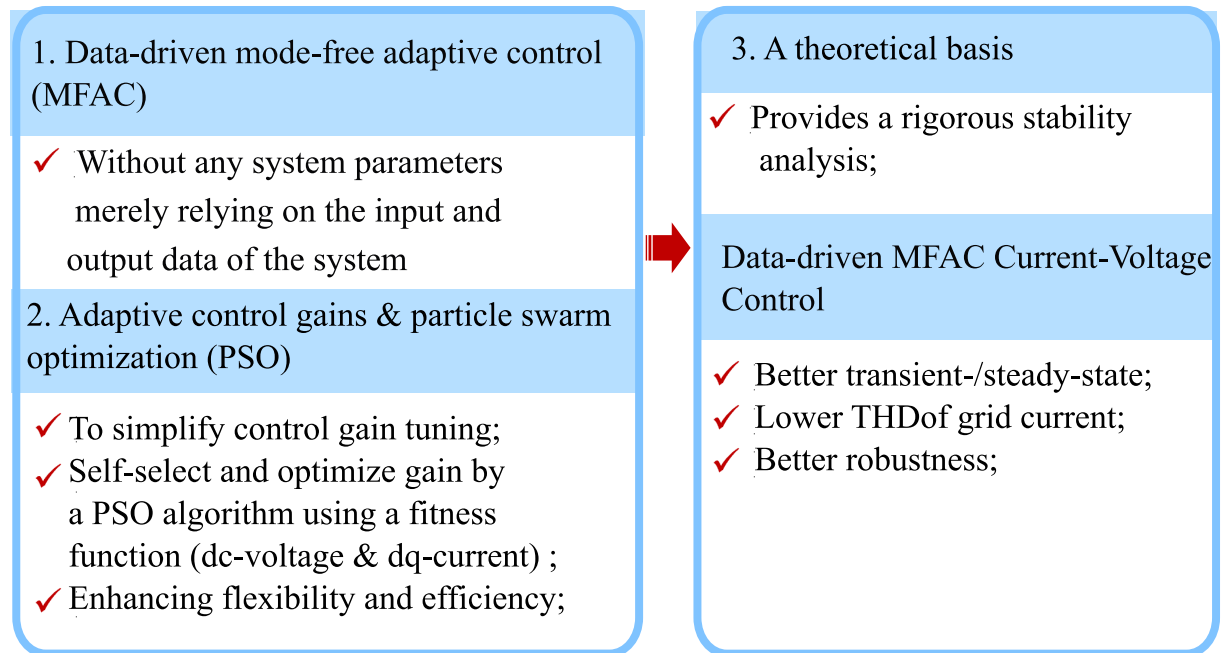


Figure D.7: Core solution idea of data-driven MFAC

D.7 Problems Addressed and Application Scenarios of Control Schemes

This dissertation aims to design robust and reliable control schemes for the 3L-NPC power converter systems in microgrids, as shown in Figure. D.8, covering: QSMC solves anti-disturbance (using discrete control for high-power grid integration). ASMC solves dc-capacitor imbalance (using constant control for renewable energy storage). STC realizes PLL-free and fast sinusoidal current control (for precise current control scenarios). Data-driven MFAC achieves without system parameters and self-selects/optimizes control gains (applicable in multi-scenario bidirectional power conversion).

D.8 Experimental Devices and Circuit Assembly

The experimental validation platform comprises a grid-connected 3L-NPC converter prototype. The platform includes:

- Grid simulator;
- Current and voltage sensors;
- Passive filter box;
- 3L-NPC power converter;

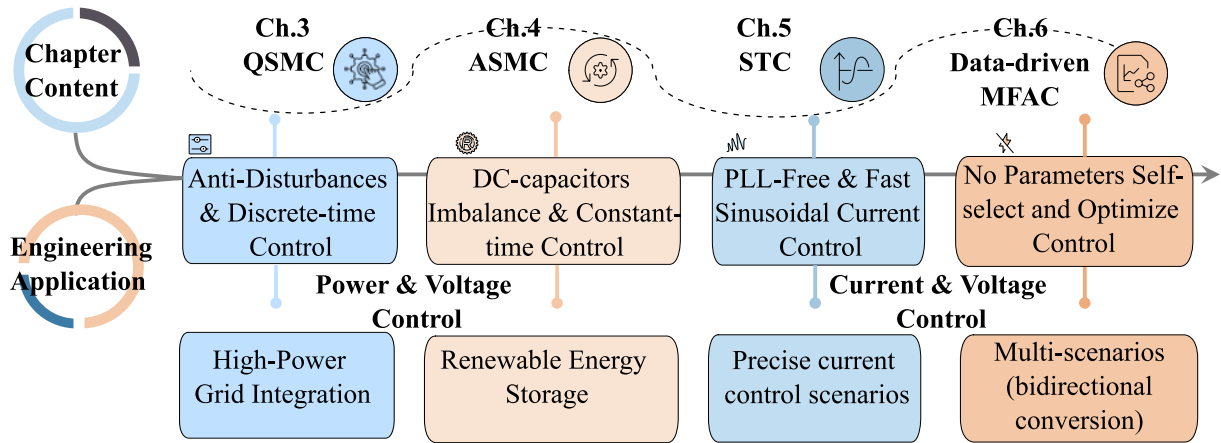


Figure D.8: Problems addressed and application scenarios of the control scheme

- DC voltage sensors and DC loads;
- Control hardware;
- Power supply;
- Hardware-in-the-Loop (HIL) platform oscilloscope;

Detailed device configurations and wiring schemes are elaborated in APPENDIX C.

D.9 Experimental Procedure

The experimental procedure of the 3L-NPC power converter involves the following steps:

- Control system design and implementation;
- Control parameter tuning and optimization;
- Execution under varied operating conditions;
- Comprehensive system performance evaluation;
- Power/current tracking performance evaluation;
- DC-link voltage regulation performance evaluation;
- Capacitor voltage balancing performance evaluation;
- Robustness assessment.

D.10 Real-Time Monitoring and Adjustment

Continuously observe system behavior during experiments.

- Monitor key performance indicators such as tracking error, voltage deviation, and oscillation.
- Conduct preliminary data analysis in real-time to detect trends and anomalies.
- Adjust parameters or experimental conditions as necessary based on observations.

D.11 Data Collection and Analysis

- Ensure rigorous data recording using digital databases for traceability. After experiments, perform detailed data analysis to assess control effectiveness under various operating scenarios.
- Evaluate the impact of tuned parameters on system performance.

D.12 Experiment Conclusion

- Terminate experiments according to planned schedules or once sufficient data is obtained.
- Shut down all equipment following safe power-down protocols.

D.13 Safety Measures

- Implement necessary safety protocols to protect personnel and equipment.
- Prepare contingency plans to immediately halt the experiment in case of unforeseen risks.

D.14 Summary

By rigorously documenting the experimental process, this work ensures repeatability, transparency, and a comprehensive understanding of the converter system's dynamic response under real-world disturbances. This section provides both theoretical insights and practical engineering guidance for deploying robust converter control systems in microgrids and other grid-interfaced applications.

APPENDIX E

Discrete-time Extended State Observer-based Quasi-sliding Mode Control Design

E.1 Discrete-time Extended State Observer-based Quasi-sliding Mode Control (DESO-QSMC) Design for Power Tracking

E.1.1 Discrete-Time Extended State Observer (DESO) Design

Based on (3.14), the following augmented state of power converter error system construct space model are obtained as,

$$\begin{cases} \xi_{j,k+1} = I_j \xi_{j,k} + J_j \mu_{j,k} + U_j \Delta g_{j,k} \\ y_{j,k} = L_j \xi_{j,k} \end{cases}, \quad (\text{E.1})$$

where $j = \{p, q\}$, $\xi_{j,k} = [e_{j,k}, g_{j,k}]^T$, $\Delta g_{j,k} = g_{j,k+1} - g_{j,k}$, $I_j = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}$, $J_j =$

$\begin{bmatrix} -O_{j,k} \\ 0 \end{bmatrix}$, $U_j = \begin{bmatrix} 0 \\ 1 \end{bmatrix}$, and $L_j = [1, 0]$, $y_{j,k}$ is the control input of system (E.1).

Design DESO for the extended system (E.1) as follows,

$$\begin{cases} \hat{\xi}_{j,k+1} = I_j \hat{\xi}_{j,k} + J_j \mu_{j,k} + \phi_j(y_{j,k} - \hat{y}_{j,k}) \\ \hat{y}_{j,k} = L_j \hat{\xi}_{j,k} \end{cases}, \quad (\text{E.2})$$

where $\hat{\xi}_{j,k} = [\hat{e}_{j,k}, \hat{g}_{j,k}]^T$, and $\phi_j \in \mathbb{R}^{2 \times 1}$ are observer gains to be designed. Define estimation error as:

$$\tilde{\xi}_{j,k} = \xi_{j,k} - \hat{\xi}_{j,k}. \quad (\text{E.3})$$

Subtracting (E.2) from (E.1) yields,

$$\tilde{\xi}_{j,k+1} = (I_j - \phi_j L_j) \tilde{\xi}_{j,k} + U_j \Delta g_{j,k}, \quad (\text{E.4})$$

if there exist the positive symmetric matrices $P_j \in \mathbb{R}^{2 \times 2}$, such that

$$(I_j - \phi_j L_j)^T P_j (I_j - \phi_j L_j) - P_j < 0, \quad (\text{E.5})$$

i.e., the eigenvalues matrix $(I_j - \phi_j L_j)$ are inside the unit circle of the z -plane. Since (E.5) is not the linear matrix, defining $\Upsilon_j = P_j \phi_j$, $\Upsilon_j \in \mathbb{R}^{2 \times 1}$ by Schur complement lemma, it can be obtained

$$\begin{pmatrix} -P_j & \Upsilon_j L_j \\ * & I_j^T P_j I_j - I_j^T \Upsilon_j I_j - L_j^T \Upsilon_j I_j - P_j \end{pmatrix} < 0, \quad (\text{E.6})$$

then the system $\tilde{\xi}_{k+1} = (I_j - \phi_j L_j) \tilde{\xi}_k$ is global stability. Suppose $\Delta g(k)$ is bounded, it can be concluded that estimate error $\tilde{\xi}_k$ is ultimately bounded, i.e., the DESO can estimate disturbance $g_{j,k}$ with ultimate boundness as $\hat{g}_{j,k}$.

E.1.2 Quasi-integral Sliding Mode Control (QSMC) Design

The quasi-sliding mode surfaces of power tracking is defined as,

$$S_{j,k} = e_{j,k}, \quad (\text{E.7})$$

Based on the aforementioned, a DESO-QSMC is designed as,

$$\mu(e_{j,k}) = \frac{1}{O_{j,k}} (\tau_j e_{j,k} + \varsigma_{jsw} \text{sign}(S_{j,k}) + \hat{g}_{j,k}), \quad (\text{E.8})$$

where $\tau_j > 0$, $\varsigma_{jsw} > 0$ are sliding mode control gains to be designed. The proof follows *Theorem 3.3.2*.

Remark E.1.1 *The tuning procedures of the DESO is shown as follows. According to (E.5), the stability of proposed DESO can be guaranteed by setting the eigenvalues (i.e., poles) of $(I_j - \phi_j L_j)$ inside the unit circle of the z -plane. It is worth mentioning that a tradeoff should also be made between the dynamic performance and robustness against the model uncertainties when selecting the pole locations in this work. E.g., the two poles of the DESO can be placed to obtain a several-times-faster dynamics compared with the open-loop system. As a result, since the two poles of $(I_j - \phi_j L_j)$ are inside the unit circle, the designed DESO is stable.*

Remark E.1.2 The parameter τ_j and ς_{jsw} of DESO-QSMC is shown as follows. 1) The parameter τ_j is often referred to as the reaching gain. It affects the convergence rate of the system to the sliding surface $S_{j,k} = 0$. A larger τ_j will make the system converge faster, but it can also introduce chattering if set too high. 2) The parameter ς_{jsw} is known as the boundary layer gain and is used to reduce chattering near the sliding surface. It should be set to a positive value, but it is often smaller than τ_j . Smaller values of ς_{jsw} reduce chattering but may slow down convergence. Thus, a trial-and-error process is still essential to attain better power tracking performance.

E.2 DESO-QSMC Design for Voltage Regulation

E.2.1 DESO Design

Augmented state space model of voltage regulation for the system (3.46) can be obtained as,

$$\begin{cases} \theta_{k+1} = X\theta_k + Yp_k^* + W\Delta d_k, \\ r_k = T\theta_k, \end{cases} \quad (\text{E.9})$$

where $\Delta d_k = d_{k+1} - d_k$, $\theta_k = [z_k, d_k]^T$, r_k is control input, $X = \begin{bmatrix} 1 & -E \\ 0 & 1 \end{bmatrix}$, $Y = \begin{bmatrix} 2E \\ 0 \end{bmatrix}$, $W = \begin{bmatrix} 0 \\ 1 \end{bmatrix}$, and $T = [1, 0]$. According to (E.9), a DESO is designed as follows,

$$\begin{cases} \hat{\theta}_{k+1} = X\hat{\theta}_k + Yp_k^* + \eta(y_k - \hat{y}_k), \\ \hat{r}_k = T\hat{\theta}_k, \end{cases} \quad (\text{E.10})$$

where $\hat{\theta}_k = [\hat{z}_k, \hat{d}_k]^T$, $\eta \in \mathbb{R}^{2 \times 1}$ is the observer gain, S_k is QSM surface, and $\hat{y}_k$ is the estimated value of y_k .

Define estimation error $\tilde{\theta}_k$ as:

$$\tilde{\theta}_k = \theta_k - \hat{\theta}_k. \quad (\text{E.11})$$

Similarly, from (E.4) to (E.6), the system $\tilde{\theta}_{k+1} = (X - \eta T)\tilde{\theta}_k$ is global stability. Suppose $\Delta d(k)$ is bounded, it can be concluded that estimate error $\tilde{\theta}_k$ is ultimately bounded, i.e., the DESO can estimate d_k with ultimate boundness as $\hat{d}_k$.

E.2.2 QSMC Design

According to (3.46), a DESO based-QSMC is designed as,

$$p_k^* = \frac{1}{2E}(-k_{z1}e_k - k_{z2}\text{sign}(S_k)) + \frac{1}{2}d_k, \quad (\text{E.12})$$

where $k_{z1} > 0$ and $k_{z2} > 0$ are sliding mode control gains to be designed. S_k is quasi-sliding mode surface of voltage regulation. The proof follows *Theorem 3.3.2*. The tuning parameters process for the DESO-QSMC in voltage regulation loop can be guided by *Remark E.1.1* and *Remark E.1.2*, not providing further details to save space.

List of Figures

1.1	The market size of microgrids and microgrid control systems.	2
1.2	A typical hierarchical control structure of an ac microgrid.	3
1.3	An overview of the power converters in microgrids.	3
1.4	A 3L-NPC power converter topology.	4
1.5	The key issues of 3L-NPC power converter control.	6
1.6	Control structure of a 3L-NPC Power Converter	7
1.7	PI control structure.	8
1.8	PR control structure.	8
1.9	PI controller applied in 3L-NPC power converter to control the current in microgrids.	9
1.10	PR controller applied in 3L-NPC power converter to control the current in microgrids.	10
1.11	SMC depicting sliding surface and two phases.	11
1.12	Chattering of system trajectory under delayed SMC.	11
1.13	System trajectory under SOSMC.	12
1.14	STA control structure.	13
1.15	AC structure.	15
1.16	An MFAC system.	15
1.17	A STC system.	16
1.18	The categorization of AC based on physical model.	17
1.19	A nonlinear power controller applied in a 3L-NPC power converter in microgrids.	18
1.20	A block diagram of OBC structure.	19
1.21	Typical classification of observers.	19
1.22	A full-order state observer.	21
1.23	A disturbance observer.	22
1.24	OBC applied in a 3L-NPC power converter in microgrids.	23
1.25	A block diagram of the entire workflow.	26
2.1	The topology of 3L-NPC power converter.	27
2.2	The topological structure of a -phase in 3L-NPC power converter.	28
2.3	When $i_a > 0$, the process of a -phase current change.	30
2.4	When $i_a < 0$, the process of a -phase current change.	31

2.5	A 3L-NPC power converter with ideal switches.	33
3.1	An overview of the power converters in microgrids.	42
3.2	The control block diagram of proposed DSTO-QISMCM scheme for the 3L-NPC power converter system.	44
3.3	Relationship between parameters ρ_j , ϱ_j and closed-loop pole z	54
3.4	The lab-constructed 3L-NPC power converter prototype of Chap. 3.	57
3.5	An overall evaluation comparison under v_{dc}^* is from 400V $\rightarrow$ 450V $\rightarrow$ 400V, q^* is from 0 $\rightarrow$ 1500 Var with $R_L = 120\Omega$	58
3.6	Comparative capacitor voltage regulation performance under voltage reference v_{dc}^* from 420V $\rightarrow$ 370V $\rightarrow$ 420V, $R_L=120\Omega$, and $q^* = 0$	61
3.7	Comparative power tracking performance with q^* from 0 to 2000Var back to 0, $v_{dc}^*=400V$, and $R_L=120\Omega$	62
3.8	The trajectory of voltage regulation under the voltage reference v_{dc}^* from 420V $\rightarrow$ 370V $\rightarrow$ 420V, $R_L=120\Omega$, and $q^*=0$	63
3.9	The trajectory of power tracking under the reactive power reference q^* from 0 $\rightarrow$ 2000Var $\rightarrow$ 0, $R_L=120\Omega$, and $v_{dc}^*=400V$	63
3.10	Comparative voltage regulation performance under dc-load R_L from 120 Ω to 80 Ω back to 120 Ω , $v_{dc}^* = 400V$, and $q^* = 0$	64
3.11	Capacitance parametric robustness comparison (transient-state) under (a) v_{dc}^* from 400V to 450V and R_L set at 120 Ω with $q^* = 0$; (b) q^* from 0 to 1000Var, v_{dc}^* set at 400V with R_L sustained at 120 Ω	64
3.12	Parameter mismatch comparison (steady-state) under v_{dc}^* is 400V and R_L is 120 Ω and $q^* = 0$. Capacitance mismatch: (a) DSTO-QISMCM, (b) DESO-QISMCM, and (c) DPI; Inductance mismatch: (d) DSTO-QISMCM, (e) DESO-QISMCM and (f) DPI.	66
4.1	Diagram of power converters in microgrids.	70
4.2	A cascade control scheme for the grid-connected 3L-NPC power converter with L filter.	73
4.3	The constructed 3L-NPC power converter prototype of Chap. 4.	86
4.4	An overall evaluation comparison when the capacitor voltage reference v_{dc}^* (160V $\rightarrow$ 140V $\rightarrow$ 160V), and the q^* (0 $\rightarrow$ 100 Var) with $R_L = 30\Omega$	88
4.5	Voltage regulation comparisons are conducted under varying dc-load R_L (30 Ω $\rightarrow$ 40 Ω) with $v_{dc}^* = 160V$ and $q^* = 0Var$	89
4.6	Dynamic response voltage v_{dc} with different AO gain γ : (a) when R_L is 40 Ω $\rightarrow$ 30 Ω with $v_{dc}^*=150V$, $q^* = 0Var$; and (b) when v_{dc}^* is 140V $\rightarrow$ 160V with $R_L = 30\Omega$ and $q^* = 0Var$	89
4.7	Comparisons are made for power tracking, varying q^* from 0Var to 200Var, with fixed values of $v_{dc}^* = 160V$ and $R_L = 30\Omega$	90
4.8	Comparisons of voltage balancing with q^* (0 Var $\rightarrow$ 100Var), $v_{dc}^* = 160V$, and $R_L = 30\Omega$	91
4.9	Dynamic response of v_{dc}^o with different ξ_2 for SMO gains ξ	91
4.10	Comparisons of steady-state voltage regulation performance: R_L varies from 40 Ω to 30 Ω , and again to 40 Ω with $v_{dc}^*=160 V$ and $q^* = 0 Var$	92

4.11	The filter inductance parameter mismatches: (a) work properly, (b) work properly, (c) work properly, (d) work improperly, (e) work properly, (f) work properly, (g) work grudgingly, (h) work properly, (j) work properly, (k) work improperly, (l) work properly, and (m) work improperly.	94
5.1	The schematic diagram of the power converter in MESs.	98
5.2	Grid-connected 3L-NPC power converter with L filter.	100
5.3	The lab-constructed 3L-NPC power converter prototype of Chap. 5.	109
5.4	An overall evaluation comparison.	111
5.5	Performance comparison of voltage regulation when the capacitor voltage reference v_{dc}^* steps from 400V to 450V again back to 400V.	111
5.6	Performance comparison of voltage regulation when the dc-link load steps from 120Ω to 80Ω again back to 120Ω	113
5.7	Performance comparison of current tracking i_α and i_β when reactive power reference steps from $q^*=0\text{Var}$ to $q^*=2000\text{Var}$	113
5.8	Capacitance parametric robustness comparison when the capacitor voltage reference v_{dc}^* steps from 400V to 450V.	114
5.9	Current errors performance comparison when exist inductance deviation: 1) LO-STA (max current error): (a) $i_\alpha^{err}/i_\beta^{err}=0.28\text{A}$, (b) $i_\alpha^{err}/i_\beta^{err}=0.35\text{A}$, and (c) $i_\alpha^{err}/i_\beta^{err}=0.49\text{A}$; 2) PR (max current error): (d) $i_\alpha^{err}/i_\beta^{err}=0.41\text{A}$, (e) $i_\alpha^{err}/i_\beta^{err}=0.48\text{A}$, and (f) work improperly.	114
5.10	Performance comparison of grid frequency variation of current error i_α^{err} and i_β^{err} with combined active power and reactive power injection under v_{dc}^* steps from 450V to 400V at 0.1s, then q^* steps from 0 Var to 2000Var at 0.6s, dc-load $R_L = 120\Omega$	115
6.1	Grid-connected 3L-NPC power converter with L filter.	122
6.2	The fitness function $W_{\min}$ with the different number of particles (part.) under iteration number=50, when Voltage reference $v_{dc}^*=160\text{V}$, dc-load $R_L = 30\Omega$ and reactive power reference $q^* = 0$	132
6.3	The range of dc-voltage and dq -current by particle (part.) of population = 100, 150, and 200 when iteration = 50 with $v_{dc}^*=160\text{V}$, $R_L = 30\Omega$, and $q^* = 0$	132
6.4	The lab-constructed 3L-NPC power converter prototype of Chap. 4.	133
6.5	An overall evaluation comparison under v_{dc}^* is from $160\text{V} \rightarrow 140\text{V} \rightarrow 160\text{V}$, q^* is from $0 \rightarrow 150\text{Var} \rightarrow 0$ with $R_L = 30\Omega$	135
6.6	The voltage performance comparison: (a) (Transient-state) v_{dc}^* from 160V to 140V back to 160V set at 30Ω with $q^* = 0$; (b) (Steady-state) grid current THD under 160V and 140V, respectively.	135
6.7	The reactive current performance comparison as q^* is from 0 to 200Var back to 0, and the grid-current THD comparison when the $q^*=200\text{Var}$ (steady-state), under the condition of $R_L=30\Omega$ and $v_{dc}^* = 160\text{V}$	136
6.8	Comparative voltage regulation performance under external dc-load variation: R_L from 30Ω to 40Ω back to 30Ω , $v_{dc}^* = 160\text{V}$, and $q^* = 0$	136
6.9	THD of grid-side current under inductance and capacitance mismatches under v_{dc}^* is 160V, R_L is 30Ω , and $q^* = 100\text{Var}$	139

6.10	Parameter mismatches comparison (steady-state) under v_{dc}^* is 160V and R_L is 30Ω and $q^* = 100\text{Var}$. RMS of the dc-link voltage under parameter mismatches (a) PSO-MFAC, (b) ANN-PI, and (c) STA; RMS of the q -current under parameter mismatches: (d) PSO-MFAC, (e) ANN-PI and (f) STA.	140
6.11	Comparative power tracking performance with q^* from 0 to 300Var back to 0, $v_{dc}^*=160\text{V}$, and $R_L=30\Omega$	141
C.1	The constructed 3L-NPC power converter prototype.	154
D.1	Engineering issues in 3L-NPC converter systems	164
D.2	Engineering issues of power and voltage control.	166
D.3	Core solution idea of DSMC.	167
D.4	Core solution idea of ASMC	168
D.5	Engineering issues of current and voltage control	168
D.6	Core solution idea of STC	169
D.7	Core solution idea of data-driven MFAC	170
D.8	Problems addressed and application scenarios of the control scheme	171

List of Tables

1.1	Comparison of SO Technologies for Power Converters	20
1.2	Comparison of DO Technologies for Power Converters	21
2.1	Switch State of a -phase	29
3.1	Nomenclatures	45
3.2	Parameters Used for Experiment	56
3.3	Controllers Parameters	57
3.4	Comparison Of Steady/Transient States Based On Figure3.5	59
3.5	Comparison Of Steady/Transient States Based Of Figure3.6	60
3.6	Comparison Of Steady/Transient States Based Of Figure3.7	60
4.1	List of Nomenclatures	72
4.2	Parameters Used for Experiment	87
4.3	Controller Parameters	87
4.4	Comparisons of AO-P, STO-ISM, STA and PI Methods	93
4.5	Comparisons of DO-ISM, STO-ISM and PI Methods	93
4.6	Comparisons of the SMO-P, STA and PI Methods	95
5.1	Nomenclatures	101
5.2	Parameters Used For Experiments	109
5.3	Controllers Parameters of Experiment	110
5.4	Performance Comparison of the Proposed ESMDO-P VS PI	112
5.5	Performance Comparisons of the Proposed LO-STA VS PR	112
5.6	The Current Static Performance Comparisons of both the Proposed LO-STA and the PR Control from Figure 5.10	116
6.1	List of Nomenclatures	123
6.2	PSO-MFAC Parameters	131
6.3	Parameters Used For Experiment	133
6.4	Controller Parameters	134
6.5	Comparisons of PSO-MFAC, ANN-PI, and STA Methods	134
6.6	Comparisons of PSO-MFAC, ANN-PI, and STA Methods	137

6.7	Comparisons of PSO-MFAC, STA and ANN-PI Methods	138
C.1	Key Features and Specifications of Current Sensors	155
C.2	Key Features and Specifications of Voltage Sensors	156
C.3	Components of Filters Box	156
C.4	Key Features and Specification 3L-NPC Power Converter PEN8018	158
C.5	Key Specifications of B-Box RCP ^{3.0}	160
C.6	Technical Specifications of RT-BOX	161
D.1	Control Classification of Power Converter	165

Bibliography

- [1] S. Ansari, A. Chandel, and M. Tariq, “A comprehensive review on power converters control and control strategies of AC/DC microgrid,” *IEEE Access*, vol. 9, pp. 17 998–18 015, 2020.
- [2] F. Mohammadi, B. Mohammadi-Ivatloo, G. B. Gharehpetian, M. H. Ali, W. Wei, O. Erding, and M. Shirkhani, “Robust control strategies for microgrids: A review,” *IEEE Systems Journal*, vol. 16, no. 2, pp. 2401–2412, 2021.
- [3] F. Mumtaz, N. Z. Yahaya, S. T. Meraj, B. Singh, R. Kannan, and O. Ibrahim, “Review on non-isolated DC-DC converters and their control techniques for renewable energy applications,” *Ain Shams Engineering Journal*, vol. 12, no. 4, pp. 3747–3763, 2021.
- [4] R. Bayindir, I. Colak, G. Fulli, and K. Demirtas, “Smart grid technologies and applications,” *Renewable and Sustainable Energy Reviews*, vol. 66, pp. 499–516, 2016.
- [5] G. Dileep, “A survey on smart grid technologies and applications,” *Renewable energy*, vol. 146, pp. 2589–2625, 2020.
- [6] O. M. Butt, M. Zulqarnain, and T. M. Butt, “Recent advancement in smart grid technology: Future prospects in the electrical power network,” *Ain Shams Engineering Journal*, vol. 12, no. 1, pp. 687–695, 2021.
- [7] A. Mohammed, S. S. Refaat, S. Bayhan, and H. Abu-Rub, “AC microgrid control and management strategies: Evaluation and review,” *IEEE Power Electronics Magazine*, vol. 6, no. 2, pp. 18–31, 2019.
- [8] N. T. Mbungu, A. A. Ismail, M. AlShabi, R. C. Bansal, A. Elnady, and A. K. Hamid, “Control and estimation techniques applied to smart microgrids: A review,” *Renewable and Sustainable Energy Reviews*, p. 113251, 2023.
- [9] D. E. Olivares, A. Mehrizi-Sani, A. H. Etemadi, C. A. Cañizares, R. Iravani, M. Kazerani, A. H. Hajimiragha, O. Gomis-Bellmunt, M. Saeedifard, R. Palma-Behnke *et al.*, “Trends in microgrid control,” *IEEE Transactions on Smart Grid*, vol. 5, no. 4, pp. 1905–1919, 2014.

- [10] E. Rokrok, M. Shafie-Khah, and J. P. Catalão, “Review of primary voltage and frequency control methods for inverter-based islanded microgrids with distributed generation,” *Renewable and Sustainable Energy Reviews*, vol. 82, pp. 3225–3235, 2018.
- [11] S. S. Madani, C. Kammer, and A. Karimi, “Data-driven distributed combined primary and secondary control in microgrids,” *IEEE Transactions on Control Systems Technology*, vol. 29, no. 3, pp. 1340–1347, 2020.
- [12] J. Liu, Y. Miura, and T. Ise, “Cost-function-based microgrid decentralized control of unbalance and harmonics for simultaneous bus voltage compensation and current sharing,” *IEEE Transactions on Power Electronics*, vol. 34, no. 8, pp. 7397–7410, 2018.
- [13] F. Gao, R. Kang, J. Cao, and T. Yang, “Primary and secondary control in DC microgrids: a review,” *Journal of Modern Power Systems and Clean Energy*, vol. 7, no. 2, pp. 227–242, 2019.
- [14] F. Garcia-Torres, C. Bordons, and M. A. Ridao, “Optimal economic schedule for a network of microgrids with hybrid energy storage system using distributed model predictive control,” *IEEE Transactions on Industrial Electronics*, vol. 66, no. 3, pp. 1919–1929, 2018.
- [15] Y. Ma, M. Zhang, H. Yang, X. Wang, J. Xu, and X. Hu, “Decentralized and coordinated scheduling model of interconnected multi-microgrid based on virtual energy storage,” *International Journal of Electrical Power & Energy Systems*, vol. 148, p. 108990, 2023.
- [16] A. Frances, R. Asensi, O. Garcia, R. Prieto, and J. Uceda, “Modeling electronic power converters in smart DC microgrids—an overview,” *IEEE Transactions on Smart Grid*, vol. 9, no. 6, pp. 6274–6287, 2017.
- [17] J. Rocabert, A. Luna, F. Blaabjerg, and P. Rodriguez, “Control of power converters in AC microgrids,” *IEEE Transactions on Power Electronics*, vol. 27, no. 11, pp. 4734–4749, 2012.
- [18] J. Hu, Y. Shan, K. W. Cheng, and S. Islam, “Overview of power converter control in microgrids—challenges, advances, and future trends,” *IEEE Transactions on Power Electronics*, vol. 37, no. 8, pp. 9907–9922, 2022.
- [19] R. Portillo, S. Vazquez, J. I. Leon, M. M. Prats, and L. G. Franquelo, “Model based adaptive direct power control for three-level NPC converters,” *IEEE Transactions on Industrial Informatics*, vol. 9, no. 2, pp. 1148–1157, 2012.
- [20] L. Liu, Z. Zhang, Y. Yin, Y. Li, H. Xie, M. Zhang, Y. Zhao, and R. Kennel, “A robust high-quality current control with fast convergence for three-level NPC converters in micro-energy systems,” *IEEE Transactions on Industrial Informatics*, vol. 11, no. 19, pp. 10 716–10 726, 2023.
- [21] L. Liu, Z. Zhang, Y. Yin, S. Vazquez, Y. Zhao, and R. Kennel, “An efficient robust power–voltage control for three-level NPC converters in microgrids,” *IEEE Transactions on Industrial Informatics*, vol. 20, no. 4, pp. 5849–5863, 2023.

- [22] L. Liu, Y. Yin, Z. Zhang, H. Chen, H. Xie, Y. Liu, Y. Zhao, and R. Kennel, "Multidisturbances compensation for three-level NPC converters in microgrids: A robust adaptive sliding mode control approach," *IEEE Transactions on Industrial Informatics*, vol. 20, no. 6, pp. 8917–8931, 2024.
- [23] L. Yacoubi, K. Al-Haddad, L.-A. Dessaint, and F. Fnaiech, "Linear and nonlinear control techniques for a three-phase three-level NPC boost rectifier," *IEEE Transactions on Industrial Electronics*, vol. 53, no. 6, pp. 1908–1918, 2006.
- [24] K. H. Ang, G. Chong, and Y. Li, "PID control system analysis, design, and technology," *IEEE Transactions on Control Systems Technology*, vol. 13, no. 4, pp. 559–576, 2005.
- [25] A. Visioli, *Practical PID control*. Springer Science & Business Media, 2006.
- [26] A. Tamaarat and A. Benakcha, "Performance of PI controller for control of active and reactive power in DFIG operating in a grid-connected variable speed wind energy conversion system," *Frontiers in Energy*, vol. 8, pp. 371–378, 2014.
- [27] S. Isik, M. Alharbi, and S. Bhattacharya, "An optimized circulating current control method based on PR and PI controller for MMC applications," *IEEE Transactions on Industry Applications*, vol. 57, no. 5, pp. 5074–5085, 2021.
- [28] R. Chattopadhyay, A. De, and S. Bhattacharya, "Comparison of PR controller and damped PR controller for grid current control of LCL filter based grid-tied inverter under frequency variation and grid distortion," in *2014 IEEE Energy Conversion Congress and Exposition (ECCE)*. IEEE, 2014, pp. 3634–3641.
- [29] M. Pérez, R. Ortega, and J. R. Espinoza, "Passivity-based PI control of switched power converters," *IEEE Transactions on Control Systems Technology*, vol. 12, no. 6, pp. 881–890, 2004.
- [30] K. Ahmed, N. Yahaya, V. Asirvadam, K. Ramani, and O. Ibrahim, "Comparison of fuzzy logic control and PI control for a three-level rectifier based on voltage oriented control," in *2016 IEEE International Conference on Power and Energy (PECon)*. IEEE, 2016, pp. 127–132.
- [31] R. Priewasser, M. Agostinelli, C. Unterrieder, S. Marsili, and M. Huemer, "Modeling, control, and implementation of DC–DC converters for variable frequency operation," *IEEE Transactions on Power Electronics*, vol. 29, no. 1, pp. 287–301, 2013.
- [32] M. Husain and I. Khan, "Ripple free AC-DC converter with PID controlled power decoupling method," in *2020 8th International Conference on Reliability, Infocom Technologies and Optimization (Trends and Future Directions)(ICRITO)*. IEEE, 2020, pp. 186–191.
- [33] G. Shen, X. Zhu, J. Zhang, and D. Xu, "A new feedback method for PR current control of LCL-filter-based grid-connected inverter," *IEEE Transactions on Industrial Electronics*, vol. 57, no. 6, pp. 2033–2041, 2010.

- [34] J. Gnanavadivel, R. Thangasankaran, N. S. Kumar, K. Krishnaveni, and M. Schlenk, "Performance analysis of PI controller and PR controller based three-phase AC-DC boost converter with space vector PWM," *International Journal of Pure and Applied Mathematics*, vol. 118, no. 24, pp. 1–16, 2018.
- [35] S. V. Drakunov and V. I. Utkin, "Sliding mode control in dynamic systems," *International Journal of Control*, vol. 55, no. 4, pp. 1029–1037, 1992.
- [36] V. I. Utkin, "Sliding mode control in discrete-time and difference systems," *Variable Structure and Lyapunov control*, pp. 87–107, 2005.
- [37] V. Utkin and H. Lee, "Chattering problem in sliding mode control systems," in *International Workshop on Variable Structure Systems, 2006. VSS'06*. IEEE, 2006, pp. 346–350.
- [38] I. D. Landau, R. Lozano, M. M'Saad *et al.*, *Adaptive control*. Springer New York, 1998, vol. 51.
- [39] I. D. Landau, R. Lozano, M. M'Saad, and A. Karimi, *Adaptive control: algorithms, analysis and applications*. Springer Science & Business Media, 2011.
- [40] R. Jager, *Fuzzy logic in control*. Rene Jager, 1995.
- [41] Z. Kovacic and S. Bogdan, *Fuzzy controller design: theory and applications*. CRC Press, 2018.
- [42] S. J. Qin and T. A. Badgwell, "An overview of industrial model predictive control technology," in *Alche symposium series*, vol. 93, no. 316. New York, NY: American Institute of Chemical Engineers, 1971-c2002., 1997, pp. 232–256.
- [43] L. Grüne, J. Pannek, L. Grüne, and J. Pannek, *Nonlinear model predictive control*. Springer, 2017.
- [44] W. H. Heemels, M. Donkers, and A. R. Teel, "Periodic event-triggered control for linear systems," *IEEE Transactions on Automatic Control*, vol. 58, no. 4, pp. 847–861, 2012.
- [45] W. P. Heemels, K. H. Johansson, and P. Tabuada, "An introduction to event-triggered and self-triggered control," in *2012 IEEE 51st IEEE Conference on Decision and Control (CDC)*. IEEE, 2012, pp. 3270–3285.
- [46] W. T. Miller, R. S. Sutton, and P. J. Werbos, *Neural networks for control*. MIT press, 1995.
- [47] T. B. Lopez-Garcia, A. Coronado-Mendoza, and J. A. Domínguez-Navarro, "Artificial neural networks in microgrids: A review," *Engineering Applications of Artificial Intelligence*, vol. 95, p. 103894, 2020.
- [48] Y. Shtessel, C. Edwards, L. Fridman, A. Levant *et al.*, *Sliding mode control and observation*. Springer, 2014, vol. 10.

- [49] Y. Pan, C. Yang, L. Pan, and H. Yu, "Integral sliding mode control: performance, modification, and improvement," *IEEE Transactions on Industrial Informatics*, vol. 14, no. 7, pp. 3087–3096, 2017.
- [50] A. Bartoszewicz, "Discrete-time quasi-sliding-mode control strategies," *IEEE Transactions on Industrial Electronics*, vol. 45, no. 4, pp. 633–637, 1998.
- [51] M.-L. Tseng and M.-S. Chen, "Chattering reduction of sliding mode control by low-pass filtering the control signal," *Asian Journal of control*, vol. 12, no. 3, pp. 392–398, 2010.
- [52] J. Zhang, X. Liu, Y. Xia, Z. Zuo, and Y. Wang, "Disturbance observer-based integral sliding-mode control for systems with mismatched disturbances," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 11, pp. 7040–7048, 2016.
- [53] V. Utkin, "Discussion aspects of high-order sliding mode control," *IEEE Transactions on Automatic Control*, vol. 61, no. 3, pp. 829–833, 2015.
- [54] G. Bartolini, A. Ferrara, A. Levant, and E. Usai, "On second order sliding mode controllers," *Variable Structure Systems, Sliding Mode and Nonlinear Control*, pp. 329–350, 1999.
- [55] A. Chalanga, S. Kamal, L. M. Fridman, B. Bandyopadhyay, and J. A. Moreno, "Implementation of super-twisting control: Super-twisting and higher order sliding-mode observer-based approaches," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 6, pp. 3677–3685, 2016.
- [56] V. Utkin, J. Guldner, and J. Shi, *Sliding mode control in electro-mechanical systems*. CRC press, 2017.
- [57] Y. M. Alsmadi, V. Utkin, M. A. Haj-ahmed, and L. Xu, "Sliding mode control of power converters: DC/DC converters," *International Journal of Control*, vol. 91, no. 11, pp. 2472–2493, 2018.
- [58] J. J. Moré, P. F. Puleston, C. Kunusch, and M. A. Fantova, "Development and implementation of a supervisor strategy and sliding mode control setup for fuel-cell-based hybrid generation systems," *IEEE Transactions on Energy Conversion*, vol. 30, no. 1, pp. 218–225, 2014.
- [59] A. Musa, L. R. Sabug, and A. Monti, "Robust predictive sliding mode control for multiterminal HVDC grids," *IEEE Transactions on Power Delivery*, vol. 33, no. 4, pp. 1545–1555, 2018.
- [60] C. Lascu, "Sliding-mode direct-voltage control of voltage-source converters with LC filters for pulsed power loads," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 12, pp. 11 642–11 650, 2020.
- [61] V. Utkin, "Sliding mode control of DC/DC converters," *Journal of the Franklin Institute*, vol. 350, no. 8, pp. 2146–2165, 2013.

- [62] J. Hu, L. Shang, Y. He, and Z. Zhu, "Direct active and reactive power regulation of grid-connected DC/AC converters using sliding mode control approach," *IEEE Transactions on Power Electronics*, vol. 26, no. 1, pp. 210–222, 2010.
- [63] H. R. Baghaee, M. Mirsalim, G. B. Gharehpetian, and H. A. Talebi, "A decentralized power management and sliding mode control strategy for hybrid AC/DC microgrids including renewable energy resources," *IEEE Transactions on Industrial Informatics*, 2017.
- [64] L. G. G. de Vicuna, M. Castilla, J. Miret, J. Matas, and J. M. Guerrero, "Sliding-mode control for a single-phase AC/AC quantum resonant converter," *IEEE Transactions on Industrial Electronics*, vol. 56, no. 9, pp. 3496–3504, 2009.
- [65] M. López, L. G. de Vicuña, M. Castilla, P. Gayà, and O. López, "Current distribution control design for paralleled DC/DC converters using sliding-mode control," *IEEE Transactions on Industrial Electronics*, vol. 51, no. 2, pp. 419–428, 2004.
- [66] I.-C. Baik, K.-H. Kim, and M.-J. Youn, "Robust nonlinear speed control of PM synchronous motor using boundary layer integral sliding mode control technique," *IEEE Transactions on Control Systems Technology*, vol. 8, no. 1, pp. 47–54, 2000.
- [67] W. Yanmin, C. Yuqing, and X. Hongwei, "Voltage and current double closed-loop terminal sliding mode control of buck converter," *J Electr Mach Control*, vol. 20, no. 8, pp. 92–97, 2016.
- [68] B. Long, P. J. Lu, K. T. Chong, J. Rodriguez, and J. M. Guerrero, "Robust fuzzy-fractional-order nonsingular terminal sliding-mode control of LCL-type grid-connected converters," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 6, pp. 5854–5866, 2021.
- [69] R. Ramos, D. Biel, E. Fossas, and F. Guinjoan, "Interleaving quasi-sliding-mode control of parallel-connected buck-based inverters," *IEEE Transactions on Industrial Electronics*, vol. 55, no. 11, pp. 3865–3873, 2008.
- [70] C. Zheng, T. Dragičević, J. Zhang, R. Chen, and F. Blaabjerg, "Composite robust quasi-sliding mode control of DC–DC buck converter with constant power loads," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 9, no. 2, pp. 1455–1464, 2020.
- [71] S. Ding, W. X. Zheng, J. Sun, and J. Wang, "Second-order sliding-mode controller design and its implementation for buck converters," *IEEE Transactions on Industrial Informatics*, vol. 14, no. 5, pp. 1990–2000, 2017.
- [72] R. Sadeghi, S. M. Madani, M. Ataei, M. A. Kashkooli, and S. Ademi, "Super-twisting sliding mode direct power control of a brushless doubly fed induction generator," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 11, pp. 9147–9156, 2018.
- [73] S. M. Rakhtala and A. Casavola, "Real-time voltage control based on a cascaded super twisting algorithm structure for DC–DC converters," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 1, pp. 633–641, 2021.

- [74] P. A. Ioannou and J. Sun, *Robust adaptive control*. PTR Prentice-Hall Upper Saddle River, NJ, 1996, vol. 1.
- [75] E. Lavretsky, “Combined/composite model reference adaptive control,” *IEEE Transactions on Automatic Control*, vol. 54, no. 11, pp. 2692–2697, 2009.
- [76] G. Joshi and G. Chowdhary, “Hybrid direct-indirect adaptive control of nonlinear system with unmatched uncertainty,” in *2019 6th International Conference on Control, Decision and Information Technologies (CoDIT)*. IEEE, 2019, pp. 127–132.
- [77] A. Navarro-Rodriguez, P. Garcia, R. Georgious, and J. García, “Adaptive active power sharing techniques for DC and AC voltage control in a hybrid DC/AC microgrid,” *IEEE Transactions on Industry Applications*, vol. 55, no. 2, pp. 1106–1116, 2018.
- [78] H. Sorouri, M. Sedighizadeh, A. Oshnoei, and R. Khezri, “An intelligent adaptive control of DC–DC power buck converters,” *International Journal of Electrical Power & Energy Systems*, vol. 141, p. 108099, 2022.
- [79] S. K. Panda and B. Subudhi, “A review on robust and adaptive control schemes for microgrid,” *Journal of Modern Power Systems and Clean Energy*, vol. 11, no. 4, pp. 1027–1040, 2022.
- [80] E. Giraldo and A. Garces, “An adaptive control strategy for a wind energy conversion system based on PWM-CSC and PMSG,” *IEEE Transactions on Power Systems*, vol. 29, no. 3, pp. 1446–1453, 2013.
- [81] S. Eren, M. Pahlevani, A. Bakhshai, and P. Jain, “A digital current control technique for grid-connected AC/DC converters used for energy storage systems,” *IEEE Transactions on Power Electronics*, vol. 32, no. 5, pp. 3970–3988, 2016.
- [82] J. Li, Z. Liu, and Q. Su, “Improved adaptive backstepping sliding mode control for a three-phase pwm AC–DC converter,” *IET Control Theory & Applications*, vol. 13, no. 6, pp. 854–860, 2019.
- [83] A. I. Dounis, P. Kofinas, G. Papadakis, and C. Alafodimos, “A direct adaptive neural control for maximum power point tracking of photovoltaic system,” *Solar Energy*, vol. 115, pp. 145–165, 2015.
- [84] H. Zhang, J. Zhou, Q. Sun, J. M. Guerrero, and D. Ma, “Data-driven control for inter-linked AC/DC microgrids via model-free adaptive control and dual-droop control,” *IEEE Transactions on Smart Grid*, vol. 8, no. 2, pp. 557–571, 2015.
- [85] X. Liu, L. Qiu, Y. Fang, and J. Rodríguez, “Predictor-based data-driven model-free adaptive predictive control of power converters using machine learning,” *IEEE Transactions on Industrial Electronics*, vol. 70, no. 8, pp. 7591–7603, 2022.
- [86] F. Donoso, A. Mora, R. Cardenas, A. Angulo, D. Saez, and M. Rivera, “Finite-set model-predictive control strategies for a 3L-NPC inverter operating with fixed switching frequency,” *IEEE Transactions on Industrial Electronics*, vol. 65, no. 5, pp. 3954–3965, 2017.

- [87] M. Siami, D. A. Khaburi, and J. Rodriguez, "Simplified finite control set-model predictive control for matrix converter-fed PMSM drives," *IEEE Transactions on Power Electronics*, vol. 33, no. 3, pp. 2438–2446, 2017.
- [88] Y. Fu, Z. Zhang, Y. Mi, Z. Li, and F. Li, "Droop control for DC multi-microgrids based on local adaptive fuzzy approach and global power allocation correction," *IEEE Transactions on Smart Grid*, vol. 10, no. 5, pp. 5468–5478, 2018.
- [89] C. Balasundar, C. K. Sundarabalan, N. Srinath, J. Sharma, and J. M. Guerrero, "Interval type2 fuzzy logic-based power sharing strategy for hybrid energy storage system in solar powered charging station," *IEEE Transactions on Vehicular Technology*, vol. 70, no. 12, pp. 12 450–12 461, 2021.
- [90] S. Li, M. Fairbank, C. Johnson, D. C. Wunsch, E. Alonso, and J. L. Proao, "Artificial neural networks for control of a grid-connected rectifier/inverter under disturbance, dynamic and power converter switching conditions," *IEEE Transactions on Neural Networks and Learning Systems*, vol. 25, no. 4, pp. 738–750, 2013.
- [91] X. Fu and S. Li, "Control of single-phase grid-connected converters with LCL filters using recurrent neural network and conventional control methods," *IEEE Transactions on Power Electronics*, vol. 31, no. 7, pp. 5354–5364, 2015.
- [92] J. Liu, S. Vazquez, L. Wu, A. Marquez, H. Gao, and L. G. Franquelo, "Extended state observer-based sliding-mode control for three-phase power converters," *IEEE Transactions on Industrial Electronics*, vol. 64, no. 1, pp. 22–31, 2016.
- [93] S. Zhuo, A. Gaillard, L. Xu, D. Paire, and F. Gao, "Extended state observer-based control of DC–DC converters for fuel cell application," *IEEE Transactions on Power Electronics*, vol. 35, no. 9, pp. 9923–9932, 2020.
- [94] D. Das, S. Madichetty, B. Singh, and S. Mishra, "Luenberger observer based current estimated boost converter for PV maximum power extraction—a current sensorless approach," *IEEE Journal of Photovoltaics*, vol. 9, no. 1, pp. 278–286, 2018.
- [95] Y. Yin, J. Liu, J. A. Sánchez, L. Wu, S. Vazquez, J. I. Leon, and L. G. Franquelo, "Observer-based adaptive sliding mode control of NPC converters: An RBF neural network approach," *IEEE Transactions on Power Electronics*, vol. 34, no. 4, pp. 3831–3841, 2018.
- [96] B. Yang, Y. Sang, K. Shi, W. Yao, L. Jiang, and T. Yu, "Design and real-time implementation of perturbation observer based sliding-mode control for VSC-HVDC systems," *Control Engineering Practice*, vol. 56, pp. 13–26, 2016.
- [97] H. H. Alhelou, M. E. H. Golshan, and N. D. Hatziargyriou, "Deterministic dynamic state estimation-based optimal LFC for interconnected power systems using unknown input observer," *IEEE Transactions on Smart Grid*, vol. 11, no. 2, pp. 1582–1592, 2019.
- [98] Y. Du, W. Cao, J. She, M. Wu, M. Fang, and S. Kawata, "Disturbance rejection and robustness of improved equivalent-input-disturbance-based system," *IEEE Transactions on Cybernetics*, vol. 52, no. 8, pp. 8537–8546, 2021.

- [99] J. Yang, W.-H. Chen, and S. Li, "Autopilot design of bank-to-turn missiles using state-space disturbance observers," pp. 1218–1223, 2010.
- [100] H. Yang, Y. Zhang, J. Liang, J. Gao, P. D. Walker, and N. Zhang, "Sliding-mode observer based voltage-sensorless model predictive power control of PWM rectifier under unbalanced grid conditions," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 7, pp. 5550–5560, 2017.
- [101] L. Liu, H. Xie, D. Kong, Y. Yin, Y. Zhao, Z. Zhang, and R. Kennel, "A disturbance observer-based integral sliding mode control for two-level power converter," in *2021 IEEE International Conference on Predictive Control of Electrical Drives and Power Electronics (PRECEDE)*. IEEE, 2021, pp. 987–992.
- [102] A. T. Woldegiorgis, X. Ge, S. Li, and M. Hassan, "Extended sliding mode disturbance observer-based sensorless control of IPMSM for medium and high-speed range considering railway application," *IEEE Access*, vol. 7, pp. 175 302–175 312, 2019.
- [103] G. Escobar, J. Leyva-Ramos, J. Carrasco, E. Galvan, R. Portillo, M. Prats, and L. Franquelo, "Modeling of a three level converter used in a synchronous rectifier application," in *2004 IEEE 35th Annual Power Electronics Specialists Conference (IEEE Cat. No. 04CH37551)*, vol. 6. IEEE, 2004, pp. 4306–4311.
- [104] R. C. Portillo, M. M. Prats, J. I. Leon, J. A. Sanchez, J. M. Carrasco, E. Galvan, and L. G. Franquelo, "Modeling strategy for back-to-back three-level converters applied to high-power wind turbines," *IEEE Transactions on Industrial Electronics*, vol. 53, no. 5, pp. 1483–1491, 2006.
- [105] H. Abu-Rub, J. Holtz, J. Rodriguez, and G. Baoming, "Medium-voltage multilevel converters—state of the art, challenges, and requirements in industrial applications," *IEEE Transactions on Industrial Electronics*, vol. 57, no. 8, pp. 2581–2596, 2010.
- [106] Z. Ye, Y. Xu, X. Wu, G. Tan, X. Deng, and Z. Wang, "A simplified pwm strategy for a neutral-point-clamped (NPC) three-level converter with unbalanced DC links," *IEEE Transactions on Power Electronics*, vol. 31, no. 4, pp. 3227–3238, 2015.
- [107] A. Lewicki, D. Kondratenko, and C. I. Odeh, "Hybridized PWM strategy for three-and multiphase three-level NPC inverters," *IEEE Transactions on Industrial Electronics*, 2023.
- [108] B. P. McGrath and D. G. Holmes, "Multicarrier PWM strategies for multilevel inverters," *IEEE Transactions on Industrial Electronics*, vol. 49, no. 4, pp. 858–867, 2002.
- [109] N. V. Nho and M.-J. Youn, "Comprehensive study on space-vector-PWM and carrier-based-PWM correlation in multilevel invertors," *IEE Proceedings-Electric Power Applications*, vol. 153, no. 1, pp. 149–158, 2006.
- [110] A. Lewicki, Z. Krzeminski, and H. Abu-Rub, "Space-vector pulsewidth modulation for three-level NPC converter with the neutral point voltage control," *IEEE Transactions on Industrial Electronics*, vol. 58, no. 11, pp. 5076–5086, 2011.

- [111] T. E. Shults, O. Husev, F. Blaabjerg, C. Roncero-Clemente, E. Romero-Cadaval, and D. Vinnikov, "Novel space vector pulsewidth modulation strategies for single-phase three-level NPC impedance-source inverters," *IEEE Transactions on Power Electronics*, vol. 34, no. 5, pp. 4820–4830, 2018.
- [112] S. B. Monge, S. Somavilla, J. Bordonau, and D. Boroyevich, "Capacitor voltage balance for the neutral-point-clamped converter using the virtual space vector concept with optimized spectral performance," *IEEE Transactions on Power Electronics*, vol. 22, no. 4, pp. 1128–1135, 2007.
- [113] X. Wu, G. Tan, Z. Ye, G. Yao, Z. Liu, and G. Liu, "Virtual-space-vector PWM for a three-level neutral-point-clamped inverter with unbalanced DC-links," *IEEE Transactions on Power Electronics*, vol. 33, no. 3, pp. 2630–2642, 2017.
- [114] S. R. Pulikanti, M. S. Dahidah, and V. G. Agelidis, "Voltage balancing control of three-level active NPC converter using SHE-PWM," *IEEE Transactions on Power Delivery*, vol. 26, no. 1, pp. 258–267, 2010.
- [115] A. Sanchez-Ruiz, M. Mazuela, H. Fernandez-Rebolleda, S. Ceballos, A. Perez-Basante, I. Ibanez-Hidalgo, M. Zubiaga, J. Pou, N. Beniwal, and G. Konstantinou, "Dc-link neutral point control for 3L-NPC converters utilizing selective harmonic elimination-PWM," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 9, pp. 8633–8644, 2021.
- [116] S. Mariéthoz and M. Morari, "Explicit model-predictive control of a PWM inverter with an LCL filter," *IEEE Transactions on Industrial Electronics*, vol. 56, no. 2, pp. 389–399, 2008.
- [117] Y. Zhang, W. Xie, Z. Li, and Y. Zhang, "Model predictive direct power control of a PWM rectifier with duty cycle optimization," *IEEE Transactions on Power Electronics*, vol. 28, no. 11, pp. 5343–5351, 2013.
- [118] H. Komurcugil, S. Biricik, and N. Guler, "Indirect sliding mode control for DC–DC SEPIC converters," *IEEE Transactions on Industrial Informatics*, vol. 16, no. 6, pp. 4099–4108, 2019.
- [119] A. Al Zawaideh and I. M. Boiko, "Analysis of stability and performance of a cascaded PI sliding-mode control DC-DC boost converter via LPRS," *IEEE Transactions on Power Electronics*, vol. 37, no. 9, pp. 10 455–10 465, 2022.
- [120] N. M. Dehkordi, N. Sadati, and M. Hamzeh, "A robust backstepping high-order sliding mode control strategy for grid-connected DG units with harmonic/interharmonic current compensation capability," *IEEE Transactions on Sustainable Energy*, vol. 8, no. 2, pp. 561–572, 2016.
- [121] J.-S. Han, T.-I. Kim, T.-H. Oh, S.-H. Lee *et al.*, "Effective disturbance compensation method under control saturation in discrete-time sliding mode control," *IEEE Transactions on Industrial Electronics*, vol. 67, no. 7, pp. 5696–5707, 2019.

- [122] S. Li, H. Du, and X. Yu, "Discrete-time terminal sliding mode control systems based on Euler's discretization," *IEEE Transactions on Automatic Control*, vol. 59, no. 2, pp. 546–552, 2013.
- [123] S. Sarpturk, Y. Istefanopulos, and O. Kaynak, "On the stability of discrete-time sliding mode control systems," *IEEE Transactions on Automatic Control*, vol. 32, no. 10, pp. 930–932, 1987.
- [124] H. Du, G. Wen, Y. Cheng, W. Lu, and T. Huang, "Designing discrete-time sliding mode controller with mismatched disturbances compensation," *IEEE Transactions on Industrial Informatics*, vol. 16, no. 6, pp. 4109–4118, 2019.
- [125] C. Zheng, Z. Gong, X. Wu, T. Dragičević, J. Rodriguez, and F. Blaabjerg, "Finite-set quasi-sliding mode predictive control of LC-filtered voltage source inverters," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 12, pp. 11 968–11 978, 2021.
- [126] T. V. Tran, K.-H. Kim, and J.-S. Lai, "Optimized active disturbance rejection control with resonant extended state observer for grid voltage sensorless LCL-filtered inverter," *IEEE Transactions on Power Electronics*, vol. 36, no. 11, pp. 13 317–13 331, 2021.
- [127] X. Wang and E. E. Yaz, "Second-order fault tolerant extended Kalman filter for discrete time nonlinear systems," *IEEE Transactions on Automatic Control*, vol. 64, no. 12, pp. 5086–5093, 2019.
- [128] T. Bernardes, V. F. Montagner, H. A. Gründling, and H. Pinheiro, "Discrete-time sliding mode observer for sensorless vector control of permanent magnet synchronous machine," *IEEE Transactions on Industrial Electronics*, vol. 61, no. 4, pp. 1679–1691, 2013.
- [129] Y. Zhang, J. Jin, and L. Huang, "Model-free predictive current control of PMSM drives based on extended state observer using ultralocal model," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 2, pp. 993–1003, 2020.
- [130] G. S. da Silva, R. P. Vieira, and C. Rech, "Discrete-time sliding-mode observer for capacitor voltage control in modular multilevel converters," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 1, pp. 876–886, 2017.
- [131] H. Akagi, E. H. Watanabe, and M. Aredes, *Instantaneous power theory and applications to power conditioning*. John Wiley & Sons, 2017.
- [132] S. Vazquez, J. A. Sanchez *et al.*, "A model-based direct power control for three-phase power converters," *IEEE Transactions on Industrial Electronics*, vol. 55, no. 4, pp. 1647–1657, 2008.
- [133] J. Davila, L. Fridman, and A. Levant, "Second-order sliding-mode observer for mechanical systems," *IEEE Transactions on Automatic Control*, vol. 50, no. 11, pp. 1785–1789, 2005.
- [134] C. Zheng, T. Dragičević, B. Majmunović, and F. Blaabjerg, "Constrained modulated model-predictive control of an LC-filtered voltage-source converter," *IEEE Transactions on Power Electronics*, vol. 35, no. 2, pp. 1967–1977, 2019.

- [135] Y. Yin, L. Liu, S. Vazquez, R. Xu, Z. Dong, J. Liu, J. I. Leon, L. Wu, and L. G. Franquelo, "Disturbance and uncertainty attenuation for speed regulation of PMSM servo system using adaptive optimal control strategy," *IEEE Transactions on Transportation Electrification*, vol. 9, no. 2, pp. 3410–3420, 2022.
- [136] W. Kang, M. Chen, Y. Guan, L. Tang, J. M. Guerrero *et al.*, "Distributed event-triggered optimal control method for heterogeneous energy storage systems in smart grid," *IEEE Transactions on Sustainable Energy*, vol. 13, no. 4, pp. 1944–1956, 2022.
- [137] D. Wang, Z. J. Shen, X. Yin, S. Tang, X. Liu, C. Zhang, J. Wang, J. Rodriguez, and M. Norambuena, "Model predictive control using artificial neural network for power converters," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 4, pp. 3689–3699, 2021.
- [138] X. Liu, L. Qiu, Y. Fang, K. Wang, Y. Li, and J. Rodriguez, "A fuzzy approximation for FCS-MPC in power converters," *IEEE Transactions on Power Electronics*, vol. 37, no. 8, pp. 9153–9163, 2022.
- [139] F. Sebaaly, H. Vahedi, H. Y. Kanaan, N. Moubayed, and K. Al-Haddad, "Design and implementation of space vector modulation-based sliding mode control for grid-connected 3L-NPC inverter," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 12, pp. 7854–7863, 2016.
- [140] J. Han, "From PID to active disturbance rejection control," *IEEE Transactions on Industrial Electronics*, vol. 56, no. 3, pp. 900–906, 2009.
- [141] W. Wu, L. Qiu, X. Liu, F. Guo, J. Rodriguez, J. Ma, and Y. Fang, "Data-driven iterative learning predictive control for power converters," *IEEE Transactions on Power Electronics*, vol. 37, no. 12, pp. 14 028–14 033, 2022.
- [142] L. Qu, W. Qiao, and L. Qu, "An extended-state-observer-based sliding-mode speed control for permanent-magnet synchronous motors," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 9, no. 2, pp. 1605–1613, 2020.
- [143] J. Lu, S. Golestan, M. Savaghebi, J. C. Vasquez, J. M. Guerrero, and A. Marzabal, "An enhanced state observer for dc-link voltage control of three-phase ac/dc converters," *IEEE Transactions on Power Electronics*, vol. 33, no. 2, pp. 936–942, 2017.
- [144] F. Umbria, F. Gordillo, F. Gomez-Estern, F. Salas, R. C. Portillo, and S. Vazquez, "Voltage balancing in three-level neutral-point-clamped converters via luenberger observer," *Control Engineering Practice*, vol. 25, pp. 36–44, 2014.
- [145] J. Yang, S. Li, and X. Yu, "Sliding-mode control for systems with mismatched uncertainties via a disturbance observer," *IEEE Transactions on Industrial Electronics*, vol. 60, no. 1, pp. 160–169, 2012.
- [146] S. Kuppusamy and Y. H. Joo, "Memory-based integral sliding-mode control for T–S fuzzy systems with pmsm via disturbance observer," *IEEE Transactions on Cybernetics*, vol. 51, no. 5, pp. 2457–2465, 2019.

- [147] K. Shi, X. Yin, L. Jiang, Y. Liu, Y. Hu, and H. Wen, "Perturbation estimation based nonlinear adaptive power decoupling control for DFIG wind turbine," *IEEE Transactions on Power Electronics*, vol. 35, no. 1, pp. 319–333, 2019.
- [148] H. K. Khalil, *Nonlinear control*. Pearson New York, 2015, vol. 406.
- [149] S. Li, J. Yang, W.-H. Chen, and X. Chen, *Disturbance observer-based control: methods and applications*. CRC press, 2014.
- [150] Y. Yin, S. Vazquez, A. Marquez, J. Liu, J. I. Leon, L. Wu, and L. G. Franquelo, "Observer-based sliding-mode control for grid-connected power converters under unbalanced grid conditions," *IEEE Transactions on Industrial Electronics*, vol. 69, no. 1, pp. 517–527, 2021.
- [151] J. Sun, "Small-signal methods for ac distributed power systems—a review," *IEEE Transactions on Power Electronics*, vol. 24, no. 11, pp. 2545–2554, 2009.
- [152] A. Rygg, M. Molinas, C. Zhang, and X. Cai, "A modified sequence-domain impedance definition and its equivalence to the dq-domain impedance definition for the stability analysis of AC power electronic systems," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 4, no. 4, pp. 1383–1396, 2016.
- [153] W. Luo, S. Vazquez, J. Liu, F. Gordillo, L. G. Franquelo, and L. Wu, "Control system design of a three-phase active front end using a sliding-mode observer," *IEEE Transactions on Systems, Man, and Cybernetics: Systems*, 2020.
- [154] D. Dong, T. Thacker, R. Burgos, F. Wang, and D. Boroyevich, "On zero steady-state error voltage control of single-phase PWM inverters with different load types," *IEEE Transactions on Power Electronics*, vol. 26, no. 11, pp. 3285–3297, 2011.
- [155] N. Altin, S. Ozdemir, H. Komurcugil, and I. Sefa, "Sliding-mode control in natural frame with reduced number of sensors for three-phase grid-tied LCL-interfaced inverters," *IEEE Transactions on Industrial Electronics*, vol. 66, no. 4, pp. 2903–2913, 2018.
- [156] R. P. Vieira, L. T. Martins, J. R. Massing, and M. Stefanello, "Sliding mode controller in a multiloop framework for a grid-connected VSI with LCL filter," *IEEE Transactions on Industrial Electronics*, vol. 65, no. 6, pp. 4714–4723, 2017.
- [157] M. Zhou, S. Cheng, Y. Feng, W. Xu, L. Wang, and W. Cai, "Full-order terminal sliding-mode-based sensorless control of induction motor with gain adaptation," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 10, no. 2, pp. 1978–1991, 2021.
- [158] L. Qu, W. Qiao, and L. Qu, "Active-disturbance-rejection-based sliding-mode current control for permanent-magnet synchronous motors," *IEEE Transactions on Power Electronics*, vol. 36, no. 1, pp. 751–760, 2020.
- [159] A. Saghafinia, H. W. Ping, M. N. Uddin, and K. S. Gaeid, "Adaptive fuzzy sliding-mode control into chattering-free IM drive," *IEEE Transactions on Industry Applications*, vol. 51, no. 1, pp. 692–701, 2014.

- [160] S. A. Khan, Y. Guo, and J. Zhu, "Model predictive observer based control for single-phase asymmetrical T-type AC/DC power converter," *IEEE Transactions on Industry Applications*, vol. 55, no. 2, pp. 2033–2044, 2018.
- [161] J. Yang, H. Wu, L. Hu, and S. Li, "Robust predictive speed regulation of converter-driven DC motors via a discrete-time reduced-order GPIO," *IEEE Transactions on Industrial Electronics*, vol. 66, no. 10, pp. 7893–7903, 2018.
- [162] S.-K. Kim and C. K. Ahn, "Self-tuning proportional-type performance recovery property output voltage-tracking algorithm for DC-DC boost converter," *IEEE Transactions on Industrial Electronics*, vol. 66, no. 4, pp. 3167–3175, 2018.
- [163] P. Lin, C. Zhang, X. Zhang, H. H. C. Iu, Y. Yang, and F. Blaabjerg, "Finite-time large signal stabilization for high power DC microgrids with exact offsetting of destabilizing effects," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 5, pp. 4014–4026, 2020.
- [164] Z. Wang, S. Li, and Q. Li, "Continuous nonsingular terminal sliding mode control of DC-DC boost converters subject to time-varying disturbances," *IEEE Transactions on Circuits and Systems II: Express Briefs*, vol. 67, no. 11, pp. 2552–2556, 2019.
- [165] A. K. Junejo, W. Xu, C. Mu, M. M. Ismail, and Y. Liu, "Adaptive speed control of pmsm drive system based a new sliding-mode reaching law," *IEEE Transactions on Power Electronics*, vol. 35, no. 11, pp. 12 110–12 121, 2020.
- [166] K. De Gusseme, D. M. Van de Sype, A. P. Van den Bossche, and J. A. Melkebeek, "Digitally controlled boost power-factor-correction converters operating in both continuous and discontinuous conduction mode," *IEEE transactions on industrial electronics*, vol. 52, no. 1, pp. 88–97, 2005.
- [167] H. Zhang, H. Li, Y. Zhou, and S. Zeng, "Cascaded proportional control with algebraic estimators for PFC AC/DC converters," *IEEE Transactions on Power Electronics*, vol. 34, no. 12, pp. 12 504–12 512, 2019.
- [168] Y. Wang, S. Mondal, C. Deng, K. Satpathi, Y. Xu, and S. Dasgupta, "Cyber-resilient cooperative control of bidirectional interlinking converters in networked AC/DC microgrids," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 10, pp. 9707–9718, 2020.
- [169] Z. Li, Z. Cheng, Y. Wang, and Q. Sui, "Distributed event-triggered fixed-time SMC-based AGC for power systems with heterogeneous frequency regulation units," *IEEE Transactions on Industrial Informatics*, vol. Early Access, DOI: 10.1109/TII.2024.3367003, 2024.
- [170] Y. Yin, J. Liu, W. Luo, L. Wu, S. Vazquez, J. I. Leon, and L. G. Franquelo, "Adaptive control for three-phase power converters with disturbance rejection performance," *IEEE Transactions on Systems, Man, and Cybernetics: Systems*, vol. 51, no. 2, pp. 674–685, 2018.

- [171] N. Tiwary, A. K. Panda, A. Narendra, R. K. Lenka *et al.*, “A robust voltage control of DAB converter with super-twisting sliding mode approach,” *IEEE Journal of Emerging and Selected Topics in Industrial Electronics*, vol. 4, no. 1, pp. 288–298, 2022.
- [172] L. Huang, H. Xin, and F. Dörfler, “ H_∞ -control of grid-connected converters: Design, objectives and decentralized stability certificates,” *IEEE Transactions on Smart Grid*, vol. 11, no. 5, pp. 3805–3816, 2020.
- [173] J. C. Doyle, B. A. Francis, and A. R. Tannenbaum, *Feedback control theory*. Courier Corporation, 2013.
- [174] L. B. G. Campanhol, S. A. O. da Silva, A. A. de Oliveira, and V. D. Bacon, “Dynamic performance improvement of a grid-tied PV system using a feed-forward control loop acting on the NPC inverter currents,” *IEEE Transactions on Industrial Electronics*, vol. 64, no. 3, pp. 2092–2101, 2016.
- [175] B. Singh, G. Bhuvaneswari *et al.*, “Position sensor-less synchronous reluctance generator based grid-tied wind energy conversion system with adaptive observer control,” *IEEE Transactions on Sustainable Energy*, vol. 11, no. 2, pp. 693–702, 2019.
- [176] C. Fu, C. Zhang, G. Zhang, J. Song, C. Zhang, and B. Duan, “Disturbance observer-based finite-time control for three-phase AC-DC converter,” *IEEE Transactions on Industrial Electronics*, vol. 69, no. 6, pp. 5637–5647, 2021.
- [177] E.-K. Kim, F. Mwasilu, H. H. Choi, and J.-W. Jung, “An observer-based optimal voltage control scheme for three-phase UPS systems,” *IEEE Transactions on Industrial Electronics*, vol. 62, no. 4, pp. 2073–2081, 2014.
- [178] X. Wu, B. Chen, M. Yang, T. Wu, S. Huang, and H. Cui, “Improved model-free predictive current control method for SPMSM with an overmodulation scheme,” *IEEE Transactions on Industrial Electronics*, vol. Early Access, DOI: 10.1109/TIE.2023.3317855, 2023.
- [179] X. Liu, L. Qiu, J. Rodríguez, W. Wu, J. Ma, Z. Peng, D. Wang, and Y. Fang, “Data-driven neural predictors-based robust MPC for power converters,” *IEEE Transactions on Power Electronics*, vol. 37, no. 10, pp. 11 650–11 661, 2022.
- [180] V. Blasko and V. Kaura, “A new mathematical model and control of a three-phase AC-DC voltage source converter,” *IEEE Transactions on Power Electronics*, vol. 12, no. 1, pp. 116–123, 1997.
- [181] C.-H. Liu and Y.-Y. Hsu, “Design of a self-tuning PI controller for a STATCOM using particle swarm optimization,” *IEEE Transactions on Industrial Electronics*, vol. 57, no. 2, pp. 702–715, 2009.