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**NANYANG
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SINGAPORE

**DECENTRALIZED CONTROL AND
STABILITY ANALYSIS FOR
ISOLATED MICROGRIDS**

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**SCHOOL OF ELECTRICAL AND ELECTRONIC
ENGINEERING**

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**DECENTRALIZED CONTROL AND
STABILITY ANALYSIS FOR
ISOLATED MICROGRIDS**

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Summary

A microgrid (MG), which combines distributed generators (DGs), energy storage devices and a cluster of loads, is a building block for the future smart grid. The increasing interest in integrating intermittent renewable energy sources (RESs) and the high penetration of power electronic converters into MGs pose two major challenges in the control of isolated MGs: one is the demand-supply balance issue under the uncertain renewable power generation and the other is the stability issue emerged from the tightly regulated power electronic converter loads.

Conventionally, centralized control methods are adopted to solve these two issues, which require a centralized controller and a communication network to coordinate the operation of local controllers. However, these methods encounter drawbacks like high communication and computation burdens as well as the single point of failure. Considering the distributed nature of the DGs in a MG and to increase system reliability and scalability, decentralized control strategies without central controllers and communications are explored in this thesis, to deal with the issues of power balance and system stability.

First, a decentralized dynamic power sharing strategy for hybrid energy storage systems in DC MGs is proposed. A hybrid energy storage system provides an effective approach to compensate the power unbalance caused by the intermittency of renewable DGs and uncertainty of loads. To fully exploit the advantages of different types of energy storage devices in the hybrid energy storage system, it is desirable that the access-oriented energy storage (e.g. supercapacitor) compensates fast power fluctuations and the capacity-oriented energy storage (e.g. battery) smoothes the low-frequency power at steady state. Unlike most existing methods with central controllers and communications, an extended droop control scheme with a virtual resistance droop controller and a virtual capacitance droop controller is proposed for the hybrid battery/supercapacitor(SC) system to achieve dynamic power sharing in a decentralized manner. In addition, considering the

bus voltage deviation issue resulted from the droop-based control approaches as well as the SC state-of-charge (SoC) recovery issue to ensure secure operation, a decentralized power management strategy is further proposed to achieve the autonomous dynamic power sharing, SC SoC recovery and bus voltage restoration at the same time. A design guideline is also developed to ensure the desired dynamics.

Next, the system stability issue is studied. Stable operation is a basic requirement for MGs. The high penetration of power electronic converter loads causes destabilizing effects in MGs, also known as the constant power load (CPL) issue. Considering the increasing complexity and the requirement of system extension and reconfiguration of the MGs, a modular approach is proposed for flexibly modeling and stability analysis. By this means, a complex system matrix can be easily integrated based on the system configuration and the overall system stability can be evaluated. The hybrid AC/DC more electric aircraft power system adopted by Boeing 787 is analyzed using the proposed technique as an example.

To further stabilize system with CPLs, a composite nonlinear controller is proposed for an on-board DC MG by integrating a nonlinear disturbance observer based feedforward compensation with the backstepping design algorithm. Unlike existing active damping methods that can only ensure stability in small signal sense, the proposed controller not only ensures global stability under large variation of the CPL, but also features fast dynamic response with accurate tracking over the wide operating range.

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Nomenclature

List of Abbreviations

AC	Alternating Current
ATRU	Auto Transformer Rectifier Unit
DC	Direct Current
BIC	Bidirectional Interlinking Converter
DG	Distributed Generation
ECS	Environmental Control System
EMA	Electromechanical Actuator
FC	Fuel Cell
GCU	Generator Control Unit
HESS	Hybrid Energy Storage System
HPF	High Pass Filter
VHPFD	Virtual High-Pass Filter based Droop
HVAC	High Voltage Alternating Current
HVDC	High Voltage Direct Current
LPF	Low Pass Filter
LVAC	Low Voltage Alternating Current
LVDC	Low Voltage Direct Current
MEA	More Electric Aircraft
MG	Microgrid
NDO	Nonlinear Disturbance Observer
RES	Renewable Energy Source
SC	Supercapacitor
SG	Synchronous Generator
SoC	State-of-Charge
TRU	Transformer Rectifier Unit
VCD	Virtual Capacitance Droop
VRD	Virtual Resistance Droop
VSI	Voltage Source Inverter

NOMENCLATURE

List of Symbols

C	Converter output capacitance
C_f	The HVDC bus filter capacitance
C_j	The virtual capacitance of the j th converter
C_{oB}	The output capacitor of the battery converter
C_{oSC}	The output capacitor of the SC converter
C_v	Virtual capacitance
d_1, d_2	The uncertain items
D_B	The duty ratio of the battery converter
D_{SC}	The duty ratio of the SC converter
E	Converter input voltage
f_s	Switching frequency
Q_B	Energy capacity of the battery
Q_c	Energy capacity of SC
R_j	Virtual resistance of the j th converter
R_{l1}	The line impedance between the battery and DC bus
R_{l2}	The line impedance between the SC and DC bus
R_v	Virtual resistance
SoC_{ini}	Initial SoC
i_{oB}	The output current of the battery converter
i_{oSC}	The output current of the SC converter
I_j	The output current of the j th converter
I_B	Battery current
I_{SC}	SC current
I_{jmax}	The current rating of the j th converter
i_{CPL}	The instantaneous value of the input current of the CPL
i_{load}	Load current
i_o	Equivalent load current of HESS
i_{RESs}	RESs current
k_i	The integral gain of the voltage compensation loop
k_1, k_2	Backstepping controller gains
L	Converter inductance
L_B	Inductor of the battery converters
L_f	The HVDC bus filter inductance
L_{SC}	Inductor of the SC converters
l_1, l_2	NDO gains
V_{bus}	Bus voltage
v_{CPL}	The instantaneous value of the input voltage of the CPL
V_j	The output voltage of the j th converter
V_B	Battery voltage
V_{Cref}	DC bus voltage reference
V_{SC}	SC voltage

NOMENCLATURE

V_{oB}	The output voltage of the battery converter
V_{oSC}	The output voltage of the SC converter
V_{nom}	The nominal bus voltage
V_{max}	Maximum bus voltage
V_{min}	Minimum bus voltage
ΔV_{VR}	Voltage compensation term
x_{ij}	State vector of module ij
y_{ij_u}	Interconnection vector from module ij to its up module
y_{ij_d}	Interconnection vector from module ij to its down module
y_{ijU_d}	Interconnection vector from the up module to the module ij
y_{ijD_u}	Interconnection vector from down module to the module ij
$Z_{oB}(s)$	The output impedance of the battery converter
$Z_{oSC}(s)$	The output impedance of the SC converter
$Z_o(s)$	The output impedance of the sources
τ	System time constant
ω_c	Corner frequency
ω_{iB}	The current control bandwidth of the battery converter
ω_{iSC}	The current control bandwidth of the SC converter
ω_n	Natural frequency of the proposed VHPFD strategy
ω_{vB}	The voltage control bandwidth of the battery converter
ω_{vSC}	The voltage control bandwidth of the SC converter

Chapter 1

Introduction

1.1 Background

1.1.1 Smart grid and microgrid technology

Electricity demand in the world increases continuously in recently years. Fig. 1.1 shows the world net electricity generation by different energy sources from 2012 to 2040 [1]. As can be observed, electricity generation will increase by an estimated 69% from 2012 to 2040 with an average annual increase of 1.9% and renewable energy sources (RESs) are the fastest-growing sources with an average annual increase of 2.9%. Due to environmental concerns and sustainable requirements, there is great interest in the utilization of the RESs (such as solar PV and wind) around the world [2]. The Sustainable Energy for All Initiative (SE4ALL) launched by the United Nations in 2012 has called for a target of doubling the share of renewable energy in the global consumption from a baseline value of 18% in 2010 to 36% in 2030 [3]. In the Manila-Paris Declaration in Climate Vulnerable Forum 2015, a coalition of 30 nations called for 100% renewable energy power supply by the year 2050 [4]. In Europe, the milestone is to increase the share of RESs to over 50% and to reduce the greenhouse gas emission by over 80% by 2050 [5]. In US, it is estimated that renewable electricity generation is sufficient to supply 80% of total electricity generation in 2050 [6].

To cope with the challenges of growing power demand and the high penetration of RESs, as well as the high requirements for system reliability and performance, electrical power system is going through the evolution of "Smart Grid" [7]. A smart grid is an electric network that can intelligently integrate all the generators and consumers efficiently by utilizing the information, communication, advanced

Figure 1.1: The world net electricity generation from 2012 to 2040 (trillion kilowatthours).

control and computational intelligence technologies, to provide clean, secure, efficient, economic, resilient, sustainable and reliable electricity supplies [8]. As the building blocks of the smart grids, microgrids (MGs) provide a promising technology for the integration of RESs into the grid with improved system efficiency and reliability as well as the possibility of islanding operation at end-user sites [9]. An MG integrates a group of distributed generators (DGs) such as RESs, fuel cells and microturbines, energy storage units such as flywheels, supercapacitors and batteries, and local loads to form a single controllable entity with respect to the power grid [10].

There are various MG related research, development and demonstrations around the world [9]. In Europe, two well-known projects are the *Microgrid: Large Scale Integration of Micro-Generation to Low Voltage Grids* (1998-2002) funded at €4.5million and *More Microgrids: Advanced Architectures and Control Concepts for More Microgrids* (2002-2006) funded at €8.5million [11]. The Microgrid project developed laboratory MGs of various functionalities. As a follow-up project of the Microgrid, the More Microgrids project realized MGs in real-world pilot sites and provided practical conclusions on operation and management of MGs based on field trials. In the USA, two major MG programs are Smart Power

Infrastructure Demonstration for Energy, Reliability and Security (SPIDERS) and Renewable and Distributed Systems Integration (RDSI); the former focused on the energy security and reliability issues, and the latter was to investigate the MG value during peak load in the utility grid [9]. In China, almost every province had its respective MG related projects and applications, and many MGs had been installed in various sites, including remote areas as well as industrial, commercial or residential sites [12]. In Singapore, the energy market authority launched an MG test-bed at Pulau Ubin island in 2013, aiming to assess the reliability of an MG using the intermittent RESs such as PV so as to bring more clean and cost-competitive electricity with improved reliability and scalability to end-users [13].

MGs can operate in both grid-connected or isolated mode by connecting or disconnecting from the grid [10]. In grid-connected mode, the frequency and voltage of the MG are built by the main grid. In isolated operation, without the support from the main grid, frequency and voltage should be maintained by DGs and the dynamics are strongly dependent on the coordination of converter interfaces of DGs. Due to the small inertia of power electronics interfaced DGs, it will be more challenging to ensure control accuracy and system stability for isolated MGs. Isolated MGs have been used in many applications, like remote sites (e.g. remote communities or remote industrial sites) or electric transportation systems (avionics, automotive, marine, etc). This thesis focuses on isolated MGs.

1.1.2 Microgrid structures and components

Based on the types of voltage in the point of common coupling, MGs can be classified as AC MGs, DC MGs and hybrid AC/DC MGs [14,15], as demonstrated in Fig. 1.2-1.4. In the AC MG, sources and loads are connected to the common AC bus. For the DC MG, sources and loads are integrated to the common DC bus. In the hybrid AC/DC MG, AC-compatible DGs and DC-compatible DGs are allocated at the AC subgrid and DC subgrid, respectively; the AC subgrid and DC subgrid are interfaced by a bidirectional interlinking converter.

Many research works have been focused on AC MGs development as conventional power systems have predominantly adopted AC due to its advantages in power transmission, distribution and transformation [16–19]. On the other hand, as many DGs like solar PV and fuel cell as well as most energy storage devices, are DC sources by nature and there are an increasing number of DC loads integrated

Figure 1.2: A typical AC MG

Figure 1.3: A typical DC MG

to the grid, including light-emitting diode (LED), communication and computation devices, motor drive systems, etc., DC MGs provide an efficient integration

Figure 1.4: A typical hybrid AC/DC MG

of sources and loads with the reduction of power conversion stages. Moreover, they do not suffer the problems encountered by their AC counterparts, including synchronization, reactive power flow, harmonics, etc. Hence DC MGs attract a great amount of interests in recent years [20–23]. The hybrid AC/DC MGs are proposed recently by combining AC subgrids and DC subgrids [14, 15, 24–26] and they inherit advantages of both AC MGs and DC MGs. Considering the increasing attention in DC MGs and hybrid AC DC MGs for the promising future distribution systems, this thesis focuses on DC MGs and the DC subgrids in hybrid AC/DC MGs.

According to the definition, an MG consists of a cluster of DGs, energy storage devices and local loads and they are briefly introduced as follows:

1) DGs

DGs refer to the power generations near the user sites, including the RESs such as PV and wind turbines, and microsources such as diesel generators [27]. They are experiencing rapid growth in recent years due to their advantages in reducing pollution, relieving energy shortage, improving system efficiency as well as smoothing the stress on power system transmission and distribution. Based on the controllability, DGs can be categorized as dispatchable units and nondispatchable

Table 1.1: Some characteristics of different energy storages

Type	Energy density (Wh/Kg)	Power density (W/Kg)	Life(cycles or years)	Response time(s)
Li-ion battery	100-200	360	500-10000	< 1/4cycles
CAES	10-30	-	20-40 years	360
Pumped hydro	0.3	-	20-60 years	10
SC	2-5	800-2000	> 100000	0.01
Flywheel	5-30	> 2000	> 20000	0.1
SMES	10-75	-	> 100000	0.01

units [28].

Dispatchable units can be fully controlled, such as diesel generators, microturbines and fuel cells.

RESs including solar PV and wind turbines are typical nondispatchable units. The RESs are commonly dependent on the weather factors and have intermittent nature; as a result, their output power is not controllable and they normally operate at the maximum power point tracking (MPPT) mode for the maximum power output to optimize the benefits. The volatile generation of the RESs challenges the power balance in MGs and normally energy storage devices are required for their integration.

2) energy storage and hybrid energy storage system

The intermittency of the RESs necessitates the presence of energy storage to compensate the power imbalance at the MGs. Different types of energy storage devices have different features in terms of power density, energy density, response rate, life cycle, cost, etc [29], as shown in Table 1.1. Generally, energy storage devices can be divided into two categories: capacity-oriented energy storage devices and access-oriented energy storage devices [30]. Capacity-oriented energy storage devices have high energy density but low power density and slow response time; they are usually used for long-term energy balancing to compensate the low-frequency fluctuation of RESs. Batteries, compressed air energy storage (CAES), pumped hydroelectric systems are considered as capacity-oriented energy storage devices. Access-oriented energy storage devices have high power density, fast dynamic response but low energy density. Therefore, they are responsible for high-frequency components of the imbalanced power in MGs. Typical access-oriented energy storage devices include supercapacitors (SCs), flywheels and superconducting magnetic energy storage (SMES), etc.

Figure 1.5: The schematic diagrams of different HESS topologies.

As can be observed, capacity-oriented energy storage devices and access-oriented energy storage devices have complementary characteristics and thus hybridization of these two types can provide fast dynamic performance, high power supply and large energy supply at the same time. The most common representative of a hybrid energy storage system (HESS) is the hybrid battery/SC system, which can be found in many applications [20, 31–36] and will be studied in this thesis.

The topologies of HESS can be divided into three categories, i.e. passive [37], semi-active [32, 38] and active [31, 33, 39, 40], as presented in Fig. 1.5 [41]. In passive topology, the battery and SC are connected directly to the DC bus and thus their power flow is uncontrollable. Two types of semi-active topologies are illustrated in Fig. 1.5b and 1.5c. For the semi-active (1) topology, the battery and SC are connected in parallel and a converter is added at the common point to actively control their output power. However, as SC output energy is related to its terminal voltage variation, which is limited by the battery terminal voltage variation in this case, SC capacity cannot be fully used. Semi-active (2) topology, in which only the SC is connected to a DC/DC converter, may be more cost effective by reducing the cost of one DC/DC converter. However, the battery terminal voltage needs to be matched with the nominal bus voltage and the battery current cannot be managed actively. Furthermore, for the fully active topology,

the battery voltage is boosted to a higher level and a smaller sized battery can be selected to reduce its cost, thus it is difficult to say the semi-active topology is more cost effective. For the fully active topology, both the battery and SC are connected to the common bus through the DC/DC converters. Then it is not necessary to match the terminal voltage of both energy storage devices with the bus voltage, giving flexible choices on the size of energy storage devices. Moreover, output power of energy storage devices can be controlled effectively to make full use of the power and energy capacities of the energy storage devices. Therefore, the active topology is most common for the HESS and will be selected in this thesis.

3) Load

Power electronic converters are widely applied in MGs as efficient interfaces for the integration of various sources and loads [42]. Power electronic converter loads, when tightly controlled, behave as constant power loads (CPLs) [43]. Typical examples of CPLs are DC/DC converter feeding resistive loads and DC/AC inverter driven electric motors [44]. CPLs have negative impedance characteristics, which may cause destabilizing effects. Instability problems arise from the interactions when CPLs are interconnected with other subsystems, even though each subsystem is designed to meet the stability requirements in its standalone mode [45, 46]. This makes stability issue more significant for MGs.

1.1.3 Microgrid control techniques

From the viewpoint of communication, control techniques of MGs can be divided into three categories: centralized, distributed and decentralized, as shown in Fig.1.6 [22, 47, 48].

In centralized control, data of each local controller will be transferred to a central controller and the central controller will process the information and give command to each local controller through the communication network, as is shown in Fig. 1.6a. The central controller coordinates the operation of local controllers and it provides best foundation to achieve many advanced control functionalities and optimize system performance. However, one obvious disadvantage is the single point of failure issue. In addition, centralized control requires extensive communication between the central controller and local controllers, and has high computation burden in the central controller, especially when a MG is at a large geographic area. Furthermore, system expansion or reconfiguration also requires

Figure 1.6: Control techniques of MGs.

the redesign of the central controller with increased complexity.

Distributed control, as shown in Fig. 1.6b, eliminates the central controller while allows the communication among local controllers. Consensus algorithms are proposed to make system performance comparable to that under the central-

ized control. This approach is immune to the single point of failure issue and system remains functionable even with the failure of some communication links. The elimination of the central controller makes it flexible for the situations of system extension and reconfiguration. However, as the communication link still exists, the physical implementation becomes more complex with the increasing number of DGs and the system performance deteriorates a lot under communication delays. Another disadvantage is its complexity in performance analysis, including the convergence rate and stability margins.

In decentralized control, each unit is controlled by its local controller and digital communication links do not exist among the units, which is shown in Fig. 1.6c. Considering the drawbacks in centralized control and distributed control, it is expected that more functions can be achieved in a decentralized manner; no communication is required and there is overall improvement in terms of economy, scalability and reliability.

1.2 Motivation

MGs, especially operating at isolated mode, encounter two major challenges [22]:

Power balance: RESs such as PV and wind, are highly affected by the weather factors, leading to the intermittency of the power generation. On the other side, there is uncertainty in the load profile considering the large number of the users. It is critical to ensure demand-supply balance and MGs should be able to accommodate the uncertain variations to keep frequency/voltage deviation within the acceptable ranges.

System stability: Unlike the conventional synchronous generators that bring large inertia for the bulk power systems, MGs show a low-inertia characteristic, especially when dominated by the power converter interfaced DGs. Furthermore, the high penetration of power electronic converter loads might decrease system damping. The power electronic converter loads, when tightly regulated, behave as CPLs with negative impedance characteristic and they have destabilizing effects.

Conventionally, centralized control methods are adopted to solve the above problems. Some distributed control approaches are also proposed recently. However, as discussed in Section 1.1.3, they have their respective drawbacks due to the communication links. To increase system reliability and scalability, decentralized control strategies without communications are explored in this thesis, to deal with the issues of power unbalance and system stability.

1.3 Major Contribution of the Thesis

The major contributions can be summarized as

1) *A decentralized dynamic power sharing strategy for hybrid energy storage systems*

To compensate the imbalanced power caused by the intermittency of RESs, a hybrid energy storage system with a battery and a supercapacitor (SC) is integrated into a DC MG. To fully exploit the advantages of HESS, it is desired for the battery to balance the steady-state power mismatch and for the SC to compensate transient power. To achieve this kind of dynamic power sharing, an extended droop control scheme with a virtual resistance droop (VRD) controller and a virtual capacitance droop (VCD) controller is proposed. Output voltage characteristics of the battery and SC are regulated by a VRD controller and a VCD controller respectively. The proposed droop controllers inherently act as a low-pass filter for the battery and a high-pass filter for SC to load response so that SC compensates high frequency fluctuations during transient and the battery provides load power at steady state. Unlike most existing methods with communication links or central controllers, the proposed strategy is a fully decentralized one. For practical implementation of this control strategy, a detailed procedure for parameter design is proposed. With the developed system dynamic model and derived transfer functions, parameters can be properly designed to ensure stable operation, bus voltage regulation and dynamic current sharing. System expansion is also demonstrated to show the scalability of the proposed method. Simulations and experiments are conducted to verify the effectiveness of the proposed strategy.

2) *A decentralized power management strategy for hybrid energy storage systems with autonomous bus voltage restoration and state-of-charge recovery*

As the previous proposed extended droop control strategy does not consider the issues of bus voltage deviation and SC state-of-charge (SoC) violation, a decentralized power management strategy is further proposed, to achieve the autonomous power split, bus voltage regulation and SC SoC recovery at the same time. A virtual high-pass filter based droop (VHPFD) controller is proposed to regulate the battery converter and a VCD controller is implemented for SC converter. The cooperation of VHPFD and VCD first achieves autonomous power split that high frequency fluctuation is buffered by SC and low frequency power is supplied by the battery. Meanwhile, the bus voltage deviation induced by the droop based power sharing is eliminated automatically at steady state. The re-

sulted bus voltage restoration simultaneously enforces SC SoC back to its nominal value and thus ensures continuous operation of SC as a power buffer without the violation of its SoC boundary. A design guideline is developed to ensure the expected system dynamics. Impact of line impedance is investigated and the scalability of the proposed approach is illustrated. The effectiveness of the proposed method and analytical results are validated by simulations and experiments.

Though the HESS studied in Chapter 2 and Chapter 3 is integrated in a DC MG, it can also be connected to a DC subgrid in a hybrid AC/DC MG or an AC MG through an inverter, and the proposed approach is flexible to be implemented.

3) A modular approach for stability analysis of complex MGs

Along with the precise control and coordination, stable operation should always be ensured for MGs, especially with the high penetration of CPLs. Considering the increasing complexity and the requirement of system extension and reconfiguration of MGs, a modular approach is proposed for effective system modeling and stability analysis. With this approach, system matrix can be easily integrated based on the system configuration and system stability can be evaluated by the eigenvalue method. The proposed module-based approach is applied in the modeling and stability analysis of the hybrid AC/DC more electric aircraft (MEA) power system, as an application example. The complex hybrid AC/DC power system is presented with models of each component and the whole system is integrated with the proposed approach. Then stability analysis is performed for the complex MEA power system with the verification of simulation results.

4) A novel composite nonlinear controller for stabilization of constant power loads in MGs.

To further stabilize system with CPLs, a composite nonlinear controller is proposed. First, the model is transformed into the Brunovsky's canonical form using the exact feedback linearization technique, to handle the nonlinearity introduced by the CPL. Second, the nonlinear disturbance observer technique is adopted to estimate the load power variation within a fast dynamic response, serving as a feedforward compensation to increase the accuracy of output voltage regulation. Then a nonlinear controller is developed by following the step-by-step backstepping algorithm with strictly guaranteed large signal stability. Unlike the conventional active damping methods that improve system stability in a small signal sense, the proposed controller not only ensures global stability under large variation of the CPL, but also features fast dynamic response with accurate tracking over the wide operating range. Both simulations and experiments are conducted

to verify the proposed strategy.

1.4 Organization of the Thesis

This thesis is organized as follows:

Chapter 1 introduces the background, motivation and contributions of this work.

Chapter 2 and Chapter 3 deal with the power unbalance issue. In Chapter 2, a decentralized dynamic power sharing strategy is proposed for a hybrid battery/SC system to compensate the unbalanced power. The proposed strategy is an extended droop control scheme with a virtual resistance droop controller and a virtual capacitance droop controller. The design procedure is presented with detailed frequency domain analysis. System stability and scalability are evaluated, and the SoC regulation of the energy storage devices are discussed. Finally, simulations and experiments are conducted to verify the effectiveness of the proposed approach and the analytical results.

Chapter 3 is an improvement for the work in Chapter 2 by considering the bus voltage restoration and SC SoC recovery. A virtual high-pass filter based droop controller is proposed to regulate the battery converter and a virtual capacitance droop controller is implemented for the SC converter. The proposed strategy can achieve the dynamic power sharing, bus voltage restoration and SC SoC recovery in a fully decentralized manner without any communication links. Next, the practical implementation and parameter design procedure are presented with a detailed frequency domain analysis. Some considerations including line impedance impacts and system expansion are discussed. Finally, simulation and experimental results are demonstrated to verify the effectiveness of the proposed strategy.

Chapter 4 and Chapter 5 focus on stability issues caused by the constant power loads in MGs. In Chapter 4, first, a modular approach is proposed for effective system modeling and stability analysis. In the second part of Chapter 4, the proposed approach is implemented in the hybrid AC/DC more electric aircraft power system as an application example. Models of subsystems are presented and the whole system model is integrated with the proposed approach. Then system stability is performed with the simulation verifications.

In Chapter 5, a composite nonlinear controller is proposed for stabilization of constant power loads. The system model is presented first. Then, a composite nonlinear control strategy based on backstepping and nonlinear disturbance ob-

server techniques is proposed. Simulation verification and experimental results are demonstrated to verify the proposed control algorithm.

Chapter 6 gives the conclusion and future works.

Chapter 2

A Decentralized Dynamic Power Sharing Strategy for Hybrid Energy Storage Systems

As introduced in Chapter 1, the intermittency of the RESs and the uncertainty of the load cause the power unbalance issue in isolated MG. This chapter deals with this issue and proposes a decentralized dynamic power sharing strategy for hybrid energy storage systems to compensate the imbalanced power. While the hybrid energy storage system is integrated into an isolated DC MG in this thesis, it can also be integrated into the DC subgrid of a hybrid AC/DC MG or an AC MG through an inverter, and the proposed approaches are flexible to be implemented as well.

2.1 Introduction

The RESs, such as PV and wind turbine, are dependent on the weather factors and their generation has intermittent nature. Energy storage is widely used in MGs for the integration of RESs to smooth the fluctuation of RESs to increase power quality, stability and reliability [20, 39]. Different types of energy storage devices have different energy density, power density, response time, life cycle, etc [29]. For example, the SC has high power density, fast dynamic response

speed but low energy density. On the other hand, the battery has comparatively high energy density but low power density and slow dynamic response. The hybrid energy storage system (HESS) inherits the complementary characteristics of energy storage devices and achieves fast dynamic performance, high power supply and large energy supply at the same time [32, 36].

How to effectively split power for different energy storage devices is a major challenge in HESS. It is desirable that the SC compensates fast fluctuations and the battery provides smooth power at steady state [32, 36]. Many power management strategies have been proposed for HESS in literatures, ranging from rule based or filter-based control approaches [33, 39], to optimization based strategies [32, 40] and model predictive control methods [49, 50]. In the filter-based approaches, the system net power is passed through high pass filter (HPF) or low pass filter (LPF) in a central controller to generate power references for different energy storage devices [35, 36, 39, 51]. In [32], a multi-objective optimization is formulated and solved by dynamic programming and a neural network to optimize power split in the battery/ultracapacitor system. In [33], the wavelet transform algorithm is employed to decouple high frequency and low frequency components of power demand for hybrid SC/battery system. A model predictive control approach is proposed in [50] to allocate fast and slow current changes to the SC and battery respectively. For all these methods, system net power should be accurately measured and a central controller is required to generate power reference for the local controller of each ES by communication links. Therefore, these methods will suffer communication delays and single point of failure issues. The expansion and reconfiguration of the MG will also lead to the redesign of the control architecture with increased complexity and computation burdens [22, 27, 47]. Considering the distributed nature of DGs in isolated MGs and to increase the system reliability and scalability, decentralized control strategies without central controllers and communications should be developed [47]. In decentralized control, each local unit makes independent control decision based on the its local information [22]. In [52], a decentralized controller is proposed for the fuel cell (FC) and ultracapacitor (UC) system, enabling the separation control of FC and UC without communication. However, the FC controller still needs to measure load current, which makes it not fully decentralized, and the control structure for each controller is complicated for practical applications. In [53], a decentralized frequency-coordinating virtual impedance method is proposed for dynamic power sharing. In this method, a proportional voltage controller is combined with an

LPF and an HPF to generate current references for the battery and SC, respectively. As a proportional controller generates steady state error, the combination of the proportional controller and HPF for the SC controller will lead to an uncontrolled leakage current at steady state, which is not desirable for the operation of the SC.

The droop control strategy is the mostly used decentralized method in MG [47, 54]. But existing droop based control methods only achieve proportional power sharing at steady state [22, 55, 56]. Considering different dynamics of ESs, power sharing based on different frequency ranges is required.

The voltage-current characteristics of droop control scheme can be seen as a virtual resistor at the output of the converter. To decouple the power for energy storage devices with different dynamics, this chapter proposes an extended droop control (EDC) scheme by extending the concept of virtual resistance droop (VRD) to virtual capacitance droop (VCD) control. The proposed VCD can regulate SC so that it only responds to fast fluctuation and will not provide power at steady state. The conventional VRD remains for the battery for consistent power supply. Each energy storage device operates at the voltage regulation mode to maintain the bus voltage and only local measurements are required for its converter controller without the communication with other energy storage devices. The energy storage devices can be located at different locations and system expansion can be flexibly achieved without the modification of existing control architectures.

Moreover, most of the power management strategies in literatures only focus on control strategies whereas the parameter design of the controller is often neglected. To ensure stable operation and improve system performance, a parameter design procedure is also proposed in this chapter based on the detailed frequency domain analysis.

This chapter is organized as follows: First, the extended droop control scheme with VRD and VCD is proposed. Next, the design procedure is presented with detailed frequency domain analysis. System stability and scalability are also evaluated. Finally, simulations and experiments are conducted to verify the effectiveness of the proposed approach and the analytical results.

Figure 2.1: An isolated DC microgrid integrating with HESS.

2.2 The Proposed Dynamic Power Sharing Strategy

The schematic diagram of an isolated DC MG is depicted in Fig. 2.1, which consists of RESs, a battery, a SC and a lumped DC load. RESs (PV and wind) are connected to DC bus through their respective power converters and work at the maximum power point tracking mode. The DC load is connected to the bus. The hybrid battery/SC system will compensate power mismatch between RESs and load to maintain the bus voltage. The battery and SC are connected to the DC bus through their respective bidirectional DC/DC converters.

In an autonomous DC MG, droop control is commonly implemented for load sharing among different energy sources. The voltage-current ($V - I$) relationship of droop control is expressed as

$$V_j = V_{nom} - R_j \cdot I_j \quad (2.1)$$

where V_j and I_j are the output voltage and current of the converter for the j th source, V_{nom} is the nominal bus voltage, and R_j is the corresponding droop coefficient, which can be seen as a virtual resistor.

The virtual resistor R_j is calculated by

$$R_j = \frac{V_{\max} - V_{\min}}{I_{j\max}} \quad (2.2)$$

where $V_{\max}$, $V_{\min}$ are the maximum and minimum allowable bus voltage and

Figure 2.2: Equivalent circuit for HESS under the EDC.

I_{jmax} is the maximum current of the j th source converter.

For the droop control, power sharing among energy sources is achieved at steady state based on their power rating. However, for the hybrid system like the battery/SC system, power sharing at different time scales is required. To solve this problem, an extended droop control (EDC) strategy is proposed by extending the conventional virtual resistance droop control (VRD) to virtual capacitance droop control (VCD).

From the view of frequency response, a capacitor acts as short circuit at high frequency and open circuit at low frequency. Therefore, a VCD controller can be applied for the SC so that it can compensate high frequency fluctuation during transient and will not provide power at steady state. The conventional droop controller remains for the battery. This is the main idea of the proposed EDC strategy.

By replacing the virtual resistor in (2.1) with a virtual capacitor, the $V - I$ relationship of the proposed virtual capacitance droop controller can be obtained as

$$V_j = V_{nom} - \frac{1}{C_j s} \cdot I_j \quad (2.3)$$

where C_j is the droop coefficient of the j th source with fast dynamics.

Fig. 2.2 shows the equivalent circuit for MG in Fig.2.1 with the implementation of EDC. R_v is the virtual resistor implemented in battery control loop and C_v is the virtual capacitor implemented in the SC control loop. R_{l1} and R_{l2} are the line impedances of the battery and SC converters.

As the droop coefficient is normally much larger than line resistance [57], the line impedance can be neglected. The output $V - I$ relationships of the battery

Figure 2.3: Step responses of i_{oB} and i_{oSC} when $R_v C_v = 1..$

and SC are expressed as:

$$\begin{cases} V_{oB} = V_{nom} - R_v \cdot i_{oB} \\ V_{oSC} = V_{nom} - \frac{1}{sC_v} \cdot i_{oSC} \end{cases} \quad (2.4)$$

where V_{oB} , V_{oSC} and i_{oB} , i_{oSC} are the output voltage and current of the battery converter and SC converter, respectively.

As the role of the HESS is to compensate the power mismatch between RESs and load, the difference of load current i_{load} and RESs current i_{RESs} can be modeled as the equivalent load current of HESS i_o :

$$\begin{cases} i_o = i_{load} - i_{RESs} \\ i_{oB} + i_{oSC} = i_o \end{cases} \quad (2.5)$$

where i_o can be either positive or negative; when i_o is positive, the HESS supplies the equivalent load current; when i_o is negative, the HESS is charged by the equivalent load current.

The current sharing between the battery and SC can be obtained as

$$\begin{cases} i_{oB} = G_B(s) \cdot i_o = \frac{1}{sR_v C_v + 1} \cdot i_o \\ i_{oSC} = G_{SC}(s) \cdot i_o = \frac{sR_v C_v}{sR_v C_v + 1} \cdot i_o \end{cases} \quad (2.6)$$

In (2.6), $G_B(s)$ acts like an LPF while $G_{SC}(s)$ acts like an HPF with the corner frequency ω_c given in (2.7). Thus, the low frequency and high frequency parts of the load current are assigned to the battery and SC respectively. Fig. 2.3 shows the dynamic response of the battery current (i_{oB}) and SC current (i_{oSC}) with a step change in the load current i_o . It can be observed that the

SC current responds immediately and gradually reduces to zero. Meanwhile, the battery current increases gradually and finally reaches i_o . The system dynamic is determined by the product of virtual resistance R_v and virtual capacitance C_v as expressed in (2.8).

$$\omega_c = \frac{1}{R_v C_v} \quad (2.7)$$

$$\tau = R_v C_v \quad (2.8)$$

Given the required corner frequency ω_c or system time constant τ , virtual capacitor C_v can be calculated by (2.2) and (2.7) or (2.2) and (2.8). As the response time of the SC is usually in milliseconds and the response time of the battery is much longer [34], SC is assumed to be able to compensate all the high frequency power immediately. To ensure that the battery can handle the power after the equivalent LPF, the time constant τ should be greater than or equal to the response time of the battery.

If the line impedances R_{l1} and R_{l2} are taken into consideration, the current sharing between the battery and SC is

$$\begin{cases} i_{oB} = G'_B(s) \cdot i_o = \frac{sR_{l2}C_v+1}{s(R_1+R_{l1}+R_{l2})C_v+1} \cdot i_o \\ i_{oSC} = G'_{SC}(s) \cdot i_o = \frac{s(R_v+R_{l1})C_2}{s(R_v+R_{l1}+R_{l2})C_v+1} \cdot i_o \end{cases} \quad (2.9)$$

It can be observed that SC still compensates the high frequency power but the battery has to take some portion of high frequency power due to the existence of line resistance R_{l2} . As the droop coefficient is normally much larger than the line resistance [57], the condition in (2.10) is usually satisfied and the high frequency power taken by the battery is negligible. To further avoid the undesirable result, it is recommended that the SC system should be connected closer to the common bus.

$$R_v + R_{l1} \gg R_{l2} \quad (2.10)$$

When the condition (2.10) is satisfied, $G'_B(s)$ and $G'_{SC}(s)$ can be seen as LPF and HPF like that in (2.6) and thus the battery and SC are allocated with the low frequency and high frequency load current respectively as expected. The time constant of the new equivalent LPF/HPF is expressed as (2.11) and it is larger than that in (2.8), which leads to a slower system dynamic response. This impact can be eliminated by tuning the virtual capacitor C_v in (2.11) to ensure the desired system dynamic response. Therefore, the proposed method can operate effectively

Figure 2.4: DC microgrid with the implementation of the proposed control strategy.

when the line resistance is taken into consideration.

$$\tau' = (R_v + R_{l1} + R_{l2})C_v \quad (2.11)$$

It is worth noting that the proposed method is not limited to the hybrid battery/SC system and can be applied in hybrid systems with sources of different dynamics, such as fuel cell/SC system or fuel cell/battery system. For general applications, the proposed VCD controller is employed for energy sources with fast dynamics and the VRD controller remains for the rest of energy sources.

2.3 The Parameter Design Procedure

The system equivalent circuit with the implementation of controller is depicted in Fig. 2.4. As the focus of this chapter is the control strategy, the battery and SC are modeled as ideal voltage sources. The system dynamic model can be derived as

$$\begin{cases} L_B \cdot \frac{dI_B}{dt} = V_B - d_B \cdot V_{oB} \\ L_{SC} \cdot \frac{dI_{SC}}{dt} = V_{SC} - d_{SC} \cdot V_{oSC} \\ C \cdot \frac{dV_{bus}}{dt} = d_B \cdot I_{oB} + d_{SC} \cdot I_{oSC} \end{cases} \quad (2.12)$$

where V_B , V_{SC} and I_B , I_{SC} are the output voltage and current of the battery and SC, respectively; V_{oB} , V_{oSC} and I_{oB} , I_{oSC} are the output voltage and current of the battery and SC converters, respectively. V_{bus} is the bus voltage. As discussed

Figure 2.5: Ideal frequency response with the implementation of EDC.

previously that the impact of the transmission line can be neglected, the transmission line is ignored here, and V_{oB} and V_{oSC} are assumed to be equal to V_{bus} . The duty ratios of g_2 and g_4 are defined as d_B and d_{SC} and their steady state values are defined as D_B and D_{SC} .

In the ideal case, each converter system is seen as an ideal voltage source, as shown in Fig. 2.2. The output impedances of the battery converter, SC converter and the parallel system $Z_{oB}^*(s)$, $Z_{oSC}^*(s)$, $Z_o^*(s)$ under the proposed EDC method can be expressed as (2.13) in the frequency domain and their corresponding frequency responses are shown in Fig.2.5.

$$\begin{cases} Z_{oB}^*(s) = R_v \\ Z_{oSC}^*(s) = (C_v s)^{-1} \\ Z_o^*(s) = \frac{R_v}{R_v C_v s + 1} \end{cases} \quad (2.13)$$

It can be observed that $Z_{oB}^*(s)$ and $Z_{oSC}^*(s)$ intersect at corner frequency ω_c and the system output impedance is dominated by the virtual resistor before ω_c and virtual capacitor after ω_c . This is coherent with (2.6) and (2.7).

However, a converter system is not an ideal voltage source in real case, and its output impedance will be affected by the system parameters and inner control

Figure 2.6: Control architecture for a single converter.

loops. For the practical implementation of the control strategy, proper parameters should be designed to make the output impedances as close to the ideal case as possible to achieve the expected power sharing strategy. Therefore, a detailed controller design procedure is proposed to meet the following requirements:

- 1) Stable operation and dynamic performance of each converter system;
- 2) Dynamic current sharing between the battery and SC;
- 3) Stable operation of the whole system.

2.3.1 Design of the converter inner control loop

Both the battery and SC subsystems are controlled as voltage sources to regulate the DC bus voltage. The most-adopted double-loop control is employed for individual converter control, as shown in Fig. 2.6. The subscript x denotes the x th energy source. For the HESS, $x=B$ denotes the battery and $x=SC$ denotes SC. As the current loop is designed to be fast enough, the open-loop transfer function of the current control, which is shown in the dashed box, can be seen as a gain of 1. Then the voltage control loop can be simplified as (2.14) based on Fig. 2.6.

$$[D_x G_{vx}(s) (V_{nom} - V_{ox}) - I_{ox}] \frac{1}{C_{ox}s} = V_{ox} \quad (2.14)$$

The open-loop and closed-loop transfer function of voltage control can be obtained as (2.15) and (2.16) respectively.

$$T_{vx.OL}(s) = G_{vx}(s) \frac{D_x}{C_{ox}s} \quad (2.15)$$

$$T_{vx.CL}(s) = \frac{G_{vx}(s) D_x}{G_{vx}(s) D_x + C_{ox}s} \quad (2.16)$$

where the voltage controller G_{vx} is expressed as

$$G_{vx}(s) = k_{vpx} + \frac{k_{vix}}{s} \quad (2.17)$$

Assume that ω_{vx} is the control bandwidth of the voltage control loop. To ensure enough phase margin, the corner frequency of the PI compensator k_{vp}/k_{vp} should be less than ω_{vx} , which can be expressed as

$$\frac{k_{vix}}{k_{vp}} = \eta\omega_{vx} \quad (2.18)$$

where η is a small value between 1/10-1/5.

To design the voltage controller, the gain of the open-loop voltage transfer function T_{vx_OL} at ω_{vx} should be zero dB. It can be expressed as

$$T_{vx_OL}(j\omega_{vx}) = \left| \frac{k_{vp}(j\omega_{vx} + \eta\omega_{vx})}{j\omega_{vx}} \frac{D_x}{C_{ox}j\omega_{vx}} \right| = 1 \quad (2.19)$$

As η is much less than 1, it can be neglected in (2.19) and the voltage controller parameters can be calculated as

$$\begin{cases} k_{vp} = \frac{\omega_{vx}C_{ox}}{D_x} \\ k_{vix} = k_{vp}\eta\omega_{vx} = \frac{\eta\omega_{vx}^2C_{ox}}{D_x} \end{cases} \quad (2.20)$$

Substituting (2.20) into (2.16), the closed-loop transfer function of the voltage controller turns out to be

$$T_{vx_CL}(s) = \frac{G_{vx}(s)D_x}{G_{vx}(s)D_x + C_{ox}s} = \frac{\omega_{vx}s + \eta\omega_{vx}^2}{s^2 + \omega_{vx}s + \eta\omega_{vx}^2} \quad (2.21)$$

where the system natural frequency ω_n and damping ratio ζ can be obtained as

$$\begin{cases} \omega_n = \sqrt{\eta}\omega_{vx} \\ \zeta = \frac{\omega_{vx}}{2\omega_n} = \frac{1}{2\sqrt{\eta}} \end{cases} \quad (2.22)$$

To have a larger damping, η can be selected as 1/10. The inner current loop can be designed according to the process similar to (2.15)-(2.20). The parameters of the current controller can be calculated as

$$\begin{cases} k_{ip} = \frac{\omega_{ix}L_x}{V_{nom}} \\ k_{ii} = k_{ip}\eta\omega_{ix} = \frac{\eta\omega_{ix}^2L_x}{V_{nom}} \end{cases} \quad (2.23)$$

where ω_{ix} is the control bandwidth of the current control loop. The ratio between the current and voltage control loop bandwidths is selected higher than 10 to ensure the effectiveness of the cascaded-loop control.

Figure 2.7: Control architecture under the EDC.

The design of the converter controller bandwidth should consider the characteristics of the energy storage device it controlled. The response time of SC is usually in milliseconds and the response time for battery is much longer [34]. To ensure fast dynamic response of the SC system, the bandwidth of the SC converter controller ω_{vSC} should be set larger than SC bandwidth. The control bandwidth of the battery converter controller ω_{vB} should be larger than the battery bandwidth and much smaller than the SC converter bandwidth ω_{vSC} to make sure that only SC compensates fast fluctuations and battery responses to slow dynamic power. By selecting proper values for the control bandwidth of the voltage controller ω_{vx} and current controller ω_{ix} , the double-loop controller can be designed based on (2.20) and (2.23) to realize bus voltage tracking with the expected dynamics and damping of the converter system.

2.3.2 Design of the proposed strategy

Fig.2.7 shows the generalized control structure for one converter with the implementation of the EDC method. G_{ox} can be a virtual resistor or a virtual capacitor, which is expressed as

$$\begin{cases} G_{oB}(s) = R_v \\ G_{oSC}(s) = (C_v s)^{-1} \end{cases} \quad (2.24)$$

Considering the inner voltage control loop, the output impedance of the x th source converter can be obtained as

$$Z_{ox}(s) = -\frac{V_{ox}}{I_{ox}} = \frac{G_{ox}(s)G_{vx}(s)D_x + 1}{G_{vx}(s)D_x + C_{ox}s} \quad (2.25)$$

By substituting (2.17), (2.20) and (2.24) into (2.25), the output impedances of the battery, SC and the integrated HESS can be obtained as (2.26)-(2.28)

respectively after simplification.

$$Z_{oB}(s) = R_v \frac{(1/R_v + \omega_{vB} C_{oB})s + \eta \omega_{vB}^2}{s^2 + \omega_{vB}s + \eta \omega_{vB}^2} = Z_{oB}^*(s) T_{s1}(s) \quad (2.26)$$

$$Z_{oSC}(s) = \frac{1}{C_v s} \frac{C_{oSC}}{C_o} \frac{s^2 + \frac{\omega_{vSC} C_{oSC}}{C_v} s + \frac{\eta C_{oSC} \omega_{vSC}^2}{C_v}}{s^2 + \omega_{vSC}s + \eta \omega_{vSC}^2} = Z_{oSC}^*(s) T_{s2}(s) \quad (2.27)$$

$$Z_o(s) = \frac{Z_{oB}(s) \cdot Z_{oSC}(s)}{Z_{oB}(s) + Z_{oSC}(s)} \quad (2.28)$$

As can be observed from (2.26) and (2.27), $Z_{oB}(s)$ and $Z_{oSC}(s)$ are shaped by $T_{s1}(s)$ and $T_{s2}(s)$. The poles of $T_{s1}(s)$ and $T_{s2}(s)$ are at the crossover frequency of the voltage control loop and they do not impact the impedances within the voltage control bandwidth. The impact of $T_{s1}(s)$ and $T_{s2}(s)$ are therefore determined by the location of zeros. $T_{s1}(s)$ has one zero at ω_{z1} and $T_{s2}(s)$ has two zeros at $\omega_{z2,3}$, which are expressed as:

$$\omega_{z1} = \frac{C_{oB} \eta \omega_{vB}^2}{1/R_1 + \omega_{vB} C_{oB}} \quad (2.29)$$

$$\omega_{z2,3} = \sqrt{\frac{C_{oSC} \eta}{C_v}} \omega_{vSC} \quad (2.30)$$

Define the frequency range within the corner frequency ω_c as the low frequency range and the frequency range between ω_c and ω_{vx} as the high frequency range. The zero of $T_{s1}(s)$ at ω_{z1} should be larger than ω_c so that $Z_{oB}(s)$ behaves as the virtual resistance in the low frequency range. As this condition can be satisfied in most cases, $Z_{oB}(s)$ preserves the characteristics of the virtual resistor. To ensure $Z_{oSC}(s)$ behaves as virtual capacitance, two zeros at $\omega_{z2,3}$ should be located beyond the high frequency range. However, it can be observed from (2.27) that the two zeros are within the voltage controller bandwidth. To minimize their impacts, they should be located as far away from ω_c as possible, i.e. larger values of C_o and ω_{vSC} and smaller value of C_v should be selected. As the virtual capacitor C_v is determined by expected dynamics as (2.7) or (2.8), the values of C_{oSC} and ω_{vSC} should be properly designed based on (2.26), (2.27) and (2.30), and the expected controller dynamics.

To illustrate the analysis above, the frequency response of $Z_{oB}(s)$, $Z_{oSC}(s)$ and $Z_o(s)$ under different values of C_{oSC} and ω_{vSC} are investigated. System parameters are listed in Table 2.1 and three sets of C_{oSC} and ω_{vSC} are shown in Table 2.2. For each set of the parameters, the converter can operate well with

CHAPTER 2. A DECENTRALIZED DYNAMIC POWER SHARING STRATEGY FOR HYBRID ENERGY STORAGE SYSTEMS

Table 2.1: System Parameters.

Parameter	Description	Value
V_{nom}	DC bus nominal voltage	270 V
ω_c	Corner frequency of EDC	$0.1 \times 2\pi$ rad/s
R_v	Virtual Resistance	1Ω
C_v	Virtual Capacitance	$1/0.2\pi$ F
V_B	Battery voltage	200 V
V_{SC}	SC voltage	100 V
L_B, L_{SC}	Inductors of battery and SC converters	0.002 H
C_{oB}	Battery Converter Output Capacitor	220 μ F
ω_{vB}	Battery Converter voltage control bandwidth	$2\pi \times 100$ Hz

Table 2.2: Cases of Parameter Design.

Cases	SC converter voltage control bandwidth(ω_{vSC})	SC Converter Output Capacitor(C_{oSC})
1	$2\pi \times 20$ Hz	4×470 μ F
2	$2\pi \times 100$ Hz	470 μ F
3	$2\pi \times 100$ Hz	4×470 μ F

Figure 2.8: Frequency response of $Z_{oB}(s)$, $Z_{oSC}(s)$ and $Z_o(s)$ for Case 1.

enough phase margin in standalone operation.

The frequency response of $Z_{oB}(s)$, $Z_{oSC}(s)$ and $Z_o(s)$ for Case 1, Case 2 and

Figure 2.9: Frequency response of $Z_{oB}(s)$, $Z_{oSC}(s)$ and $Z_o(s)$ for Case 2.

Figure 2.10: Frequency response of $Z_{oB}(s)$, $Z_{oSC}(s)$ and $Z_o(s)$ for Case 3.

Case 3 are shown in Fig. 2.8, Fig. 2.9 and Fig. 2.10 respectively. It can be observed that, $Z_o(s)$ is dominated by $Z_{oB}(s)$ and behaves as a virtual resistor in low frequency range for all three cases. However, in the high frequency range, if

Figure 2.11: Peak magnitude of $Z_o(s)$ in high frequency range with variation of C_{osc} and ω_{vsc} .

C_{osc} and ω_{vsc} are not properly designed, $Z_{osc}(s)$ is shaped and may not behave as a virtual capacitor. For Case 1 where a small ω_{vsc} is selected, $|Z_{osc}(s)|$ is close to $|Z_{oB}(s)|$ in the high frequency range and $Z_o(s)$ is not dominated by $Z_{osc}(s)$, as shown in Fig. 2.8. This means that both the battery and SC systems will respond for the high frequency fluctuations during load change and as a result, dynamic current sharing is not achieved. Moreover, peak gain of $Z_o(s)$ in the high frequency range exceeds its gain in the low frequency range. This is also an undesirable result as it may introduce disturbances and cause oscillations in the bus voltage. Similar undesirable result can be observed in Fig. 2.9 for Case 2 with an improper value of C_{osc} . For Case 3, $Z_o(s)$ is dominated by $Z_{osc}(s)$ in the high frequency range, as shown in Fig. 2.10. This means that only the SC converter system will respond for high frequency fluctuations and thus the control objective is achieved. Meanwhile, since $|Z_o(s)|$ at high frequency range is suppressed below its low frequency gain, disturbance in the bus voltage can be suppressed. Therefore, Case 3 provides a desirable set of C_{osc} and ω_{vsc} .

It can also be observed that when $|Z_o(s)|$ in the high frequency range is suppressed below its low frequency gain, $Z_o(s)$ is dominated by $Z_{osc}(s)$ at the high frequency range. This result can be seen as a criterion to select C_{osc} and ω_{vsc} . Fig. 2.11 shows the peak magnitude of $Z_o(s)$ at the high frequency range with the

Figure 2.12: Equivalent source-load model for HESS feeding the CPL.

variation of C_{oSC} and ω_{vSC} , and its comparison with low frequency gain, which is the magnitude of virtual resistor (i.e. the value is unity in this chapter). With the increase of C_{oSC} and ω_{vSC} , the peak gain of $Z_o(s)$ will decrease to below unity. Therefore, larger C_{oSC} and ω_{vSC} are preferred. But they cannot be too large considering physical constraints and system dynamics. By properly selecting a set of C_{oSC} and ω_{vSC} that makes the high frequency gain of $Z_o(s)$ less than its low frequency gain, the transient power sharing can be achieved and oscillations in the bus voltage can be suppressed.

To better illustrate the effectiveness of the design guideline for the proposed controller, time domain simulation results will be presented in case studies in Section 2.5.

2.4 Some Considerations

2.4.1 Stability analysis

Recently, power electronic converters are widely applied in DC MG. Tightly regulated power converter loads behave as constant power loads (CPLs) and the interactions of CPLs with other subsystems may degrade system stability. To investigate the interactions, the system is usually divided into a source subsystem and a load subsystem and the stability is determined by the ratio of their impedances [22, 45]. To assess system stability, there are many stability criteria, which have been reviewed in [45]. The output impedance of the HESS derived in (2.26)-(2.28) can be used for stability analysis.

As the load model is not the focus in this work, the HESS feeding an ideal CPL P_{CPL} is studied here. Fig. 2.12 shows the equivalent circuit, where $Z_o(s)$

is the equivalent output impedance of the source and $Z_i(s)$ is the equivalent input impedance of the load. System stability can be determined by the ratio of $Z_o(s)/Z_i(s)$ and according to the Middlebrook stability criterion [45], $Z_o(s)$ should be much smaller than $Z_i(s)$ in all frequency range to ensure system stability, i.e.

$$|Z_o(s)| < |Z_i(s)| \quad (2.31)$$

Since the ideal CPL model is adopted [58], its load impedance is expressed as

$$Z_i(s) = -\frac{V_{bus}^2}{P_{CPL}} \quad (2.32)$$

As discussed in the previous section, to ensure the dynamic current sharing, the parameters should be properly designed that the high frequency gain of $Z_o(s)$ is less than its low frequency gain, i.e.

$$|Z_o(s)|_{\max} = R_v \quad (2.33)$$

Combination of (2.31)-(2.33) gives the condition in (2.34) and maximum load that is allowed is calculated as (2.35) when V_o is at $0.5V_{nom}$.

$$|Z_i(s)| > R_v \quad (2.34)$$

$$P_{\max} = \frac{V_{nom} - V_{bus}}{R_v} \cdot V_{bus} = \frac{V_{nom}^2}{4R_v} \quad (2.35)$$

Considering the operation requirements of these two converters, the bus voltage cannot be lower than the battery voltage V_B or SC voltage V_{SC} . Therefore, the maximum possible output power is modified as

$$P_{\max} = \frac{V_{nom} - V_{bus_min}}{R_v} \cdot V_{bus_min} \quad (2.36)$$

where

$$V_{bus_min} = \max \left\{ 0.5V_{nom}, V_B, V_{SC} \right\} \quad (2.37)$$

It should be noticed that the proposed method helps to stabilize the system at the same time. If only the conventional droop control is implemented, the output impedance of the source is obtained as $Z_{oB}(s)$ in (2.26). The frequency response of $Z_{oB}(s)$ can be observed from Fig. 2.8, Fig. 2.9 and Fig. 2.10. In some

frequency range, $|Z_{oB}(s)|$ exceeds its low frequency gain, i.e.

$$|Z_{oB}(s)|_{\max} > R_v \quad (2.38)$$

Then the maximum allowable power will be less than that in (2.36). This is because the virtual capacitance controller helps to suppress the output impedance gain at high frequency range, as discussed in Section 2.3.2. Therefore, the proposed method can achieve the dynamic power sharing and improve the system stability margin in the meantime.

It should be mentioned that the Middlebrook criterion is criticized to be conservative as it only considers magnitudes of impedances. But for the ideal CPL model whose input impedance is a constant value with phase value of -180° , as shown in (2.32), the stability result is no difference if a more liberal criterion is adopted. For example, the Gain Margin Phase Margin (GMPM) criterion relaxes the conservativeness by allowing $Z_o(s) > Z_i(s)$ at some frequency range when the phase difference is less than 180° . However, as can be observed from Fig. 2.11, the peak value of $|Z_o(s)|$ is at the low frequency range with phase of 0° . Thus the phase difference of $Z_o(s)$ and $Z_i(s)$ at low frequency part is always 180° . Considering (2.33), to satisfy the GMPM criterion, the stability condition in (2.34) should be met, which is the same as the Middlebrook criterion result. As Middlebrook criterion is not conservative in this case and it is simple for the design, it is chosen in this chapter. If a specific load model is adopted such as a converter load model, more liberal criteria in [45] can be selected with the source impedance derived in (2.26)-(2.28) to investigate the system stability.

2.4.2 State-of-charge regulation of energy storage devices

It should be mentioned that the proposed approach is for the battery and SC operating within their state-of-charge (SoC) boundaries. When the battery and SC are out of SoC boundaries, they will switch to other operating modes. For example, when SC SoC reaches its lower boundary, the SC converter controller will switch to the charging mode and the battery will charge SC until its SoC reaches nominal value, and then SC converter controller will switch back to the dynamic power sharing mode [59,60]. But there is a strict requirement for seamless mode transition and when SC operates in the SoC regulation mode, it cannot compensate load fluctuations. These drawbacks are overcome by the proposed approach in Chapter 3 that can achieve dynamic power sharing and SC SoC

restoration autonomously.

The proposed method is for real-time operation of the hybrid battery/SC system to compensate residual power of RESs generation and load consumption. As the energy capacity of the battery bank is much higher than that of SC and the battery is scheduled as the energy source to compensate the steady-state residual power, the SoC variation of the battery is much lower than that of SC in real-time operation. In this case, the battery is assumed to operate within its SoC limitation and there is no active regulation for the battery SoC in the proposed control algorithm. To regulate the battery SoC, other energy sources, like diesel generators, should be available as the alternative energy sources and the proposed strategy is flexible to combine with high level control algorithms with communication networks such as [61] and [62] to deal with this situation. For example, when the battery SoC reaches the higher boundary, the battery changes to the floating mode and RESs switch to the voltage control mode to support the bus voltage; when the battery SoC drops to the lower boundary and the output of RESs is also low, the alternative energy sources should be available to support the bus voltage and charge the battery, or otherwise, load shedding scheme is required.

2.4.3 System expansion with multiple energy storage devices

The proposed control method is flexible for system expansion. Fig. 2.13 demonstrates a hybrid system with M batteries and N SCs under the EDC method. It can be simplified to Fig. 2.2 by taking into account that:

$$\frac{1}{R_{v,eq}} = \sum_{m=1}^M \frac{1}{R_{vm}}, C_{v,eq} = \sum_{n=1}^N C_{vn} \quad (2.39)$$

where R_{vm} is the virtual resistance of the m th battery; C_{vn} is the virtual capacitance of the n th SC; $R_{v,eq}$ and $C_{v,eq}$ are the equivalent virtual resistance and virtual capacitance.

When the capacity of the system is to be expanded, the virtual resistance of the $(M + 1)$ th battery, $R_{v(M+1)}$, can be calculated by (2.1) based on its power rating. To maintain the same system dynamics, the virtual capacitance of the new SC branch ($C_{v(N+1)}$) is given by

$$C_{v(N+1)} = \frac{R_{v,eq}}{R_{v(M+1)}} \cdot C_{v,eq} \quad (2.40)$$

Figure 2.13: Equivalent circuit with the integration of M batteries and N SCs.

Substitution of (2.40) into (2.39) yields to

$$R'_{v,eq} C'_{v,eq} = \left(\frac{R_{v,eq} \cdot R_{v(M+1)}}{R_{v,eq} + R_{v(M+1)}} \right) \cdot \left(1 + \frac{R_{v,eq}}{R_{v(M+1)}} \right) \cdot C_{v,eq} = R_{v,eq} C_{v,eq} \quad (2.41)$$

It can be observed from (2.41) that the system dynamics will not change when the virtual resistance of the new battery branch is given by (2.40). Thus, the system capacity can be easily expanded without affecting the existing system dynamics.

2.5 System Studies

The system diagram in Fig. 2.4 is implemented in PLECS simulation and the experimental platform to verify the proposed EDC method and analytical results. The parameters in Table 2.1 and Table 2.2 are adopted. I_{oB} and I_{oSC} are the output current of the battery converter and SC converter respectively.

Figure 2.14: Simulation result for normal operation.

2.5.1 Simulation results

Fig. 2.14 shows the simulation result with the properly designed parameters in Case 3. When the load increases suddenly, I_{oSC} increases rapidly and later decreases gradually to zero; I_{oB} increases gradually and after several seconds becomes constant to supply the total load. Similarly, when the load steps down, SC only responds for transient fluctuations whereas the battery responds gradually to provide steady state load power. This shows the effectiveness of the proposed controller and the designed parameters.

Fig. 2.15 shows the simulation result of the unstable situation. According to (2.36) and (2.37), the maximum power supply is calculated as 14kW and the DC voltage should be maintained above 200V. When the load is 13.5 kW, system remains stable. At 15s, when the load is increased to 14.5 kW, the bus voltage will gradually drop to 200V and the system starts to oscillate until the voltage and current limitations are reached; the converter cannot work properly. In the real situation, the bus voltage deviation is limited as (2.1) and therefore, condition (2.36) and (2.37) will not be violated and the system always operates in stable region.

The impact of the transmission line is simulated in Fig. 2.16 with the line resistances R_{l1} and R_{l2} both set at 0.5Ω . Other parameters are the same as the normal operation without line resistances in Fig. 2.14 and the same load change is applied. Fig. 2.16 shows that dynamic current sharing between the battery and SC is still achieved, i.e. SC compensates fast fluctuations and the battery supplies

Figure 2.15: Simulation result for unstable operation.

Figure 2.16: Simulation result considering transmission line.

steady state power. Compared with the simulation result in Fig. 2.14, the system dynamics is slightly slower and this impact can be minimized by reducing the virtual capacitance C_v . This validates the effectiveness of the proposed method when the line resistance is considered.

To demonstrate the scalability of the proposed method, the HESS in Fig. 2.4 is expanded to a system with two batteries and two SCs and the simulation result for the expanded system is shown in Fig. 2.17. The parameters are the same as that in normal operation with transmission lines in Fig. 2.16. The additional

Figure 2.17: Simulation result for the expanded system with two batteries and two SCs.

battery and SC are the same as the original ones and their droop coefficients are set at 1Ω and $1/(2\pi \times 0.1) \text{ F}$ to maintain the same system dynamic response. The DC load increases from 0.2kW to 2.2kW at 5s and decreases from 2.2kW to 0.2kW at 10s. It can be observed that SCs respond to load change immediately and the batteries supply load at steady state. System dynamics are the same with those in Fig. 2.16. This validates the effectiveness of the proposed method for system expansion.

2.5.2 Experimental results

A laboratory test platform is built to verify the effectiveness of the proposed EDC method, as shown in Fig. 2.18. A 195V lead-acid battery bank represents the battery in this chapter as it has slower dynamic response. A 95V Li-ion battery bank represents SC as it has faster dynamic response and higher power density. Two bidirectional DC/DC converters work as interface converters. They are built by IGBT modules and controlled by TI TMS320F28335 DSP respectively. A DC

Figure 2.18: Experiment setup.

Figure 2.19: Experiment result with desirable parameter (Case 3).

power supply and a resistive load are connected to the DC bus as the equivalent load of HESS.

1) Validation of proposed design procedure

Fig. 2.19 shows the experiment results with the properly designed parameters in Case 3. When the load steps up/down, I_{oSC} increases rapidly and later decreases gradually to zero, whereas I_{oB} increases/decreases gradually and at steady state becomes constant to supply total load. Fig. 2.20 shows the dynamics for Case 1 when the parameter is improperly selected. During transient, the battery also responds immediately, which cannot meet the control objective. This verifies the proposed design procedure and criterion.

2) Different dynamic response

As illustrated that the dynamic response of current sharing is determined by the time constant or corner frequency of the EDC strategy, which is related

Figure 2.20: Experiment result with undesirable parameter (Case 1).

Table 2.3: Parameters for Different Power Sharing Dynamics.

Cases	Corner Frequency of EDC (ω_c)	Virtual Resistor (R_v)	Virtual Capacitor (C_v)
3	$2\pi \times 0.1$ rad/s	1Ω	$1/(2\pi \times 0.1)$ F
4	$2\pi \times 0.5$ rad/s	1Ω	$1/(2\pi \times 0.5)$ F
5	$2\pi \times 0.1$ rad/s	5Ω	$1/(2\pi \times 0.1)$ F

Figure 2.21: Experiment result for Case 4.

with product of the virtual resistor and virtual capacitor. Parameters for different power sharing dynamics are shown in Table 2.3. Other parameters are from Table 2.1 and Table 2.2 Case 3 to ensure the EDC implementation.

The experiment result for Case 3 is already shown in Fig. 2.19. Fig. 2.21 shows the result for Case 4 and compared with Fig.2.19, it turns out that the current sharing dynamics increase to 5 times when corner frequency increases

Figure 2.22: Experiment result for Case 5.

Figure 2.23: Experiment result for negative power compensation.

to 5 times. Fig. 2.22 shows the dynamics for Case 5. Compared with Fig. 2.19, the current sharing response is almost the same as the corner frequency of EDC strategy remains the same. Therefore, as mentioned in (2.7) and (2.8), the expected power sharing dynamic response can be realized by selecting the proper EDC parameters.

3) HESS in charging mode

The above simulation and experiment validate the effectiveness of the proposed method for the HESS when it supplies the power. As the role of HESS is to compensate the imbalanced power between the load and source (demand minus generation), which can be either positive or negative, the negative case with the implementation of the proposed method should be studied as well. Fig. 2.23 shows the experiment result when the HESS absorbs the extra net power. Initially only

resistive load is connected to the bus and the load current is only supplied by battery. The parameters are the same as that in Fig. 2.19. At about 4s, the DC power source supplies power to the DC bus and its value exceeds the load demand. To absorb the imbalanced power, the SC converter output current I_{oSC} increases in the negative direction immediately and decreases to zero gradually. Meanwhile, I_{oB} reduces gradually to a negative value and at steady state, the battery absorbs all the extra power. Thus the dynamic power sharing between the battery and SC is achieved. A similar dynamic response can be observed when the DC power source is disconnected at 12s. Therefore, the effectiveness of the proposed method is validated when HESS operates in the charging mode and absorbs residual power.

2.6 Conclusion

This chapter proposes a decentralized control strategy for dynamic current sharing in HESS. The output voltage characteristics of the battery and SC are regulated by a virtual resistance droop controller and a virtual capacitance droop controller respectively. The proposed droop controllers inherently act as an LPF for the battery and an HPF for SC to the load response so that SC compensates high frequency fluctuations during transient and the battery provides load power at steady state. For practical implementation of this control strategy, a detailed procedure for parameter design is proposed. With the developed system dynamic model and derived transfer functions, the parameters can be properly designed to ensure stable operation, bus voltage regulation and dynamic current sharing. System expansion is also demonstrated to show the scalability of the proposed method. Simulations and experiments are conducted to validate the effectiveness of the proposed control strategy and analytical results.

Chapter 3

A Decentralized Power Management Strategy for Bus Voltage Restoration and SoC Recovery

This chapter is a continuation of the work in Chapter 2. In Chapter 2, a decentralized control strategy is proposed for dynamic power sharing of the hybrid energy storage system. However, this approach does not consider the issues of the bus voltage deviation and SC SoC violation. For further improvement, a decentralized power management strategy is proposed in this chapter to achieve the dynamic power split, bus voltage restoration and SC SoC recovery.

3.1 Introduction

As introduced in Chapter 2, the hybrid energy storage system provides an effective solution for the integration of RESs into MGs and to fully exploit the advantages of the complementary energy storage devices in HESS, it is desirable that the SC compensates fast fluctuations and the battery provides smooth power at steady state [32, 36]. Moreover, considering the low energy density of SC, its SoC should be regulated to avoid boundary violation and operating failure [20]. The bus voltage should also be regulated at its nominal value to improve power quality and ensure lifetime operation of the voltage sensitive load [56]. To achieve these

objectives, many power management strategies have been proposed in literatures [32,33,39,40,49,50]. However, most of them are centralized or hierarchical control methods which require a centralized controller to process the information from local units and make the decisions. The communication delay or failure may cause system reliability issues and the expansion or reconfiguration of the system will lead to redesign of the central controller with increased complexity.

To increase system reliability and scalability, decentralized control strategies without central controllers and communications should be developed. Ref [63] proposes a voltage droop based power sharing strategy for a grid connected converter and an energy storage device during normal and islanding operations. With a high droop gain and an HPF implemented for the energy storage device, the energy storage device is ensured to absorb the high frequency components of load to smooth the power supplied by the grid connected converter. In [64], two DC voltage control strategies are proposed for the coordination of the line side converter (LSC) and the energy storage device, one is that a PI controller is employed for the LSC and a high droop gain controller is adopted for the energy storage device, and the other is that a single PI controller is employed for the voltage loop with its output separated into the high and low components for the energy storage device and the LSC by an LPF. However, for the above methods, the SoC regulation of energy storage device is not considered and thus, the continuous operation of the energy storage devices may lead to the violation of SoC limitation and unsafe situations. Furthermore, the bus voltage deviation resulted from the droop-based control during islanding mode may degrade power quality. In [53], a frequency-coordinating virtual impedance strategy is proposed for autonomous operation of the hybrid battery/SC system by implementing a proportional controller with an LPF for the battery and a proportional controller with an HPF for SC. The SC voltage/SoC is also restored by an extra PI controller. However, the impact of the SC SoC loop on the dynamic power sharing performance is not considered and the bus voltage deviation issue remains to be solved. In [59], a virtual impedance power management strategy is proposed for SC to ensure its role as a transient power buffer. To regulate SC SoC, an SoC regulation loop with mode switching strategy is employed. However, there is a strict requirement for seamless mode transition and when SC operates in SoC regulation mode, it cannot compensate load fluctuations. In addition, the bus voltage deviation, which may impact voltage sensitive load and cause power quality issues, is still ignored.

In Chapter 2, a decentralized extended droop control strategy is proposed to

enable autonomous power sharing of hybrid battery/SC system. A virtual resistance droop (VRD) controller is implemented for battery and a virtual capacitance droop (VCD) controller is employed for SC so that SC responds to fast fluctuations immediately and the battery provides total power at steady state. However, the issues of the bus voltage restoration and SC SoC recovery are not taken into consideration. For improvement, this chapter proposes a control strategy consisting of a virtual high pass filter based droop (VHPFD) controller for the battery converter and a VCD controller for the SC converter. It can achieve the dynamic power sharing, bus voltage restoration and SC SoC recovery in a decentralized manner without communications. A guideline for parameter design is also illustrated to ensure expected system dynamics. Main improvements compared with previous works include:

1)*Decentralized bus voltage restoration:* The cooperation of the proposed VHPFD and VCD ensures that SC will not supply/absorb any current at steady state, and as a result, the bus voltage is regulated by the battery converter and can be restored to its nominal value at steady state by the proposed HPFD.

2)*Autonomous SC SoC recovery:* The SC SoC is enforced to its initial value autonomously along with the restoration of the bus voltage.

This chapter is organized as follows: First, the decentralized power management strategy for dynamic power sharing, bus voltage restoration and SC SoC recovery is proposed. Second, the practical implementation and parameter design procedure are presented with a detailed frequency domain analysis. Then, some considerations including line impedance impacts, system expansion and battery SoC regulation are discussed. Finally, simulations and experiments are conducted to verify the effectiveness of the proposed strategy and the analytical results.

3.2 The Proposed Power Management Strategy

The isolated DC MG in Chapter 2 is continuously studied and it is replotted in Fig. 3.1 for convenience. RESs operate in maximum power point tracking mode with output current i_{RESs} and the current of the lumped DC loads is denoted as i_{Load} . The hybrid battery/SC system will compensate the power mismatch between RESs and load to maintain the bus voltage. The current relationship at

Figure 3.1: Schematic diagram of DC MG with the integration of HESS.

Figure 3.2: Equivalent circuit for autonomous power split under VRD and VCD.

the DC bus is expressed as

$$\begin{cases} i_o = i_{load} - i_{RESs} \\ i_{oSC} + i_{oB} = i_o \end{cases} \quad (3.1)$$

where i_o is the equivalent load current of HESS and can be either positive or negative; when i_o is positive, the HESS supplies the equivalent load current; when i_o is negative, the HESS is charged by the equivalent load current.

For the power management of HESS, it is expected that SC buffers fast fluctuations and the battery compensates the low frequency power mismatch. Meanwhile, the bus voltage should be regulated at the nominal value and SC SoC should also be taken care of to ensure continuous and reliable operation.

3.2.1 Dynamic power sharing

To achieve the dynamic power sharing, the extended droop control scheme of VCD and VRD, which is proposed in Chapter 2, is implemented first. Fig. 3.2 shows the equivalent circuit with the implementation of VRD and VCD.

The output voltage-current ($V-I$) relationship of the battery converter under the VRD control is given by

$$V_{oB} = V_{nom} - R_v \cdot I_{oB} \quad (3.2)$$

where V_{nom} is the nominal voltage; V_{oB} and I_{oB} are the output voltage and current of the battery converter. The virtual resistance R_v is determined by the maximum bus voltage deviation ΔV_{max} and converter current rating $I_{oB,max}$, i.e.

$$R_v = \frac{\Delta V_{max}}{I_{oB,max}} \quad (3.3)$$

The V-I relationship of the SC converter under the VCD control is derived as

$$V_{oSC} = V_{nom} - \frac{1}{sC_v} \cdot i_{oSC} \quad (3.4)$$

where V_{oSC} and i_{oSC} are the output voltage and current of the SC converter; C_v is the virtual capacitance.

As the line impedance is much smaller than the droop coefficient, the voltage drop across the transmission line is negligible compared with the voltage drop caused by the droop control [57]. Hence,

$$V_{oB} = V_{oSC} = V_o \quad (3.5)$$

where V_{bus} is the voltage of the common bus.

Based on (3.1), (3.2), (3.4) and (3.5), the current sharing relationship between the battery and SC is derived as

$$\begin{cases} i_{oB} = G_{B0}(s) \cdot i_o = \frac{1}{R_v C_v s + 1} \cdot i_o \\ i_{oSC} = G_{SC0}(s) \cdot i_o = \frac{R_v C_v s}{R_v C_v s + 1} \cdot i_o \end{cases} \quad (3.6)$$

As can be observed from (3.6), the load current is automatically split into the low and high frequency components for the battery and SC by G_{B0} and G_{SC0}

with the cutoff frequency expressed as

$$\omega_c = \frac{1}{R_v C_v} \quad (3.7)$$

Therefore, the implementation of VCD and VRD enables the autonomous power split for SC and battery, as presented in Chapter 2. But there exists bus voltage deviation resulted from the droop-based technique and the SoC regulation of SC is not taken into consideration. The main improvements of the work in this chapter will be presented in the following sections.

3.2.2 Bus voltage restoration

As the VRD strategy results in output voltage deviation, an compensation voltage can be added to the reference voltage of the battery converter with the VRD loop for the voltage restoration along with the autonomous power sharing. The $V - I$ characteristics in (3.2) is modified as

$$V_{oB} = V_{nom} - R_v \cdot i_{oB} + \Delta V_{VR} \quad (3.8)$$

To generate the compensation voltage, the output voltage of the battery converter is compared with its nominal value and the difference is processed by an integrator controller, i.e.

$$\Delta V_{VR} = \frac{k_i}{s} (V_{nom} - V_{oB}) \quad (3.9)$$

The output V/I characteristics of the battery converter before and after the implementation of the voltage compensation loop are shown in Fig. 3.3, denoted as Curve 1 and Curve 2. To provide the equivalent load current i_o , output voltage drops to V_o when the compensation loop is not implemented, as described by Curve 1.

To restore the bus voltage, the reference voltage is increased by ΔV_{VR} and the new curve is obtained as Curve 2. Then the output voltage can be regulated at the nominal value V_{nom} . It is similar if i_o is negative. Thus the bus voltage restoration is completed.

By substituting (3.9) into (3.8) and rearranging it, the V-I characteristics of

Figure 3.3: V-I characteristics with and without voltage restoration.

the battery converter is derived as

$$V_{oB} = V_{nom} - \frac{R_v s}{s + k_i} \cdot i_{oB} \quad (3.10)$$

It can be observed from (3.10) that the insertion of voltage restoration term in (3.8) is equivalent to adding an HPF to the virtual resistance in the conventional VRD. Thus the proposed strategy is named as virtual HPF based droop (VHPFD) strategy. As the HPF blocks DC component, the output voltage of the battery converter returns to its nominal value at steady state and the voltage deviation caused by conventional VRD is eliminated.

The VCD strategy in (3.4) remains for the SC converter regulation so that it can operate as short circuit during transient and open circuit at steady state. As the SC converter has zero current sharing at steady state, its output voltage follows the output voltage of the battery converter. Thus the autonomous bus voltage restoration is achieved with the cooperation of VHPFD and VCD strategies.

3.2.3 State-of-Charge recovery

The proposed strategy has another advantage that SC SoC is automatically recovered along with the restoration of the bus voltage and no additional controller is needed.

Based on the definition, SC SoC can be obtained as

$$SoC = SoC_{ini} - \frac{\int i_{SC} dt}{Q_c} \quad (3.11)$$

where SoC_{ini} , i_{SC} and Q_c are the initial SoC, terminal current and capacity of SC.

At steady state, the bus voltage is regulated at its nominal value and according

to (3.5), it yields

$$V_{oB} = V_{oSC} = V_{bus} = V_{nom} \quad (3.12)$$

Combination of (3.4) and (3.12) gives

$$-\frac{1}{C_v} \cdot \int_{t_0}^{t_s} i_{oSC} dt = 0 \quad (3.13)$$

where t_0 is the starting time of transient and t_s is the time when the bus voltage is restored to the nominal value at steady state.

Neglecting the conversion loss, the relationship of the input and output power of the SC converter is expressed as

$$V_{SC} i_{sc} = V_{oSC} i_{oSC} \quad (3.14)$$

where V_{SC} is the SC voltage and can be assumed as a constant value considering the slow variation of its SoC during transient [26].

Based on (3.11)(3.13) and (3.14), the variation of the SC SoC from the start of load variation to the completeness of the voltage restoration is expressed as

$$\Delta SoC = -\frac{\int_{t_0}^{t_s} i_{SC} dt}{Q_C} = -\frac{C_v V_{oSC}}{Q_e V_{SC}} \left(\frac{1}{C_v} \cdot \int_{t_0}^{t_s} i_{oSC} dt \right) = 0 \quad (3.15)$$

Eqn. (3.15) reveals that the SC SoC recovery is achieved automatically with the restoration of the bus voltage without any additional controllers. Once the original SC SoC is set at the reference value, SC SoC can be regulated around its reference value.

For the battery SoC regulation, as the energy capacity of the battery bank is much higher than that of SC and the battery is scheduled as the energy source to compensate the steady-state residual power, the SoC variation of the battery is much lower than that of SC in real-time operation. Therefore, the battery is assumed to operate within its SoC limitation and there is no active regulation for the battery SoC in the proposed control algorithm. The SoC management of battery can be left for long term energy management (with a time scale of 15 minutes to several hours). The focus of this thesis is on realtime control with a time scale of several seconds to minutes. More detailed discussions have been presented in Section 2.4.2.

Figure 3.4: System model and control architecture under the proposed strategy.

3.3 Practical Implementation and Parameter Design

Fig.3.4 shows the control architecture for HESS under the proposed power management strategy. The proposed VHPFD strategy is employed for the battery controller and the VCD strategy is for the SC converter regulation. The classical cascaded PI control is employed for the inner voltage regulation and current tracking of the two bidirectional DC/DC converters.

For the inner voltage and current loops, the cascaded PI controllers are employed and the detailed parameter design procedure can follow Section 2.3.1 in Chapter 2. For the outer loops, their cutoff frequencies are determined by the expected transient power sharing requirement and the battery dynamics, which would be much lower than the bandwidths of the inner converter control loops. Therefore, the dynamics of the inner loops can be neglected when designing the outer loops.

As the proposed VHPFD scheme shapes the conventional VRD with an HPF, the enabled voltage/SoC restoration may impact the autonomous transient power split for the SC and battery. Parameters should be carefully designed to decouple the two processes to ensure the expected system dynamics. Considering (3.12) and combining (3.1), (3.4) and (3.10), current sharing of the battery and SC can be obtained as

$$i_{oB} = G_B(s) \cdot i_o = \frac{\frac{1}{R_v C_v} s + \frac{k_i}{R_v C_v}}{s^2 + \frac{1}{R_v C_v} s + \frac{k_i}{R_v C_v}} \cdot i_o \quad (3.16)$$

$$i_{oSC} = G_{SC}(s) \cdot i_o = \frac{s^2}{s^2 + \frac{1}{R_v C_v} s + \frac{k_i}{R_v C_v}} \cdot i_o \quad (3.17)$$

Equations (3.16) and (3.17) show that the load current is split by second-order filters $G_B(s)$ and $G_{SC}(s)$ for the battery and SC. As system dynamics should ensure the smooth response of the battery, the cutoff frequency of $G_B(s)$ should be properly designed. To minimize the impact of the voltage/SC SoC restoration process on autonomous power split and achieve the expected system dynamics, the cutoff frequency of $G_B(s)$ should be selected the same as ω_c in (3.7) with the VRD and VCD approach.

Equations (3.16) and (3.17) can be rewritten in the standard form as

$$i_{oB} = G_B(s) \cdot i_o = \frac{2\xi\omega_n s + \omega_n^2}{s^2 + 2\xi\omega_n s + \omega_n^2} \cdot i_o \quad (3.18)$$

$$i_{oSC} = G_{SC}(s) \cdot i_o = \frac{s^2}{s^2 + 2\xi\omega_n s + \omega_n^2} \cdot i_o \quad (3.19)$$

with the natural frequency ω_n and damping ratio ξ expressed as

$$\omega_n = \sqrt{\frac{k_i}{R_v C_v}} \quad (3.20)$$

$$\xi = \frac{1}{2R_v C_v \omega_n} \quad (3.21)$$

The definition of cutoff frequency gives the condition that the gain of $G_B(s)$ at ω_c should be -3dB [65], i.e.

$$|G_B(j\omega_c)| = \frac{\sqrt{1 + 4\xi^2 \frac{\omega_c^2}{\omega_n^2}}}{\sqrt{\left(1 - \frac{\omega_c}{\omega_n}\right)^2 + 4\xi^2 \frac{\omega_c^2}{\omega_n^2}}} = \frac{1}{\sqrt{2}} \quad (3.22)$$

By solving (3.22), the natural frequency ω_n is derived as

$$\omega_n = \frac{\omega_c}{\left[(1 + 2\xi^2) + \sqrt{(1 + 2\xi^2) + 1}\right]^{\frac{1}{2}}} \quad (3.23)$$

Based on (3.20) and (3.21), and given the expected cutoff frequency and damping ratio, the parameters k_i and C_v can be designed as

$$C_v = \frac{1}{2R_v \xi \omega_n} \quad (3.24)$$

$$k_i = \omega_n^2 R_v C_v \quad (3.25)$$

(a) $G_B(s)$

(b) $G_{SC}(s)$

Figure 3.5: Bode plots for $G_B(s)$ and $G_{SC}(s)$ with different cutoff frequencies.

With the design guideline in (3.23), (3.24) and (3.25), the desired system dynamics can be ensured.

Fig.3.5 shows the bode diagrams for $G_B(s)$ and $G_{SC}(s)$ with the expected cutoff frequencies at 0.2Hz, 0.4Hz, 0.6Hz and the damping ratio is fixed at 0.8. The results reveal that the expected cutoff frequencies are obtained based on the design guideline. For each pair of the bode diagrams of $G_B(s)$ and $G_{SC}(s)$, $G_B(s)$ only allows signals below its cutoff to pass and $G_{SC}(s)$ allows signals above its cutoff frequency to pass. Thus the load power is automatically split into low and high frequency components for the battery and SC.

Figure 3.6: Bode plots for $G_B(s)$ with different damping.

It can be observed from Fig. 3.5a that at some frequency range near the cutoff frequency, the magnitude of $G_B(s)$ is above 0 dB, i.e. the magnitude of the battery current is larger than the load current and the battery is charging SC (or reverse if the load current is negative). This is the frequency range where the voltage restoration and SC SoC recovery operate. The higher the peak magnitude of $G_B(s)$ exceeds 0 dB, the higher the maximum charging/discharging current and thus the faster the SoC recovery speed will be. Fig. 3.5a also reveals that the peak gains of $G_B(s)$ at different cutoff frequencies are the same, which indicates that the maximum charging current from the battery to SC is the same under the same damping ratio. As the SoC recovery and voltage restoration share the same process, voltage restoration has the same feature. Since their operating frequency range is below the cutoff frequency of transient power split, they will not impact transient dynamics of SC and the two loops are decoupled.

Fig. 3.6 shows the bode diagrams with the damping ratios set at 0.6, 0.8, 1 and 1.2, and the cut off frequency at 0.4 Hz. With the increase of the damping ratio, the peak gain of $G_B(s)$ decreases. Thus a smaller damping ratio leads to faster speed of voltage restoration and SoC recovery. A proper damping ratio should be selected considering the tradeoff between the damping and dynamics.

Therefore, with the designed parameters, the proposed strategy can achieve the expected power split, voltage restoration and SoC recovery at the same time with negligible interactions.

3.4 Some Considerations

3.4.1 Impact of line impedance

In practice, there exists line impedance between the output terminal of the battery converter and DC bus R_{l1} and line impedance between the output terminal of the SC converter and DC bus R_{l2} . In this case, as the battery supplies all the current and SC will not supply any current at steady state, there is a voltage deviation between the battery converter output voltage and bus voltage, whereas the SC converter output voltage still equals to the bus voltage. Then the steady state voltage relationship in (3.12) is modified as

$$V_{oB} = V_{bus} - R_{l1}I_{oB}, V_{oSC} = V_{bus}, V_{oB} = V_{nom} \quad (3.26)$$

As the voltage regulation is at the battery output converter, if the impedance R_{l1} is large, the accuracy of the voltage regulation and SC SoC recovery will be suffered. Thus, the impedance between the battery system and DC bus should be as small as possible i.e. $R_{l1} \approx 0$, and it is recommended that the battery system should be located close to the DC bus.

If the impedance between SC system and DC bus is considered, (3.4) is modified as

$$V_{oSC} = V_{nom} - \frac{1}{sC_v} \cdot i_{oSC} - R_{l2}i_{oSC} \quad (3.27)$$

The current sharing of SC in (3.19) is modified by combining (3.10) and (3.27) as

$$i_{oSC} = G_{SC}(s) \cdot i_o = \frac{s^2}{s^2 + \frac{1}{(R_v + R_{l2})C_v}s + \frac{k_i}{(R_v + R_{l2})C_v}} \cdot i_o \quad (3.28)$$

with the new natural frequency ω_n expressed as

$$\omega_n = \sqrt{\frac{k_i}{(R_v + R_{l2})C_v}} \quad (3.29)$$

As can be observed, SC still only compensates the high frequency power as expected whereas the current sharing dynamics is a little slower due to the existence of the line impedance. The slower dynamic response can be eliminated by tuning the virtual capacitance value C_v to ensure the desired dynamic response. Therefore, the proposed method is still effective with the line impedance between the SC and DC bus.

It is also worth mentioning that if the line impedance between the battery and DC bus cannot be ignored, the proposed strategy is flexible to work with higher level control strategies such as hierarchical control in [20] and distributed control in [56] to compensate the bus voltage deviation and inaccurate current sharing caused by line impedance.

3.4.2 System expansion with multiple energy storage devices

The proposed strategy can be extended to a hybrid system with multiple energy storage devices. When multiple batteries are integrated, their virtual resistances are determined by their respective power ratings as expressed in (3.3). For multiple integrated SCs, their transient power sharing can be designed in proportional to their capacities $Q_{c,j}$, i.e.

$$C_{v,1} : C_{v,2} : \dots : C_{v,n} = Q_{c,1} : Q_{c,2} : \dots : Q_{c,n} \quad (3.30)$$

where $C_{v,j}$ and $Q_{c,j}$ are virtual capacitance and capacity of j th SC.

Then the system can be simplified to the battery/SC system in Fig. 3.2 with equivalent virtual resistance and capacitance as

$$\begin{cases} \frac{1}{R_v} = \sum_{i=1}^m \frac{1}{R_{v,i}} \\ C_v = \sum_{j=1}^n C_{v,j} \end{cases} \quad (3.31)$$

where m and n are the number of batteries and SCs; $R_{v,i}$ is the virtual resistance of i th battery.

Similar to the extended droop scheme in Chapter 2, the proposed approach is not limited for battery/SC system and can be applied for hybrid system with different characteristics such as fuel cell/battery system or fuel cell/SC system that VRD and voltage restoration loop regulate energy sources with slow dynamics and VCD controller regulate energy sources with fast dynamics.

3.5 System Studies

3.5.1 Simulation results

The HESS with control architecture in Fig. 3.4 is constructed in PLECS to validate the effective of proposed strategy. Parameters are listed in Table 3.1.

CHAPTER 3. A DECENTRALIZED POWER MANAGEMENT STRATEGY FOR BUS VOLTAGE RESTORATION AND SOC RECOVERY

Table 3.1: System Parameters.

Variable	Description	Value
V_{nom}	DC bus nominal voltage	270 V
V_{SC}	SC nominal voltage	200 V
V_B	Battery nominal voltage	100 V
Q_{SC}	SC energy capacity	20 Wh
Q_B	Battery energy capacity	16 kWh
P_B	Battery converter power rating	1Ω
f_s	Switching frequency	10 kHz
L_B, L_{SC}	Inductors of battery and SC converters	0.002H
C_{oB}, C_{oSC}	Output capacitors of battery and SC converters	220 μF, 1100 μF
$\omega_{vB}, \omega_{vSC}$	Voltage control bandwidths of battery and SC	$2\pi \times 100$, $2\pi \times 20$
ω_{iB}, ω_i	Current control bandwidths of battery and SC	$2\pi \times 1000$ Hz
SoC_{ini}	Initial SoC of SC	0.5
R_v	Virtual resistance	1 Ω

Table 3.2: Parameters for Different Cases.

Case	Expected cutoff frequency (ω_c)	Designed damping (ξ_c)	VRD (R_v)	VCD (C_v)	Voltage regulation loop gain(k_i)
1	$0.4 \times 2\pi$	1	1	0.3979	0
2	$0.4 \times 2\pi$	0.8	1	0.5431	0.7192
3	$0.2 \times 2\pi$	0.8	1	1.0862	0.3596
4	$0.4 \times 2\pi$	1.2	1	0.4656	0.3729
5	$0.4 \times 2\pi$	0.8	1; 0.5	0.181; 0.362	0.7192

Cases in Table 3.1 are investigated with different expected cutoff frequencies and damping ratios. The same load change is applied for Case 1-4 that the load steps up from 0.1kW to 1.1kW at 10s and steps down from 1.1kW to -0.9kW at 22.5s. I_{oSC} and I_{oB} refer to the output currents of the SC and battery converters respectively.

Fig. 3.7 shows the effect of the autonomous power split for Case 1 with the implementation of VCD and VRD controllers without the SoC regulation of the battery and SC. At the time of load change, SC responds immediately and battery responds gradually; at steady state, SC will not supply power and the battery supplies total demand. Thus autonomous power split for the battery and SC is achieved whereas the issues of the bus voltage regulation and SoC deviation are unsolved. It can also be observed from the SoC variations of the battery and SC that, with the change of the HESS equivalent load, the maximum deviation

Figure 3.7: Simulation results for Case 1.

Figure 3.8: Simulation results for Case 2.

of the battery SoC to its nominal value is 0.0006, while the maximum deviation of the SC SoC is 0.2. This verifies the discussion in Section 3.2.3 that the SoC

Figure 3.9: Simulation results for Case 3.

variation of the battery is much lower than that of SC in real-time operation and the battery can be assumed to operate within its SoC limitation in this thesis.

Fig. 3.8 demonstrates the system response with the proposed strategy for Case 2. The VHPFD and VCD loops are implemented with the designed parameters to minimize the interactions. With the same expected cutoff frequency, the effective power split is achieved and transient dynamics of the battery and SC are almost the same as that in Fig. 3.7 by observing the dynamics at 10s-11s and 22.5s-23.5s. At the end of the transient power sharing process, the voltage/SoC restoration process becomes dominant and then the bus voltage is restored along with the recovery of SC SoC. Therefore, the effectiveness of the proposed strategy is verified.

The effectiveness of the design guideline is examined by the comparison of simulation results in Fig. 3.9 for Case 3 and Fig. 3.8 for Case 2. As can be observed in Fig. 3.9, autonomous power split, voltage regulation and SoC recovery are also achieved. The time of transient power sharing is around 2.4s, which is twice as that in Fig.8. The result is coherent with the relationship of their expected cutoff frequencies, which validates the effectiveness of the design guideline.

The impact of damping ratio is investigated by comparing simulation results in Fig. 3.10 for Case 4 and Fig. 3.8 for Case 2. As the expected cutoff frequency is the same, same transient power sharing between battery and SC is achieved, which can be verified by dynamic response at 10s-11s or 22.5s-23.5s. The bus voltage/SC SoC restoration time for Case 2 and Case 4 are around 3s and 6s, respectively. With a higher damping ratio, the restoration time becomes longer

Figure 3.10: Simulation results for Case 4.

Figure 3.11: Simulation results for Case 2 with the original value of SC SoC at 0.4.

and this result is in agreement with the analysis for Fig. 3.6.

Fig. 3.11 shows the simulation result for Case 2 with a different initial SC SoC value set at 0.4. As analyzed in Section 3.2.3 that the SC SoC is restored to its initial value along with the DC voltage restoration. The simulation results in Fig. 3.11 and Fig. 3.8 validate that after load variations, the SC SoC can always return to its original value and the dynamics of its SoC recovery is not affected by the original value of the SC SoC. Therefore, once the original value of the SC SoC is set at the reference value, SC can always operate around its reference SoC condition.

(a) $R_{l1}=0.2 \Omega$, $R_{l2}=0 \Omega$

(b) $R_{l1}=0 \Omega$, $R_{l2}=0.2 \Omega$

Figure 3.12: Simulation results for Case 2 with line impedances between output of converters and DC bus.

The impact of line impedances on the proposed method is investigated in Fig. 3.12. The line impedance between the battery system and DC bus is denoted as R_{l1} and the line impedance between SC system and DC bus is denoted as R_{l2} . The impact of R_{l1} is shown in Fig. 3.12a with R_{l1} set at 0.2Ω , and R_{l2} at 0Ω . Other parameters are the same as that in Fig. 3.8 for Case 2. As can be observed, the transient power sharing between the battery and SC is still achieved, whereas there are small deviations at the bus voltage and SC SoC to their respective reference values due to the voltage drop caused by the line impedance between battery and DC bus. But compared with the existing VRD and VCD method in

Figure 3.13: Simulation results for Case 5 with two batteries and two SCs.

Figure 3.14: Simulation results for Case 2 with a series of pulse loads.

Fig. 3.7 without the voltage regulation, these deviations are relatively small. To minimize these deviations, the line impedance between the output terminal of the

battery converter and DC bus R_{l1} should be as small as possible i.e. $R_{l1} \approx 0$, and it is recommended that the battery system should be located close to the DC bus. Fig. 3.12b presents the impact of impedance R_{l2} by setting R_{l1} at 0Ω , and R_{l2} at 0.2Ω . It can be observed that the transient sharing between the battery and SC, DC bus voltage restoration and SC SoC recovery are all achieved, whereas the system dynamic response is slighter slower compared to that in Fig. 3.8 without the line impedance. This slower dynamic response can be compensated by tuning the virtual capacitance value to reach the same natural frequency and thus similar dynamic result, as that in Fig. 3.8. Therefore, the proposed method is still effective when the line impedances are considered.

The proposed strategy for the system with multiple energy storage devices is studied in Fig. 3.13 with the parameters in Case 5 for two batteries and two SCs. A battery with twice power capacity of P_B and an SC with twice energy capacity of Q_{SC} are integrated into the existing battery/SC system. The load steps up from 0.2kW to 2.2kW at 10s and steps down from 2.2kW to -1.8kW at 22.5s . As can be observed, the high frequency power is compensated by SCs and the low frequency power is supplied by batteries with the desired cutoff frequency. Steady state power sharing between batteries are based on their virtual resistances (1:2). Transient power sharing between SCs are based on their virtual capacitances (1:2). The voltage restoration and SoC regulation are also achieved as expected. SoC recovery process of two SCs are almost the same. Moreover, as the expected frequency and damping ratio are the same as those in Case 2, the system dynamics for the power sharing and voltage regulation are almost the same with that in Fig. 3.8. This validates the application of the proposed strategy in the system with multiple energy storage devices.

Considering the different fluctuation frequencies of the equivalent load (the residual power of RESs and load) in practical operation, a series of pulse loads with different frequencies are investigated in Fig. 3.14 with the parameters in Case 2. The pulse frequency is set at 0.5Hz , 0.2Hz and 0.1Hz during $0\text{s} - 10\text{s}$, $10\text{s} - 20\text{s}$ and $20\text{s} - 35\text{s}$, respectively. In each cycle, the load steps up from 0W to 500W and steps down from 500W to 0W . As can be observed, with each step change, SC responses immediately and the battery responses gradually; the bus voltage value and SC SoC value are regulated around their reference values, respectively. This further verifies the effectiveness of the proposed strategy.

To verify the proposed approach in a more realistic situation, a stochastic time domain simulation is performed in Fig. 3.15 with the fluctuations of the

Figure 3.15: Simulation results for Case 2 with the variation of the renewable power generation.

power output of RESs. The renewable power is modeled as a series of uniformly distributed random numbers between 0W and 1000W at the sampling time of 1s. The load is fixed at 500W. The battery/SC system is to compensate the residual power of RESs and load. The control parameters in Case 2 is adopted. It can be observed from Fig. 3.15 that, with the variations of the renewable power, SC responds fast and the battery provides smooth power; the bus voltage value and SC SoC value are regulated around their reference values respectively. The result is in consistent with previous results and the effectiveness of the proposed approach is validated.

3.5.2 Experimental results

An experiment platform for HESS is constructed to validate the proposed strategy, as shown in Fig. 3.16. The SC is represented by a Li-ion battery bank of 195V and the battery is represented by a lead-acid battery bank of 95V. The interface bidirectional DC/DC converters are built by IGBT modules with TI TMS320F28335 DSPs, respectively. A resistive load is adopted to emulate the

Figure 3.16: Experiment Setup.

Figure 3.17: Experiment results for Case 1.

load variation. The same load change is employed for the cases with a step up from 270W to 800W and a step down from 800W to 270W. The SoC value is obtained by the math function in the LeCroy oscilloscope based on (3.11).

Fig. 3.17 shows the experiment result for Case 1 when the VRD and VCD strategies are implemented. As can be observed, the transient power sharing between the battery and SC is achieved, whereas steady state deviations of the bus voltage and SC SoC are unsolved.

Fig. 3.18 illustrates the effectiveness of the proposed strategy with the designed parameters in Case 2. When the load steps up, I_{oSC} increases immediately and then decreases slowly along with the gradual increase of I_{oB} . The transient

Figure 3.18: Experiment results for Case 2.

Figure 3.19: Experiment results for Case 3.

power sharing between the battery and SC is almost the same as that in Fig. 3.17 under the same expected cutoff frequency. At the end of the transient dynamics, the voltage/SoC restoration process starts to take the role. I_{oSC} continues to reduce to a negative value and the battery starts to charge SC. Finally, I_{oSC} returns to zero and battery provides the total load current at steady state. In the meantime, the voltage restoration and SC SoC recovery are completed. Similar dynamics can be observed when the load steps down. Therefore, the proposed

strategy can achieve the autonomous power split, voltage restoration and SoC recovery.

The comparison of Fig. 3.19 for Case 3 and Fig. 3.18 for Case 2 verify the effectiveness of the design guideline. For both cases, the desired objectives are achieved. The transient time of the autonomous power split for the battery and SC in Fig. 3.19 is twice as that in Fig. 3.18, which meets the relationship of their expected cutoff frequencies. Therefore, with the developed design guideline, the proposed strategy can achieve the desired dynamic performance.

3.6 Conclusion

In this chapter, a decentralized strategy is proposed for HESS to achieve the autonomous power split, voltage regulation and SoC recovery. An VHPFD controller is proposed for the battery converter which is equivalent to adding an HPF to the conventional VRD. The proposed VHPFD automatically eliminates the output voltage deviation. A VCD controller is implemented for the SC converter to ensure zero current sharing of SC at steady state. The cooperation of VHPFD and VCD strategies enables simultaneous bus voltage restoration and SC SoC recovery. A design guideline is developed to obtain the expected transient power sharing dynamics with negligible interactions with the voltage/SoC restoration process. Impact of line impedance is investigated and the scalability of the proposed approach is illustrated. Simulations and experiments are conducted to validate the effectiveness of the proposed strategy and analytical results.

Chapter 4

A Modular Approach for Stability Analysis of Microgrids

In Chapter 2-3, the power balance issue is addressed with the proposed decentralized control strategies. Along with the precise control and system coordination, stable operation should always be ensured for MGs. Therefore, in this chapter and the next chapter, we will focus on system stability issue. Considering the increasing complexity as well as the requirement of the system extension and re-configuration of MGs, a modular approach is proposed in this chapter for effective system modeling and stability analysis.

4.1 Introduction

Stable operation is a basic requirement for the MGs and it should be always ensured during all the operating conditions [22]. In MGs, power electronic converters provide efficient interfaces for the integration of various sources and loads [42]. The extensive interconnection of power electronics converters, however, lead to a large variety of dynamic interactions and may cause unstable issues [45]. Power electronic converter loads, when tightly controlled, behave as constant power loads (CPLs) [43]. CPLs have negative impedance characteristics, which may destabilize system. Instability problems arise from the interactions when CPLs are interconnected with other subsystems, even though each subsystem is designed to meet stability requirements in its standalone mode [45,46]. Due to the increasing penetration of CPLs in the MGs, it is significantly important to understand the interactions and evaluate system stability.

In the past, extensive simulations under all kinds of operating conditions are necessary to ensure system stable operation [66]. Considering drawbacks of real-time simulations in terms of complexity, cost and time, averaging models have been developed for stability analysis using averaging modeling methods, including the state space averaging (SSA) method [67, 68] and generalized state space averaging (GSSA) approach [69]. Based on the models built by SSA or GSSA approach, in order to do stability analysis, the system model should be integrated and then linearized into the state space form with the system matrix. The eigenvalue method is then employed to examine eigenvalues of system matrix and assess system stability. With the increasing number of loads integrated in the MGs, the system model becomes more complex. Moreover, system reconfiguration is required for different failure contingencies and operating conditions. Within the existing approaches, the system model has to be reformulated and re-linearized for each contingency or operating condition. Though modeling approaches for standalone components have been reported in many literatures, there is no general approach for system level modeling and integration. Considering system complexity and the requirement of extension and reconfiguration, a general and flexible approach for system integration should be explored.

The Middlebrook methodology and its various extensions provide another approach for stability analysis which are based on impedance models [45, 70]. It has the advantages of modularity and flexibility. However, it is a conservative method and how to select the proper interface to divide source/load system may also affect stability analysis results, especially for the complex system with multiple buses. In addition, impedances of some components are difficult to derive and some existing impedance models are not easily applicable for AC distribution systems with parallel connected multiple power converters.

This chapter proposes a modular approach, which is a continuation of the eigenvalue method and also possesses the advantages of modularity and flexibility as Middlebrook methodology. A general module is defined as a state-space matrix with interconnection vectors in order to model the dynamic characteristics of a general system component. Modules can be connected in series or parallel by matching the interconnection vectors of modules. By this means, the system matrix can be easily integrated based on system configuration. To investigate system stability under different operating conditions and failure contingencies, the system matrix can be modified by adding, removing or replacing relevant modules. To achieve different accuracy of system stability analysis, different component mod-

els such as SSA, GSSA or reduced order models can be used to form component modules. The proposed approach also allows seamless integration of an increasing number of loads in MGs and provides a flexible solution for system extension and reconfiguration.

Recently, the aircraft industry is moving toward more electric aircraft (MEA) [71, 72]. As a typical complex autonomous MG, there are extensive interconnections of power electronic converters in the MEA power system and its stability is of great concern [73, 74]. Therefore, in the second part of this chapter, the proposed modular approach is applied in the modeling and stability analysis of the hybrid AC/DC MEA power system, as an application example.

This chapter is organized as follows: First, a modular approach is proposed with the formulation of different connection types. Next, the proposed approach is implemented in the hybrid AC/DC MEA power system as an application example. The complex hybrid AC/DC MEA power system is presented with models of each component and the whole system is integrated with the proposed approach. Then stability analysis is performed for the complex MEA power system with the verification of simulation results.

4.2 The Proposed Modular Approach

In the conventional eigenvalue stability analysis, the system dynamic model is established by integrating all component models and then linearized to obtain the system matrix for eigenvalue evaluation. The procedure of system modeling and linearization is tedious and time consuming for complex system reconfiguration. To simplify modeling process, a general dynamic module is proposed to represent dynamics of system components by dividing system into modules. A general dynamic module consists of a module matrix, which is derived from a small signal model to show dynamic characteristics of a component, and interconnection vectors, which show the connections with other modules. The model of each module can be obtained by various methods such as SSA method, GSSA approach and etc. Then the system matrix can be obtained by the interconnection of component module matrices.

Each module is derived from its dynamic model in the form of

$$\dot{x} = f(x, u) \quad (4.1)$$

Figure 4.1: The proposed module.

Fig. 4.1 shows a general module ij with its interconnection vectors $y_{ij,u}$, $y_{ij,d}$, $y_{ijU,d}$, $y_{ijD,u}$. The module is connected between node i and node j and its state vector is denoted as x_{ij} . The interconnection vectors represent the information exchange between modules, which are usually current or voltage vectors related to state vectors. The bold line represents the physical connections and the dashed arrows represent the exchanged information vectors. For module ij with its state vector x_{ij} , it provides $y_{ij,u}$ and $y_{ij,d}$ to its up modules ijU and down modules ijD respectively; and it obtains $y_{ijU,d}$ and $y_{ijD,u}$ from its up modules ijU and down modules ijD respectively. The subscripts u and d denote information flow directions, up and down. For a module, its up modules and down modules can be more than one, which will be discussed in the case of the parallel connections later.

By performing small signal disturbance, the small signal model of module ij with interconnection vectors can be expressed as

$$\Delta \dot{x}_{ij} = A_{ij} \Delta x_{ij} + \sum B_{ij,ijU} \Delta y_{ijU,d} + \sum B_{ij,ijD} \Delta y_{ijD,u} \quad (4.2)$$

$$\Delta y_{ij,u} = C_{iju} \Delta x_{ij} \quad (4.3)$$

$$\Delta y_{ij,d} = C_{ijd} \Delta x_{ij} \quad (4.4)$$

where all the matrices are partial derivatives obtained from the dynamic model at the operating point, i.e. $A_{ij} = \partial f / \partial x_{ij}$, $B_{ij,ijU} = \partial f / \partial y_{ijU,d}$, $B_{ij,ijD} = \partial f / \partial y_{ijD,u}$, $C_{iju} = \partial y_{ij,u} / \partial x_{ij}$, $C_{ijd} = \partial y_{ij,d} / \partial x_{ij}$; the second term in (4.2) represents the impact of interconnection vectors from its up modules ijU ; the third term in (4.2) represents the impact of interconnection vectors from its down modules ijD .

There are three types of modules, i.e. source modules, load modules and intermediate modules. If module ij is a source module, node i can be seen as a reference node (ground) and $\Delta y_{ijU,d}$ can be seen as the input vector from the outside environment, e.g. speed reference and torque from engine. Similarly, if module ij is a load module, node j can be seen as a reference node (ground) and

Figure 4.2: Series connected modules.

Δy_{ijD_u} can be seen as the input vector from the outside disturbance, e.g. speed reference and torque from load machine. As these input vectors from outside are not related to the system state vectors, it will not be included for the system matrix integration. For intermediate modules, they will exchange information with both up module and down modules.

Modules can be connected in series or parallel to form the system. The procedure of the system model formulation for different module connections are presented below.

4.2.1 Series connected modules

Fig. 4.2 shows a system of series connected module ij and module jk . The small signal model of module jk is similar to module ij in (4.2)-(4.3) and is obtained by replacing the subscript ij with jk as

$$\Delta \dot{x}_{jk} = A_{jk} \Delta x_{jk} + \sum B_{jk,jkU} \Delta y_{jkU_d} \quad (4.5)$$

$$\Delta y_{jk_u} = C_{jku} \Delta x_{jk} \quad (4.6)$$

In this system, module ij is a source module and module jk is a load module, so Δy_{ijU_d} and Δy_{jkD_u} can be seen as input vector from the outside environment and thus be ignored. As module jk is the down module of ij and module ij is the up module of jk , the subscripts ijD and jkU are replaced by jk and ij , respectively, and the following relationships hold

$$\sum B_{ij,ijD} \Delta y_{ijD_u} = B_{ij,jk} \Delta y_{jk_u} \quad (4.7)$$

$$\sum B_{jk,jkU} \Delta y_{jkU_d} = B_{jk,ij} \Delta y_{ij_d} \quad (4.8)$$

Substitution of (4.6) and (4.7) into (4.2), (4.4) and (4.8) into (4.5), and com-

Figure 4.3: Parallel connected up modules.

bination of (4.2) and (4.5) give the system model as

$$\begin{bmatrix} \Delta \dot{x}_{ij} \\ \Delta \dot{x}_{jk} \end{bmatrix} = \begin{bmatrix} A_{ij} & B_{ij,jk}C_{jku} \\ B_{jk,ij}C_{ijd} & A_{jk} \end{bmatrix} \begin{bmatrix} \Delta x_{ij} \\ \Delta x_{jk} \end{bmatrix} \quad (4.9)$$

4.2.2 Parallel connected up modules

Fig. 4.3 shows a two-source system in order to illustrate the integration situation when a module has several up modules at one node. In this case, module ij and module ilj are source modules parallel connected at node j and module jk is the down module. The small signal model of module ilj can be obtained similar to the expressions in (4.2) and (4.4) by replacing the subscript ij with ilj .

$$\Delta \dot{x}_{ilj} = A_{ilj} \Delta x_{ilj} + \sum B_{ilj,iljD} \Delta y_{iljD-u} \quad (4.10)$$

$$\Delta y_{ilj-d} = C_{iljd} \Delta x_{ilj} \quad (4.11)$$

Based on the topology, the subscript jkU represents module ij and ilj , which yields the relationship in (4.12); the subscripts ijD and $iljD$ are substituted by jk , which gives the relationship in (4.7) and (4.13).

$$\sum B_{jk,jkU} \Delta y_{jkU-d} = B_{jk,ij} \Delta y_{ij-d} + B_{jk,ilj} \Delta y_{ilj-d} \quad (4.12)$$

$$\sum B_{ilj,iljD} \Delta y_{iljD-u} = B_{ilj,jk} \Delta y_{jk-u} \quad (4.13)$$

Substitution of (4.4) and (4.8) into (4.2), (4.11) and (4.12) into (4.5), (4.6) and (4.13) into (4.10), and combination of (4.2), (4.5) and (4.10) give the system

Figure 4.4: Parallel connected down modules.

model as

$$\begin{bmatrix} \Delta \dot{x}_{ij} \\ \Delta \dot{x}_{jk} \\ \Delta \dot{x}_{i1j} \end{bmatrix} = \begin{bmatrix} A_{ij} & B_{ij,jk}C_{jku} & 0 \\ B_{jk,ij}C_{ijd} & A_{jk} & B_{jk,ij1}C_{i1jd} \\ 0 & B_{i1j,jk}C_{jku} & A_{i1j} \end{bmatrix} \begin{bmatrix} \Delta x_{ij} \\ \Delta x_{jk} \\ \Delta x_{i1j} \end{bmatrix} \quad (4.14)$$

4.2.3 Parallel connected down modules

To illustrate the case when a module has several down modules connected at one node, a two-load system in Fig. 4.4 is modeled as an example. Module ij is the source module and module jk and $jk1$ are parallel connected load modules. The small signal model of module $jk1$ can be obtained similar to the expressions in (4.2)-(4.3) by replacing the subscript ij with $jk1$

$$\Delta \dot{x}_{jk1} = A_{jk1} \Delta x_{jk1} + \sum B_{jk1,jk1U} \Delta y_{jk1U_d} \quad (4.15)$$

$$\Delta y_{jk1_u} = C_{jk1u} \Delta x_{jk1} \quad (4.16)$$

As module ij is the up module of module jk and $jk1$, the subscript ijD consists of module jk and $jk1$ and the relationship in (4.17) is obtained; module jk and $jk1$ are down modules of module ij , i.e. the subscripts jkU and $jk1U$ are equivalent to ij and the interconnection relationship is matched as (4.8) and (4.18).

$$\sum B_{ij,ijD} \Delta y_{ijD_u} = B_{ij,jk} \Delta y_{jk_u} + B_{ij,jk1} \Delta y_{jk1_u} \quad (4.17)$$

$$\sum B_{jk1,jk1U} \Delta y_{jk1U_d} = B_{jk1,ij} \Delta y_{ij_d} \quad (4.18)$$

Substitution of (4.6), (4.16) and (4.17) into (4.2), (4.6) and (4.8) into (4.5), (4.4) and (4.18) into (4.15), and combination of (4.2), (4.5) and (4.15) give the system model as

$$\begin{bmatrix} \Delta \dot{x}_{ij} \\ \Delta \dot{x}_{jk} \\ \Delta \dot{x}_{jk1} \end{bmatrix} = \begin{bmatrix} A_{ij} & B_{ij,jk}C_{jku} & B_{ij,jk1}C_{jk1u} \\ B_{jk,ij}C_{ijd} & A_{jk} & 0 \\ B_{jk1,ij}C_{ijd} & 0 & A_{jk1} \end{bmatrix} \begin{bmatrix} \Delta x_{ij} \\ \Delta x_{jk} \\ \Delta x_{jk1} \end{bmatrix} \quad (4.19)$$

4.2.4 Matrix characteristics and system reconfiguration

Based on the system matrix formulation of different types of module connections, some characteristics of the system model can be observed from (4.9), (4.14) and (4.19), which help to simplify the process of system integration. The m^{th} block in the diagonal line in system matrix is the A_m matrix of module m . If module m and n are interconnected up and down modules at one node (eg. module ij and module jk are interconnected at node j), then the $(m,n)^{th}$ block and $(n,m)^{th}$ block in the system matrix should be $B_{m,n}C_{nu}$ and $B_{n,m}C_{md}$, respectively; otherwise, the $(m,n)^{th}$ block and $(n,m)^{th}$ block in the system matrix will be zero matrix. This characteristic can help to integrate system matrix once the interconnected module/node relationship is obtained from the system topology. Moreover, this characteristic also indicates that the system matrix can reflect the interconnection relationship of modules and nodes, which in turn provides a verification way for the integrated system matrix.

Matrices for the small signal model of each component with its interconnection vectors are extracted individually and can be reused in different scenarios and different systems with only modification of parameters as long as components remain the same. Each module can be modeled with sufficient accuracy and reduced order models are not necessary for simplification. Moreover, if the system reconfiguration occurs, system matrices can be obtained by just adding, removing or modifying matrices associated with the relevant modules. For example, the two-load system in Fig. 4.4 can be seen as adding the module $jk1$ at node j as another down module of module ij in the simple series system in Fig. 4.2. Then based on the matrix in (4.9) for the system in Fig. 4.2, the system matrix of paralleled system can be obtained by adding matrices of module $jk1$, as expressed in (4.19). If module jk in Fig. 4.4 is replaced by another load $jk2$, only sub-matrices with subscript of jk should be replaced by the new ones with subscript of $jk2$ in the system matrix, as expressed in (4.20). If module jk is

shed because of emergency events, system matrix can be easily reformulated as (4.21) by removing sub-matrices related with module jk (setting them as zero sub-matrices). If another load module $jk3$ is added at node j as another down module of module ij , the system matrix can be conveniently obtained as (4.22) by adding sub-matrices related to module $jk3$.

$$\begin{bmatrix} A_{ij} & B_{ij,jk2}C_{jk2u} & B_{ij,jk1}C_{jk1u} \\ B_{jk2,ij}C_{ijd} & A_{jk2} & 0 \\ B_{jk1,ij}C_{ijd} & 0 & A_{jk1} \end{bmatrix} \quad (4.20)$$

$$\begin{bmatrix} A_{ij} & 0 & B_{ij,jk1}C_{jk1u} \\ 0 & 0 & 0 \\ B_{jk1,ij}C_{ijd} & 0 & A_{jk1} \end{bmatrix} \quad (4.21)$$

$$\begin{bmatrix} A_{ij} & B_{ij,jk}C_{jku} & B_{ij,jk1}C_{jk1u} & B_{ij,jk3}C_{jk3u} \\ B_{jk,ij}C_{ijd} & A_{jk} & 0 & 0 \\ B_{jk1,ij}C_{ijd} & 0 & A_{jk1} & 0 \\ B_{jk3,ij}C_{ijd} & 0 & 0 & A_{jk3} \end{bmatrix} \quad (4.22)$$

This method can be applied to interconnected systems with any modules. For a given system topology, it can be divided by modules with the interconnected nodes and the order of the nodes can be determined by the depth-first search [75]. For each module, we have the information of its node orders and matrices derived from its dynamic model. By writing a programming to visit all the nodes of modules in the system, we can form the up module and down module relationships. The operating point can be obtained by the steady state power flow calculation. Then the system dynamic matrix can be integrated by the proposed approach according to the system topology, and the operating point can be substituted into the system matrix for the eigenvalue stability analysis. For the case of system reconfiguration, only the modules or the connections of the nodes that differ from the previous system need to be considered and the system matrix modification can be performed similar to the process in (4.20), (4.21) and (4.22). The operating point under the new configuration can be recalculated by power flow analysis. Then the variables related to the operating point in the system matrix will be modified according to the new operating point for the eigenvalue analysis.

4.2.5 Limitations

The proposed approach is a linear technique and could not ensure system stability at large disturbances, which may be caused by faults, protection, sudden load connections or disconnections, etc. For the large signal stability, normally simulation-based approaches are employed. But these methods are time consuming and may need additional resources for complex systems. Some analytical methods to investigate large signal stability have been proposed in literatures and there is a detailed review on these methods in [76]. However, these large signal analysis methods are only applicable for simple systems, such as inverter-fed motor drive system with DC power supply. They are not able to handle the complex aircraft power system with multiple power converters at the current stage of development. Therefore, a large signal stability analysis will be left as our future work.

The proposed approach will be applied and illustrated in the next section to obtain system matrices for stability analysis of the complex MEA power system.

4.3 The Application Example in the More Electric Aircraft Power System

In this part, an overview of the MEA technology and the state of art of the MEA system are presented first. Next, the complex hybrid AC/DC power system is presented with models of each component and then, the whole system is integrated with the proposed method. Finally, stability analysis is performed for the complex MEA power system with the simulation verification.

4.3.1 An overview of the More Electric Aircraft

Driven by the demand to reduce gas emissions and decrease operational cost as well as to provide safe and comfortable services, the aircraft industry is moving towards more electric aircraft (MEA) [71, 73]. MEA is to use electrical power to drive onboard loads, which are conventionally driven by a combination of hydraulic, pneumatic, electrical and mechanical power. In a conventional aircraft, electrical loads only take about 10% of the total non-propulsive loads. With the development of MEA technology, a large amount of electrical loads are integrated in the MEA power system, which leads to the increased capacity and complexity of electrical network [72, 77]. Therefore, advanced power network architecture is

required to ensure high reliability, stability and efficiency as well as light weight of MEA [72, 73]. For example, the hybrid AC/DC architecture with 230V AC/360-800Hz and ± 270 V DC is adopted to the latest MEA platform Boeing B787 [78, 79].

The MEA power system is a typical isolated MG with the extensive integration of CPLs [74]. It requires greater use of power electronic converters and motor drive systems for functions as flight control actuation, fuel pump, environmental control system, engine start, etc. As mentioned in the background part in Chapter 1, the negative impedance characteristics of the CPLs may destabilize system operation. Due to the extensive integration of CPLs in MEA, it is significantly important to understand their interactions and evaluate system stability.

Stability analysis for the MEA power system has been reported in many literatures [67–69, 74, 80–83]. But most of them only consider a simplified system with one load supplied by the AC bus or DC bus. A complex hybrid AC/DC system normally consists of different AC/DC voltage levels i.e. 230V/380Hz-800Hz AC, ± 270 V DC, 115V/400Hz AC and 28V DC [78–80]. The interactions for such a complex system has not been reported in the literature. To fully understand the interactions of the MEA power system and guarantee its system stability, the loads at each bus should be considered and the whole system should be investigated with the variation of system parameters and operating modes.

4.3.2 System modeling under the proposed approach

The more electric aircraft power system under study is shown in Fig. 4.5 to illustrate the modeling technique of a general MEA power system [74, 80]. This structure is a hybrid high voltage alternating current (HVAC) and high voltage directing current (HVDC) structure adopted by Boeing 787. The primary 230V HVAC bus is powered by a variable frequency synchronous generator. Frequency insensitive loads such as de-icing system and essential loads are directly connected to the HVAC bus. Electromechanical actuators (EMAs), which are used for flight control actuation, are connected to the HVAC bus through a PWM rectifier. The 540V HVDC bus is connected to HVAC bus through an autotransformer rectifier unit (ATRU). The environmental control system (ECS), which is the largest power consumers in MEA, is connected to the HVDC bus. To power various 115V/400Hz AC (LVAC) and 28V DC (LVDC) legacy loads, the conventional 115V/400Hz AC and 28V DC distributions are preserved in this architecture, which is converted from the HVAC bus by using a two-stage inversion power supply and a transformer

Figure 4.5: The MEA power system under study.

Table 4.1: Interconnection vectors for each module

Module	Module type	Vector to up module	Vector to down module	Vector from up module	Vector from down module
SG	Source	-	y_{sg_d}	-	y_{sgD_u}
ATRU	Intermediate	y_{a_u}	y_{a_d}	y_{aU_d}	y_{aD_u}
Rectifier	Intermediate	y_{r_u}	y_{r_d}	y_{rU_d}	y_{rD_u}
EMA/ECS	Load	y_{m_u}	-	y_{mU_d}	-
Inverter	Load	y_{inv_u}	-	y_{invU_d}	-
TRU	Load	y_{t_u}	-	y_{tU_d}	-
Line	Intermediate	y_{fl_u}	y_{fl_d}	y_{flU_d}	y_{flD_u}

rectifier unit (TRU) respectively. Considering the impact of the long transmission line in the distribution network, the line impedance between SG and HVAC bus is also considered as an example. As various loads are connected to the distribution network through different converters, the whole system stability should be investigated by considering the interactions of subsystems and destabilizing effects of CPLs, in order to ensure stable operation of the MEA power system.

The system shown in Fig. 4.5 is modeled with the proposed technique described in Section 4.2, which consists of SG, ATRU, ECS, PWM rectifier, EMA, two-stage inversion subsystem, TRU and transmission line. Each component is seen as a module. Table 4.1 summarizes the module type and interconnection vectors for each module. The dynamic model and the corresponding small signal model of each module are derived and then the system matrix is integrated for stability analysis. The s-domain circuit diagrams and state space models of the modules are presented in Appendix A.

Generator module

A wound field brushless synchronous generator (SG) model is adopted and the dynamics of stator, field and two damper windings are considered. The stator variables are aligned with rotor reference frame using Park transformation and the whole model is built in dq reference frame. A double loop PI controller is applied for the generator control unit (GCU) that the outer voltage loop regulates SG output voltage at 230V and fast dynamic inner current loop generates field voltage for the SG. State variables consist of the d-axis stator current I_{sd} , field current I_F , d-axis damper winding current I_{kd} , q-axis stator current I_{sq} , voltage loop state X_e , current loop state X_i , d-axis output voltage V_{sd} and q-axis output voltage V_{sq} . The dynamic model of SG with GCU can be referred to [84], and the equivalent circuit and the corresponding state space model are shown in Appendix A.1.1. As SG is a source module, only interconnection with its down module is considered, as shown in Table 4.1. The linearized small signal model of SG with its interconnection vectors is expressed as

$$\Delta \dot{x}_{sg} = A_{sg} \Delta x_{sg} + B_{sg,sgD} \Delta y_{sgD-u} \quad (4.23)$$

$$\Delta y_{sg-d} = C_{sgd} \Delta x_{sg} \quad (4.24)$$

where

$$\Delta x_{sg} = \begin{bmatrix} \Delta I_{sd} & \Delta I_F & \Delta I_{kd} & \Delta I_{sq} & \Delta I_{kq} & \Delta X_e & \Delta X_i & \Delta V_{sd} & \Delta V_{sq} \end{bmatrix}_{1 \times 9}^T$$

$$\Delta y_{sg-d} = \begin{bmatrix} \Delta V_{sd} & \Delta V_{sq} \end{bmatrix}_{1 \times 2}^T$$

ATRU module

A 12-pulse ATRU is adopted to convert wild frequency 230V AC voltage to 540V DC voltage, which is modeled as an equivalent 6-pulse rectifier with 15° phase shift and increased magnitude by 1.035 [85]. The impedance between the source and rectifier consists of the transmission line impedance and transformer equivalent impedance, which will cause rectifier commutation effect. The commutation effect is represented by a voltage drop of a resistance r_μ on the DC side and then the rectifier is represented by a time-varying transformer in dq reference frame. The detailed modeling process can be referred to [86] and the dynamic model can be derived in Appendix A.1.2 with the equivalent circuit in Fig. A2. The state variables consist of the line current in d- and q-axis I_{ad} , I_{aq} , input terminal voltage rectifier V_{ad} , V_{aq} , dc bus voltage V_{dc} and current I_{dc} . It is an intermediate module and has interconnection vectors with the up module and down module. The linearized small signal model of ATRU with its interconnection vectors is expressed as

$$\Delta \dot{x}_a = A_a \Delta x_a + B_{a,aU} \Delta y_{aU.d} + B_{a,aD} \Delta y_{aD.u} \quad (4.25)$$

$$\Delta y_{a.u} = C_{au} \Delta x_a \quad (4.26)$$

$$\Delta y_{a.d} = C_{ad} \Delta x_a \quad (4.27)$$

where

$$\Delta x_a = \begin{bmatrix} \Delta I_{ad} & \Delta I_{aq} & \Delta V_{ad} & \Delta V_{aq} & \Delta I_{dc} & \Delta V_{dc} \end{bmatrix}_{1 \times 6}^T$$

$$\Delta y_{a.u} = \begin{bmatrix} \Delta I_{ad} & \Delta I_{aq} \end{bmatrix}_{1 \times 2}^T$$

$$\Delta y_{a.d} = [\Delta V_{dc}]_{1 \times 1}^T$$

PWM rectifier module

PWM rectifier is modeled in dq reference frame and the double-loop PI control is implemented. The outer voltage loop is used to regulate the output DC voltage and generate reference input current in d-axis I_{rd}^* . The inner current loop is used to regulate the input current to be in phase with the input voltage to achieve nearly unity power factor. The detailed modeling process can be referred to [87] and the dynamic model can be derived accordingly, as is shown in Appendix A.1.3 with equivalent circuit in Fig. A3. The state variables consist of the rectifier input current in d- and q- axis I_{rd} , I_{rq} , voltage control loop state X_{re} , d- and q-axis

current control loop state X_{rd} , X_{rq} , and dc bus voltage V_{dc1} . As the PWM rectifier is an intermediate module, interconnections with the up module and down module are included in the model. The linearized small signal model of PWM rectifier with its interconnection vectors is expressed as

$$\Delta \dot{x}_r = A_r \Delta x_r + B_{r,U} \Delta y_{rU_d} + B_{r,D} \Delta y_{rD_u} \quad (4.28)$$

$$\Delta y_{r_u} = C_{ru} \Delta x_r \quad (4.29)$$

$$\Delta y_{r_d} = C_{rd} \Delta x_r \quad (4.30)$$

where

$$\begin{aligned} \Delta x_r &= \begin{bmatrix} \Delta I_{rd} & \Delta I_{rq} & \Delta X_{re} & \Delta X_{rd} & \Delta X_{rq} & \Delta V_{dc1} \end{bmatrix}_{1 \times 6}^T \\ \Delta y_{r_u} &= \begin{bmatrix} \Delta I_{rd} & \Delta I_{rq} \end{bmatrix}_{1 \times 2}^T \\ \Delta y_{r_d} &= [\Delta V_{dc1}]_{1 \times 1}^T \end{aligned}$$

EMA/ECS module

EMAs connected to HVAC bus through PWM rectifiers are responsible for functions such as flight control actuation. They are lumped as one EMA. Environmental control system (ECS) connected to the HVDC bus is also driven by EMAs. To distinguish them, EMAs that drive the ECS are lumped as one EMA and denoted as ECS. EMA is a servo-motor drive system which can be modeled as an inverter fed induction machine. The motor is controlled using the well-known field oriented control method [88]. The detailed derivation process and dynamic model can be referred to [89]. The equivalent circuit of EMA and the corresponding state space model are illustrated in Appendix A.1.4. The state variables consist of the rotating frequency ω_r , q-axis stator current I_{sqm} , filtered DC-link voltage V_{f1} , q-axis reference stator voltage V_{sqm}^* and q-axis reference stator current I_{sqm}^* respectively. It is a load module and only exchange interconnection vectors with up module. The linearized small signal model of EMA with its interconnection vectors is expressed as

$$\Delta \dot{x}_m = A_m \Delta x_m + B_{m,U} \Delta y_{mU_d} \quad (4.31)$$

$$\Delta y_{m_u} = C_{mu} \Delta x_m \quad (4.32)$$

where

$$\Delta x_m = \begin{bmatrix} \Delta \omega_r & \Delta I_{sqm} & \Delta V_{f1} & \Delta V_{sqm}^* & \Delta I_{sqm}^* \end{bmatrix}_{1 \times 5}^T$$

$$\Delta y_{m.u} = [\Delta I_m]_{1 \times 1}^T$$

To distinguish ECS and EMA in Fig. 4.5, the subscript m represents the EMA module and the subscript $m1$ will be used to represent the ECS module.

Two-stage inversion subsystem

A two-stage inversion power supply is used to provide 115V/400Hz bus, which consists of an autotransformer unit (ATU), a PWM rectifier and an inverter. ATU is a step down transformer and can be modeled as a step ratio k . The PWM rectifier model is the same with the one in (4.28)-(4.30), to be distinguished from it, subscript $r1$ is used for the PWM rectifier in the two-stage power inversion subsystem. The inverter model is similar to the PWM rectifier model by reversing its input and output terminal. The double-loop PI control is implemented such that the outer voltage loop is used to regulate the inverter output voltage and the inner current loop is used to regulate generate modulation indices. As the LVAC load is dominantly resistive, the load is modeled as a resistor in the inverter model. The detailed modeling process can be referred to [90]. The corresponding state space model with the equivalent circuit is demonstrated in Appendix A.1.5. The state variables consist of the inverter output current and voltage in dq reference frame I_{ind} , I_{inq} and V_{ind} , V_{inq} , filtered DC link voltage V_{inf} , state of d- and q-axis voltage control loop X_{vd} and X_{vq} , state of d- and q-axis current control loop X_{id} and X_{iq} . The inverter with resistive load can be seen as a load module and as a result, only interconnections with up module are included. The linearized small signal model of the inverter with its interconnection vector is expressed as

$$\Delta \dot{x}_{inv} = A_{inv} \Delta x_{inv} + B_{inv,invU} \Delta y_{invU.d} \quad (4.33)$$

$$\Delta y_{inv.u} = C_{inu} \Delta x_{inv} \quad (4.34)$$

where

$$\Delta x_{inv} = \begin{bmatrix} \Delta I_{ind} & \Delta I_{inq} & \Delta V_{ind} & \Delta V_{inq} & V_{inf} & \Delta X_{vd} & \Delta X_{vq} & \Delta X_{id} & \Delta X_{iq} \end{bmatrix}_{1 \times 9}^T$$

$$\Delta y_{inv.u} = [\Delta I_{inv}]_{1 \times 1}^T$$

TRU module

The LVDC bus is connected to the HVAC bus through the transformer rectifier unit (TRU). The TRU model is similar with the ATRU model except that the transformer turn ratio for TRU is 1.0520 and for ATRU is 1.036. Therefore, the TRU model similar to the ATRU model in (4.25)-(4.27) by modifying the ratio value. To distinguish them, the subscript a is used to represent ATRU and the subscript t is used to represent TRU. The LVDC load is represented by a resistor, which can be modeled in TRU and this leads to removal of y_{tD-u} in the TRU model.

Line module

The transmission line can also be modeled as a module so that the system can be flexibly reformulated with/without the line impedances. The line module consists of the line resistance R_{fl} , line inductance L_{fl} and line capacitance C_{fl} . The LC filter at AC side can also share the same form as this module. The state variables are the inductor current in d- and q-axis I_{fld} , I_{flq} , and capacitor voltage in d- and q-axis V_{fld} , V_{flq} . The equivalent circuit and the corresponding state space model are shown in Appendix A.1.6.

The linearized small signal model of line module with its interconnection vectors is expressed as

$$\Delta \dot{x}_{fl} = A_{fl} \Delta x_{fl} + B_{fl,flD} \Delta y_{flD-u} + B_{fl,flU} \Delta y_{flU-d} \quad (4.35)$$

$$\Delta y_{fl-u} = C_{flu} \Delta x_{fl} \quad (4.36)$$

$$\Delta y_{fl-d} = C_{fld} \Delta x_{fl} \quad (4.37)$$

where

$$\begin{aligned} \Delta x_{fl} &= \begin{bmatrix} \Delta I_{fld} & \Delta I_{flq} & \Delta V_{fld} & \Delta V_{flq} \end{bmatrix}_{1 \times 4}^T \\ \Delta y_{flD-u} &= \begin{bmatrix} \Delta I_{fld} & \Delta I_{flq} \end{bmatrix}_{1 \times 2}^T \\ \Delta y_{flU-d} &= \begin{bmatrix} \Delta V_{flind} & \Delta V_{flinq} \end{bmatrix}_{1 \times 2}^T \end{aligned}$$

Figure 4.6: System integration under the proposed module-based method.

System integration

Based on the derived small signal model of each module with the interconnection vectors in Table 4.1 and according to the system architecture in Fig. 4.5, the system model can be integrated by matching the interconnection vectors of modules, as is shown in Fig. 4.6.

At HVAC interface, it can be obtained that

$$\Delta y_{flD_u} = \Delta y_{r_u} + \Delta y_{a_u} + \Delta y_{t_u} + k\Delta y_{r1_u} \quad (4.38)$$

$$\Delta y_{fl_d} = \Delta y_{rU_d} = \Delta y_{aU_d} = \frac{1}{k}\Delta y_{r1U_d} \quad (4.39)$$

Similarly, other equations based on interconnection matching can also be obtained from Fig. 4.6. By combining (4.23)-(4.38) and based on Fig. 4.6, the system matrix for the MEA power system under study is obtained as (4.40), with the state vector given by (4.41). It is a 56^{th} order system. The submatrices are shown in Appendix A.2. Then eigenvalue method can be employed to investigate

Figure 4.7: System eigenvalue plot at preset values.

stability of such a complex system.

$$A_{sys} = \begin{bmatrix} A_{sg} & B_{sg,fl}C_{flu} & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ B_{fl,sg}C_{sgd} & A_{fl} & B_{fl,r}C_{ru} & 0 & B_{sg,a}C_{au} & 0 & kB_{fl,r1}C_{r1u} & 0 & B_{fl,t}C_{tu} \\ 0 & B_{r,fl}C_{fld} & A_r & B_{r,m}C_{mu} & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & B_{m,r}C_{rd} & A_m & 0 & 0 & 0 & 0 & 0 \\ 0 & B_{a,fl}C_{fld} & 0 & 0 & A_a & B_{a,m1}C_{m1u} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & B_{m1,a}C_{ad} & A_{m1} & 0 & 0 & 0 \\ 0 & \frac{1}{k}B_{r1,fl}C_{fld} & 0 & 0 & 0 & 0 & A_{r1} & B_{r1,inv}C_{invu} & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & B_{inv,r1}C_{r1d} & A_{inv} & 0 \\ 0 & B_{t,fl}C_{fld} & 0 & 0 & 0 & 0 & 0 & 0 & A_t \end{bmatrix} \quad (4.40)$$

$$\Delta x_{sys} = [\Delta x_{sg} \quad \Delta x_{fl} \quad \Delta x_r \quad \Delta x_m \quad \Delta x_a \quad \Delta x_{m1} \quad \Delta x_{r1} \quad \Delta x_{inv} \quad \Delta x_t]_{1 \times 56}^T \quad (4.41)$$

4.3.3 Stability analysis and simulation verification

Based on the model developed in Section 4.3.2, the MEA system shown in Fig. 4.5 can be studied by examining eigenvalues of A_{sys} in (4.40). The system is stable if and only if all the eigenvalues are in the left side of the complex plane.

To investigate system stability and impacts of load at each bus, different operating scenarios are considered as shown in Table 4.2. The preset value at each bus is based on Boeing 787 data from [91]. The CPL at the HVDC bus is the ECS load and the CPL at the HVAC bus is the EMA load. The corresponding torques of ECS and EMA are 230 Nm and 350 Nm respectively at the speed of 975 r/min. The loads at the LVAC bus and LVDC bus are resistive. It should be mentioned that the total load will not exceed the generator capacity of 125kW and each component operates within its rated capacity so that the system instability will

Figure 4.8: Eigenvalue loci with load variation at each bus. (a) with variation of the HVDC CPL; (b) with variation of the HVAC CPL; (c) with variation of the LVAC load; (d) with variation of the LVDC load.

not be caused by the loss of power supply or power rating violation, but only the interactions. System parameters are shown in Appendix A.3.

Table 4.2: Operating scenarios

Scenarios	HVDC CPL	HVAC CPL	LVAC load	LVDC load
Preset value	45kW	28kW	21kW	4kW
Scenario I	1-60kW	14kW	10.5kW	2kW
Scenario II	22.5kW	1-90kW	10.5kW	2kW
Scenario III	22.5kW	14kW	1-87kW	2kW
Scenario IV	22.5kW	14kW	10.5kW	1-67kW

The eigenvalue plot for the integrated system at preset values is shown in Fig. 4.7. System is stable as all the eigenvalues are in the left side of the plane.

The impacts of the load at each bus are investigated in four scenarios with the variation of the load at that particular bus whereas other loads are fixed at half of the preset values. The eigenvalues with high negative real parts are not shown as they will not impact system stability. In Scenario I, when the load torque of the HVDC CPL (i.e. ECS) varies from 0 to 500Nm at a step increase of 10Nm, the dominant eigenvalues move toward the right half plane, which is shown in Fig. 4.8a. It can be observed that when the ECS load torque exceeds 370Nm, some eigenvalues move to the right half of the plane and the system becomes unstable. The corresponding HVDC bus power at the stability boundary is 45.8kW. In Scenario II, when the load torque of the HVAC dynamic CPL (i.e. EMA) varies from 10Nm to 640Nm at a step increase of 10Nm and the corresponding power varies from 2kW to 90kW, all eigenvalues stay in the left hand plane as shown in Fig. 4.8b and the system is stable. In Scenario III when the LVAC load varies from 1kW to 87 kW at a step increase of 1kW, the system is always stable as shown in Fig. 4.8c. In Scenario IV, when the LVDC load varies from 1kW to 67 kW at a step increase of 1kW, the system stays stable as shown in Fig. 4.8d. Based on the results of different operating scenarios, it turns out that the interactions on the HVAC bus, LVAC bus and LVDC bus will not make system unstable. It is the CPL on HVDC bus that can destabilize the system.

The hybrid MEA power system in Fig. 4.5 is built in MATLAB/Simulink to verify the analytical results. Fig. 4.9 shows the simulation results for Scenario I when only the HVDC dynamic CPL changes. Fig. 4.9a illustrates the dynamic response of the ECS torque load (T_{ECS}), angular frequency (ω_r), HVDC bus voltage (V_{HVDC}) and current (I_{HVDC}). Fig. 4.9b presents the dynamic response of power at each bus. At 2s, the ECS load steps up from 300Nm to 360Nm and the system is still stable. At 3.5s, when the ECS load torque changes from

(a)

(b)

Figure 4.9: Simulation results for Scenario I. (a) Dynamic response on HVDC bus; (b) Dynamic response of power at each bus.

Figure 4.10: Simulation results for Scenario II, III and IV.

Figure 4.11: Eigenvalue loci with the variation of system frequency.

360Nm to 370Nm and the corresponding power changes from 45.8kW to 47.3kW, the oscillation at the HVDC bus increases with time and finally the whole system becomes unstable. This validates the analytical result in Fig. 4.8a.

Fig. 4.10 shows the dynamic response of power at each bus to examine the analytical results for Scenario II, III and IV. At 1.5s, when the EMA load torque

Figure 4.12: Simulation results with different system frequencies.

increases from 500Nm to 635Nm and the corresponding EMA input power PEMA increases from 70kW to 90kW, the system is stable, which verifies the analytical result for Scenario II in Fig. 4.8b. At 3.5s, when the LVAC load increases from 43kW to 87kW, the system still operates well and this result validates the analytical result for Scenario III in Fig. 4.8c. The result in Fig. 4.8d can also be verified when the LVDC load increases from 2 kW to 67kW at 5s and the system remains stable.

Based on the analysis and simulation results, the CPL at the HVDC bus (ECS) is the critical load in the MEA power system. To further validate the effectiveness of the developed model for the system stability analysis, some case studies are conducted to investigate the impacts of some parameters with both theoretical analysis and simulation verifications. The system reconfiguration with another ECS at the HVDC bus is also studied to validate the effectiveness of the proposed method under different configurations.

Impacts of parameters

1) Effect of system frequency (generator rotational speed)

As the frequency of MEA power system varies from 360Hz to 800Hz, the effect

Figure 4.13: Eigenvalue loci with the variation of HVDC bus inductance values.

of the frequency variation is investigated. Fig. 4.11 shows the eigenvalue loci when the frequency varies from 360Hz to 800Hz whereas other parameters remain the same as the preset value scenario. It can be observed that with the increase of frequency, the dominant eigenvalues move to the left and away from the imaginary axis. This indicates the damping effect with the increase of frequency. Based on the analytical model, the unstable boundaries in terms of the ECS load torque are predicted as 370Nm, 420Nm and 470Nm for system frequencies at 400Hz, 600Hz and 800Hz, respectively. Simulations with different system frequencies are conducted with the results shown in Fig. 4.12. It can be observed that with the increase of the ECS load torque, the system with higher frequency becomes unstable at larger load torque and the unstable boundaries meet the analytical results.

2) Effect of filter parameters

The different filters may affect system stability and the impact of the HVDC bus filter is investigated here as an example, which is represented by L_f and C_f in Fig. A2.

Fig. 4.13 shows the loci of dominant eigenvalues with the filter inductance L_f varies from 1mH to 3mH at a step increase of 0.05mH. As can be observed, the eigenvalues move towards the right hand side of the plane and the system stability margin decreases with the increase of inductance value. When the inductance L_f is set at 1.5mH, 2mH and 2.5mH, the corresponding unstable boundary in terms of ECS load torque is predicted as 480Nm, 370Nm and 300Nm respectively based on the developed analytical model. This can be validated by the simulation result in Fig. 4.14, which presents the system dynamic response of the HVDC bus voltage at different inductance values with the increase of the ECS load torque.

The impact of the HVDC bus filter capacitance C_f is investigated using the

Figure 4.14: Simulation results with different HVDC bus inductance values.

Figure 4.15: Eigenvalue loci with the variation of HVDC bus capacitance values.

developed model in Fig. 4.15, which illustrates the loci of eigenvalues with the variation of C_f from 0.3mF to 2.3mF. It shows that a larger HVDC bus capacitor helps to stabilize the system. When the capacitance value C_f is set at 0.5mF, 0.8mF and 1mF, the corresponding unstable boundary in terms of the ECS load torque is predicted as 370Nm, 400Nm and 430Nm, respectively. Fig. 4.16 shows the system dynamic response under different capacitance values with the increase of the ECS load torque, and the simulation results satisfy the analytical results which are predicted by the developed model.

Figure 4.16: Simulation results with different HVDC bus capacitance values.

System reconfiguration

As mentioned before, the proposed modular approach is flexible for the system extension and reconfiguration. The system configuration with another ECS connected at the HVDC bus is studied. The additional ECS has the same parameters as the existing ECS and it is denoted as the ECS2 module with the matrices represented by subscript $m2$. Taking consideration of the previous system matrix in (4.40) and the proposed approach for parallel modules in (4.19), the new system model is obtained as (4.42) and (4.43).

$$A_{sys} = \begin{bmatrix} A_{sg} & B_{sg,fl}C_{flu} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ B_{fl,sg}C_{sgd} & A_{fl} & B_{fl,r}C_{ru} & 0 & B_{sg,a}C_{au} & 0 & kB_{fl,r1}C_{r1u} & 0 & B_{fl,t}C_{tu} & 0 \\ 0 & B_{r,fl}C_{fld} & A_r & B_{r,m}C_{mu} & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & B_{m,r}C_{rd} & A_m & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & B_{a,fl}C_{fld} & 0 & 0 & A_a & B_{a,m1}C_{m1u} & 0 & 0 & 0 & B_{a,m2}C_{m2u} \\ 0 & 0 & 0 & 0 & B_{m1,a}C_{ad} & A_{m1} & 0 & 0 & 0 & 0 \\ 0 & \frac{1}{k}B_{r1,fl}C_{fld} & 0 & 0 & 0 & 0 & A_{r1} & B_{r1,inv}C_{invu} & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & B_{inv,r1}C_{r1d} & A_{inv} & 0 & 0 \\ 0 & B_{t,fl}C_{fld} & 0 & 0 & 0 & 0 & 0 & 0 & A_t & 0 \\ 0 & 0 & 0 & 0 & B_{m2,a}C_{ad} & 0 & 0 & 0 & 0 & A_{m2} \end{bmatrix} \quad (4.42)$$

Figure 4.17: Eigenvalue loci with the variation ECS load torque.

$$\Delta x_{sys} = [\Delta x_{sg} \quad \Delta x_{fl} \quad \Delta x_r \quad \Delta x_m \quad \Delta x_a \quad \Delta x_{m1} \quad \Delta x_{r1} \quad \Delta x_{inv} \quad \Delta x_t \quad \Delta x_{m2}]_{1 \times 61}^T \quad (4.43)$$

Fig. 4.17 shows the eigenvalue loci when the the ECS load torque varies from 0Nm to 350Nm at a step increase of 10 Nm and the ECS2 load torque remains at 200Nm. Other parameters are the same as that in the preset value scenario. It can be observed that the eigenvalue loci move toward the right hand side of the plane with the increase of the ECS torque and the unstable boundary in terms of the ECS load torque is 300Nm. Under the same process, the unstable boundary is predicted as 250Nm and 340Nm respectively, when the ECS2 load torque is set at 250Nm and 150Nm. These analytical results are validated by simulation results in Fig. 4.18, which shows the system dynamic response with different ECS2 load torque values (250Nm, 200Nm and 150Nm). Compared with the simulation results in Fig. 4.9 that only one ECS is connected at the HVDC bus and the unstable boundary is 370Nm, the unstable boundary of each ECS decreases, but the total load boundary increases.

Based on the above analytical and simulation results, the effectiveness of the proposed approach is verified with different parameters and configurations.

4.4 Conclusion

This chapter proposes a modular modeling approach for modeling and stability analysis complex MGs. Each module is obtained in the form of matrix with interconnection vectors, and system can be integrated by matching the interconnection vectors based on system configuration. The integration process for series, parallel and overall system are detailed. The modifications of the system matrix under reconfiguration scenarios are presented. The proposed method is applied for sta-

Figure 4.18: Simulation results with different load torque values of ECS2.

bility analysis of a complex hybrid MEA power system. The small signal model of each component in the system is derived with its interconnection vectors and the whole system matrix is integrated. The impacts of load at each bus are investigated by the eigenvalue method. It turns out that interactions on the HVDC bus cause instability issues whereas interactions on the HVAC, LVAC and LVDC bus have no negative impact. The impacts of the system frequency, HVDC bus filter inductance and capacitance are also investigated. All the analytical results are validated by simulations in MATLAB/Simulink. This proves the effectiveness of the developed analytical model for predicting unstable region and investigating the impacts of parameters. In conclusion, the proposed approach provides an effective and flexible solution for small signal modeling of complex MGs.

Chapter 5

A Novel Composite Nonlinear Controller for Stabilization of Constant Power Loads

The previous chapter proposes a modular approach for system modeling and stability analysis. The system stability can be guaranteed by properly designing parameters based on the stability analysis. To further stabilize the system with CPLs, a composite nonlinear controller is proposed in this chapter.

5.1 Introduction

Power-electronics based DC distribution systems are becoming increasingly common in vehicular power systems, due to their advantages in terms of weight, volume, efficiency, flexibility, isolation, controllability, etc. [92, 93]. However, as already introduced in Chapter 4, power electronic converter loads, when tightly controlled, behave as constant power loads (CPLs) and have destabilizing effects [44, 92, 94]. Therefore, it is significantly important to explore strategies that can achieve fast dynamic response with the guaranteed system stability.

Numerous strategies have been proposed for the DC/DC power electronic converters loaded by CPLs. Passive damping strategies, such as adding the necessary capacitor or resistor [95] or designing LC filters [96] are simple and effective, whereas they are costly and limited by physical constraints. Active damping methods stabilize the system by modifying the control loops to emulate the passive elements, such as virtual capacitor [97], virtual resistor [98] and virtual

impedance [99]. These active damping methods are equivalent to injecting stabilizing power into the CPL to increase damping, which may in turn lead to the compromise of load performance. To avoid the compromise of load performance, [43] proposes a stabilizing method by modifying the control loop of the source converter to emulate a virtual resistor. But the original converter control loop is still modified and thus the system dynamic response is affected. Moreover, all these active damping methods are based on small signal models, they can only ensure small signal stability near the operating point. Therefore, the results are only local and when a large disturbance happens, these linear control methods may become ineffective and the system may be unstable.

Taking into consideration of the nonlinearity of converters and the negative-impedance characteristics of CPLs, nonlinear control technologies should be implemented to stabilize the system in the large signal sense. In [100], a hybrid model predictive control (MPC) method is proposed for a boost converter feeding a CPL, but the online computational burden of MPC prohibits its practical implementation. Sliding mode control (SMC) is well-known for its strong robustness with guaranteed large signal stability [101]. However, the variable switching frequency requirement will lead to chattering issues and the requirement for measuring the output filter capacitor current will lead to a current sensor in series with the capacitor, resulting in a higher equivalent series resistance (ESR), which in turn degrades the ripple filtering effect and increases the output impedance. In [102], a sliding-mode duty-ratio controller is proposed for the DC/DC buck converter with CPL at fixed switching frequency. The proposed controller is able to stabilize the DC converter system over the entire operating range with the significant variation in load power and input voltage. However, the measurement of the DC capacitor current is still required, resulting in the aforementioned weakness. A passivity-based controller strategy is also employed for designing DC/DC boost converter feeding CPLs [103]. This technique utilizes the physical properties of the system to achieve a simple control law for stabilization. However, there is a high requirement for model accuracy and a tracking error exists with the variation of the operating point. This motivates the design of a nonlinear controller that can stabilize the system and achieve the accurate tracking against uncertainties with easy implementation.

An adaptive backstepping technique is one of the most effective nonlinear control design tools for solving stabilization and tracking problems [104] and it has been applied for power electronic converter systems [105–107]. However, the ex-

isting literatures only consider the resistive load and the implementation of the backstepping technique in the case of CPL has not been reported. In addition, due to the nonminimum phase nature of boost converters, the backstepping algorithm cannot deal with the objective of output voltage regulation directly and it is indirectly achieved by current/power tracking, which may lead to a non-zero steady-state tracking error due to the uncertain load variation around the nominal resistance. Nonlinear disturbance observer (NDO) is an effective technique to estimate uncertainties/disturbances online for nonlinear systems with minimum information of system dynamics [108, 109]. It attracts much interest as it provides a promising solution for disturbance rejection and uncertainty compensation, which is a key objective during the controller design [110].

Based on adaptive backstepping and NDO techniques, this chapter proposes a composite nonlinear controller for stabilization of a DC/DC boost converter system feeding CPLs. The dynamic model of the boost converter is first converted to the Brunovsky's canonical form for the standard backstepping design. The NDO technique is employed to estimate the uncertain load power variation within a fast dynamic response, which serves as a feedforward compensation for the tracking reference and consequently, achieves fast and accurate tracking of the DC bus voltage under large load variations. Then the proposed controller is recursively constructed step by step following the backstepping algorithm. The designed control law not only ensures a large signal stability, but also features fast dynamics and accurate tracking at a wide operating range. The proposed design procedure is also applicable for the stabilization of other DC/DC converters.

This chapter is organized as follows: First, the system model with the stability problem is presented. Next, a composite nonlinear control strategy based on the backstepping and NDO techniques is proposed. Finally, simulation verification and experimental results are demonstrated to verify the proposed control algorithm.

5.2 System Description and Problem Formulation

A typical DC onboard MG is shown in Fig. 5.1 [43]. The main DC bus is supplied by a DC source through the DC/DC source converter. As can be observed, there are a great number of DC/DC converter loads and inverter fed motor drive loads integrated in the system. For a DC/DC converter feeding resistive load, as long as the converter output voltage is tightly regulated, the output power of the converter

Figure 5.1: A typical DC onboard microgrid

is constant and in turn the input power is almost constant. For an inverter-motor drive system, when its speed is tightly controlled, as torque remains constant for a certain operating period, the output power is constant and the input power of the inverter is constant. Therefore, these loads are constant power loads (CPLs) and can be represented by a lumped CPL [44].

The model of the lumped CPL is a controlled current source [44], given by

$$i_{CPL} = \frac{P_{CPL}}{v_{CPL}} \quad (5.1)$$

where P_{CPL} is the power of CPL, i_{CPL} and v_{CPL} are the instantaneous values of the input current and voltage of the CPL.

The CPL has negative impedance characteristics and this will pose negative impact on system performance [94], especially when the CPL is connected in cascaded with a source converter, it will decrease system damping or even cause instability.

Fig. 5.2 shows a simplified DC onboard MG, i.e. an energy source provides DC power through a DC/DC boost converter, feeding a CPL. The lumped resistive load is denoted as R and the lumped CPL is represented by a controlled current source based on the power of CPL P_{CPL} and bus voltage according to (5.1). The

Figure 5.2: The simplified DC distribution system

dynamic model can be represented as

$$\begin{cases} L \frac{di_L}{dt} = E - (1 - u)v_C \\ C \frac{dv_C}{dt} = (1 - u)i_L - \frac{v_C}{R} - \frac{P_{CPL}}{v_C} \end{cases} \quad (5.2)$$

where i_L and v_C are the instantaneous inductor current and capacitor voltage of boost converter; E is the voltage of the energy source; and u is the duty ratio of the switch, which represents the control input signal.

The control objective is that the DC-link capacitor voltage v_C should track the reference voltage v_{Cref} even under large disturbances. As a resistive load has damping effect and CPL will decrease damping, the worst case in terms of system stability is that the DC load is pure CPL [95]. Since CPL introduces nonlinearity into the system, conventional linear controllers are not sufficient to guarantee system stability under large signal disturbances [103]. This motivates the development of a novel nonlinear controller to stabilize CPLs, which will be illustrated in the next section.

5.3 Design of the Proposed Controller

The design procedure of the proposed composite controller is illustrated in this section. First, the state-space model in (5.2) is transformed to the canonical form for standard backstepping design. Second, the NDO is employed to estimate the load variation to improve the tracking accuracy and system dynamics. Finally the proposed controller is designed by the step-by-step backstepping algorithm with guaranteed globally asymptotical stability.

5.3.1 Coordinate transformation

For the design of the proposed backstepping controller, the system model in (5.2) should first be transformed into the standard form. Based on the result of the exact feedback linearization of DC/DC converters [111], a coordinate transformation is adopted in which the new coordinate has the meaning of the total stored energy and rate of change of stored energy, respectively.

The total energy stored in the dynamic system is expressed as

$$x_1 = \frac{1}{2}Li_L^2 + \frac{1}{2}Cv_C^2 \quad (5.3)$$

Taking the derivative of (5.3) yields

$$\dot{x}_1 = Li_L\dot{i}_L + Cv_C\dot{v}_C = Ei_L - \frac{v_C^2}{R} - P_{CPL} \quad (5.4)$$

Based on (5.4), a new state x_2 and an uncertain term d_1 can be defined as follows

$$x_2 = Ei_L - \frac{v_C^2}{R_0} \quad (5.5)$$

$$d_1 = -P_{CPL} + \frac{v_C^2}{R_0} - \frac{v_C^2}{R} \quad (5.6)$$

where R_0 represents the nominal resistance of the resistive load.

Taking the derivative of both sides in (5.5) yields

$$\begin{aligned} \dot{x}_2 = & \frac{E^2}{L} + \frac{2v_C^2}{R_0^2C} - \left(\frac{Ev_C}{L} + \frac{2i_Lv_C}{R_0C} \right) (1-u) \\ & + \frac{2}{R_0C} \left(P_{CPL} - \frac{v_C^2}{R_0} + \frac{v_C^2}{R} \right) \end{aligned} \quad (5.7)$$

Now define the intermediate control law v and the uncertain term d_2 as

$$v = \frac{E^2}{L} + \frac{2v_C^2}{R_0^2C} - \left(\frac{Ev_C}{L} + \frac{2i_Lv_C}{R_0C} \right) (1-u) \quad (5.8)$$

$$d_2 = \frac{2}{R_0C} \left(P_{CPL} - \frac{v_C^2}{R_0} + \frac{v_C^2}{R} \right) \quad (5.9)$$

Then the system in (5.2) can be transformed to a canonical form as

$$\begin{cases} \dot{x}_1 = x_2 + d_1 \\ \dot{x}_2 = v + d_2 \end{cases} \quad (5.10)$$

In the system expressed by (5.10), x_1 and x_2 are state variables, d_1 and d_2 are uncertain items and v is the intermediate control signal. As can be observed, the system model is expressed in the Brunovsky's canonical form, which is suitable for the employment of the backstepping algorithm.

For the control of the DC/DC boost converter, the control objective is to design the control law u so that the DC bus voltage v_C can track its reference value v_{Cref} asymptotically. After the coordinate transformation, the voltage tracking objective is converted to designing the intermediate control law v to achieve the asymptotic tracking of state x_1 to its reference value x_1^* , which is given by

$$x_1^* = \frac{1}{2}Li_{Lref}^2 + \frac{1}{2}Cv_{Cref}^2 = \frac{1}{2}L\left(\frac{P_{ref}}{E}\right)^2 + \frac{1}{2}Cv_{Cref}^2 \quad (5.11)$$

where P_{ref} is the value of the total load power, expressed as

$$P_{ref} = P_{CPL} + \frac{v_{Cref}^2}{R} \quad (5.12)$$

Once the intermediate control signal v in (5.10) is designed, the final control law u in the original system (5.2) can be obtained according to (5.8) as

$$u = 1 - \left(\frac{E^2}{L} + \frac{2v_C^2}{R_0^2C} - v \right) / \left(\frac{Ev_C}{L} + \frac{2i_L v_C}{R_0C} \right) \quad (5.13)$$

5.3.2 Load power estimation via nonlinear disturbance observer

It can be observed from (5.11) and (5.12) that the reference value x_1^* varies with the load power, which is an uncertain value and this may cause tracking errors. As NDO is an effective technique to estimate uncertainties [110], it is employed to estimate uncertain items d_1 , d_2 and the total load power in (5.12) to achieve accurate tracking and fast dynamic response.

Note that the uncertainties d_i ($i = 1, 2$) are related to the load power according to (5.6) and (5.9), hence from a practical point of view, their values and derivatives should be bounded. Moreover, the load power is seen as constant at steady state. Therefore, the following assumptions can be made:

Assumption 1: The uncertain variables d_i , $\dot{d}_i$, ($i = 1, 2$) for the system (5.10) satisfy the following two conditions:

$$d_i(t) \in L_\infty, \dot{d}_i(t) \in L_\infty \quad (5.14)$$

$$\lim_{t \rightarrow \infty} \dot{d}_i = 0 \quad (5.15)$$

According to NDO in [110], the uncertain item d_1 is estimated by

$$\begin{cases} \hat{d}_1 = l_1(x_1 - p_1) \\ \dot{p}_1 = x_2 + \hat{d}_1 \end{cases} \quad (5.16)$$

where p_1 is an auxiliary state of the NDO and l_1 is a positive constant, denoted as the NDO gain.

Similarly, the uncertain item d_2 is estimated by

$$\begin{cases} \hat{d}_2 = l_2(x_2 - p_2) \\ \dot{p}_2 = v + \hat{d}_2 \end{cases} \quad (5.17)$$

where p_2 is an auxiliary state of the NDO and l_2 is a positive constant.

According to (5.16) and (5.17), the estimation error of d_i ($i=1, 2$) and its derivative are expressed as

$$\tilde{d}_i = d_i - \hat{d}_i \quad (5.18)$$

$$\dot{\tilde{d}}_i = \dot{d}_i - \dot{\hat{d}} = \dot{d}_i - l_i \tilde{d}_i \quad (5.19)$$

Based on (5.6), (5.12) and (5.16), the total load power P_{ref} in (5.12) can be estimated as

$$P_{ref} = \frac{v_{Cref}^2}{R_0} - \hat{d}_1 \quad (5.20)$$

Then the reference value of state x_1 in (5.11) is reformulated as

$$x_1^* = \frac{1}{2} \frac{L}{E^2} \left(\frac{v_{Cref}^2}{R_0} - \hat{d}_1 \right)^2 + \frac{1}{2} C v_{Cref}^2 \quad (5.21)$$

5.3.3 Recursive backstepping design

Based on the standard model in (5.10) and fload power estimated by NDO, the proposed controller can be designed by following the adaptive backstepping algorithm in this section. As the design objective is to enforce the state variables x_1 and x_2 to track their desired values x_1^* and x_2^* asymptotically, a new set of coordinates is introduced as

$$\begin{aligned} z_1 &= x_1 - x_1^* \\ z_2 &= x_2 - x_2^* \end{aligned} \quad (5.22)$$

Step 1:

Taking the derivative of z_1 along its trajectory yields

$$\dot{z}_1 = z_2 + x_2^* + d_1 - \dot{x}_1^* \quad (5.23)$$

Consider the Lyapunov function defined as

$$V_1 = \frac{1}{2}z_1^2 + \frac{1}{2}\tilde{d}_1^2 \quad (5.24)$$

The derivative of (5.24) along its trajectory is

$$\dot{V}_1 = z_1(x_2^* + z_2 + \hat{d}_1 - \dot{x}_1^*) + z_1\tilde{d}_1 - l_1\tilde{d}_1^2 + \tilde{d}_1\dot{d}_1 \quad (5.25)$$

To stabilize (5.23) with the Lyapunov function (5.24), the virtual control law x_2^* is designed as

$$x_2^* = -k_1z_1 - \hat{d}_1 + \dot{x}_1^* \quad (5.26)$$

Then the resulting expression of $\dot{V}_1$ is obtained as

$$\dot{V}_1 = -k_1z_1^2 + z_1z_2 + z_1\tilde{d}_1 - l_1\tilde{d}_1^2 + \tilde{d}_1\dot{d}_1 \quad (5.27)$$

As is seen from (5.27), there is no guarantee that $\dot{V}_1$ is negative definite. In the next step, the intermediate control signal v is selected such that this condition is satisfied.

Step 2:

Again taking the derivative of z_2 along its trajectory yields

$$\dot{z}_2 = v + d_2 + k_1\dot{z}_1 + \dot{d}_1 - \ddot{x}_1^* \quad (5.28)$$

Select the Lyapunov function

$$V_2 = V_1 + \frac{1}{2}z_2^2 + \frac{1}{2}\tilde{d}_2^2 \quad (5.29)$$

whose derivative is expressed as

$$\begin{aligned} \dot{V}_2 = & -k_1z_1^2 + z_1z_2 + z_1\tilde{d}_1 - l_1\tilde{d}_1^2 + \tilde{d}_1\dot{d}_1 + z_2[v + d_2 \\ & + k_1(z_2 - k_1z_1 + \tilde{d}_1) + l_1\tilde{d}_1 - \ddot{x}_1^*] - l_2\tilde{d}_2^2 + \tilde{d}_2\dot{d}_2 \end{aligned} \quad (5.30)$$

To stabilize (5.28) with the Lyapunov function (5.29), the intermediate control

law v is designed as

$$v = -k_2 z_2 - \hat{d}_2 + \ddot{x}_1^* \quad (5.31)$$

Theorem 1: The closed-loop system consisting of (5.2), (5.10), (5.22), (5.26) and (5.31) is globally asymptotically stable and the DC bus voltage v_C can track its reference value v_{Cref} asymptotically.

Proof. With Lemma 3 in Appendix B, the following relations hold

$$z_1 z_2 \leq 0.5 z_1^2 + 0.5 z_2^2 \quad (5.32)$$

$$z_1 \tilde{d}_1 \leq 0.5 z_1^2 + 0.5 \tilde{d}_1^2 \quad (5.33)$$

$$\tilde{d}_1 \dot{d}_1 \leq 0.5 \tilde{d}_1^2 + 0.5 \dot{d}_1^2 \quad (5.34)$$

$$z_2 k_1 (z_2 - k_1 z_1) \leq 0.5 z_1^2 + \left(\frac{k_1^4}{2} + \frac{1}{2k_1^2} \right) z_2^2 \quad (5.35)$$

$$z_2 (k_1 + l_1) \tilde{d}_1 \leq 0.5 \tilde{d}_1^2 + \frac{(k_1 + l_1)^2}{2} z_2^2 \quad (5.36)$$

$$\tilde{d}_2 \dot{d}_2 \leq 0.5 \tilde{d}_2^2 + 0.5 \dot{d}_2^2 \quad (5.37)$$

By substituting (5.32)-(5.37) into (5.30) and doing some algebraic manipulations, eqn. (5.30) can be derived as

$$\begin{aligned} \dot{V}_2 \leq & -(k_1 - 1.5) z_1^2 - \left[k_2 - 0.5 - \left(\frac{k_1^4}{2} + \frac{1}{2k_1^2} \right) - \frac{(k_1 + l_1)^2}{2} \right] \\ & \cdot z_2^2 - (l_1 - 1.5) \tilde{d}_1^2 - (l_2 - 0.5) \tilde{d}_2^2 + \frac{1}{2} \dot{d}_1^2 + \frac{1}{2} \dot{d}_2^2 \end{aligned} \quad (5.38)$$

By selecting $l_1 \geq 3$, $l_2 \geq 2$, $k_1 \geq 2.5$ and

$$k_2 \geq \frac{k_1^4}{2} + \frac{1}{2k_1^2} + \frac{(k_1 + l_1)^2}{2} + 1.5$$

the derivative of V_2 can be obtained as

$$\begin{aligned} \dot{V}_2 \leq & -(k_1 - 1.5) z_1^2 - \left[k_2 - 0.5 - \left(\frac{k_1^4}{2} + \frac{1}{2k_1^2} \right) - \frac{(k_1 + l_1)^2}{2} \right] \\ & \cdot z_2^2 - (l_1 - 1.5) \tilde{d}_1^2 - (l_2 - 0.5) \tilde{d}_2^2 + \frac{1}{2} \dot{d}_1^2 + \frac{1}{2} \dot{d}_2^2 \end{aligned} \quad (5.39)$$

Since V_2 involves V_1 , its stabilization automatically guarantees the stability of the first loop in Step 1. Considering Lemma 1 in Appendix B and Assumption 1, and taking $\dot{d}_i$ ($i=1, 2$) as perturbation inputs, eqn. (5.39) provides the proof

Figure 5.3: Control architecture of the proposed controller

that the closed-loop system of (5.2), (5.10), (5.22), (5.26) and (5.31) is globally input-to-state stable (ISS).

By using Lemma 2 in Appendix B and Assumption 1 with (5.39), the closed-loop ISS system is globally asymptotically stable with $\lim_{t \rightarrow \infty} z_1 \rightarrow 0$. This implies the asymptotic tracking of x_1 at x_1^* and thus the DC bus voltage v_C can track its reference value v_{Cref} asymptotically. $\square$

Based on the procedure developed above, the control architecture of the proposed controller is depicted in Fig. 5.3.

Remark: It is worth mentioning that the design procedure for the proposed controller is also applicable to other DC/DC converter topologies such as buck converter and buck-boost converter. The system model is first transformed into canonical form in (5.10) based on the exact feedback linearization technique. Then the nonlinear controller for that specific system can be designed by following the NDO estimation procedure in (5.16)-(5.21) and the backstepping procedure in (5.22)-(5.39). With the NDO and backstepping algorithm, the designed composite controller can achieve the fast dynamic response and accurate bus voltage tracking with guaranteed large signal stability.

Table 5.1: System Parameters

Variables	Description	Value
V_{Cref}	DC bus voltage reference	50V
E	Converter input voltage	25V
f_s	Switching frequency	20kHz
L	Inductance	1mH
C	Capacitance	0.47mF

5.4 Simulation Results

The system model described by Fig. 5.2 with the proposed controller in Fig. 5.3 is simulated in Matlab/Simulink to validate the effectiveness of the proposed controller. The system parameters are listed in Table 5.1. As the worst case in terms of stability is pure CPL without the resistive load, the nominal resistive load is 0W and R_0 is seen as infinite (set at a very large value such as 1000000Ω). First, the procedure for tuning design parameters is illustrated and then the simulation results under different operating conditions will be presented.

5.4.1 Tuning of design parameters

The proposed composite nonlinear controller consists of two NDOs for the load estimation and a backstepping controller for the DC bus voltage regulation. As the NDOs provide the reference value for the backstepping controller, NDOs should be designed first with the control parameters k_1 and k_2 roughly selected as 200 and 2000. Fig. 5.4 shows the dynamic response of the observed load variation when CPL increases from 50W to 100W under different values of l_1 ($l_1=200, 500, 2000$ and 5000), while the NDO gain l_2 is fixed at 200. As can be observed from Fig. 5.4, the NDOs can track the load variation within a very short transition (in milliseconds) and a larger l_1 will result in a faster convergence rate. Fig. 5.5 shows the simulation results under different values of l_2 ($l_2 = 200, 500, 2000$ and 5000) when l_1 is fixed at 2000. It reveals that the impact of l_2 on system performance is much smaller than that of l_1 .

Next, the gains k_1 and k_2 are tuned while the NDO gains l_1 and l_2 are fixed at 2000 and 200. As the objective of the backstepping controller is to regulate the bus voltage to its desired value, the dynamic responses of the bus voltage with different controller gains are investigated. The simulation results with different values of k_2 ($k_2=200, 500, 2000$ and 5000) are shown in Fig. 5.6 when k_1 is fixed

Figure 5.4: Load estimation response with different value of l_1

Figure 5.5: Load estimation response with different value of l_2

Figure 5.6: Voltage tracking response with different values of k_2

Figure 5.7: Voltage tracking response with different values of k_1

at 200. It is shown that a larger k_2 will lead to a shorter settling time for the voltage tracking. Fig. 5.7 shows the simulation results with different values of k_1 ($k_1 = 200, 500, 1000, 2000$ and 5000) when k_2 is fixed at 2000. It can be observed that a larger k_1 will cause a faster convergence rate, however, an extremely large value may lead to large overshoot.

It should be noticed that, as a common problem of the existing high gain

Figure 5.8: Simulation results with resistive load and CPL

control methods, a larger gain will cause deterioration of system robustness against the measurement noise. Therefore, we should select proper values based on the required system dynamics. Furthermore, as the NDO provides the reference value for the backstepping controller, the dynamic response of load estimation should be designed faster than the voltage tracking dynamics. For example, if the required voltage tracking settling time is within 20ms, the dynamics of load estimation can be expected as around 10ms. Thus l_1 and l_2 can be selected as 500 and 200, while the gains of k_1 and k_2 can be selected at 200 and 2000. This set of parameters will be used in the simulations and experiments below to verify the effectiveness of the proposed controller and to show its advantage in terms of stability by comparing with the conventional PI controller with similar dynamics.

5.4.2 Simulation verification

First, the situation with the integration of both resistive load and CPL is considered and the simulation result is shown in Fig. 5.8. Originally, there is 25W resistive load at DC bus. Then 50W CPL is connected at 0.2s and disconnected at 0.4s. As can be observed, at each load change period, the bus voltage can be enforced to the nominal value within 10ms and the control signal u , which is the duty ratio of the converter, is maintained at 0.5 as expected. Moreover, since

Figure 5.9: Simulation results with CPL variation

Figure 5.10: Simulation results with output voltage variation

the total load power is estimated as (5.20) by NDO, the effectiveness of the NDO estimator is validated by the fourth plot in Fig. 5.8. It indicates that after each

Figure 5.12: Experiment results with both resistive load and CPL under the proposed controller

5.5.1 System performance under the proposed strategy

Fig. 5.12 shows the dynamic response of the system under the proposed controller with the integration of both resistive load and CPL. Originally, 25W resistive load is connected at the DC bus. At 0.4s, 50W CPL is integrated and at 1.48s, the CPL is disconnected. It can be observed that at each time, the voltage responds immediately and can track its reference value accurately with the settling time less than 40ms.

Then the resistive load is removed and the system is loaded by pure CPL. Fig. 5.13 shows the dynamic response of the output voltage, input voltage and inductor current. The original load is set as 100W. Then the CPL increases to a heavy load condition of 160W. It is shown that the output voltage accurately tracks its reference value at 50V within 40ms and the system is stable.

The impact of the output voltage variation is shown in Fig. 5.14. Only the CPL is connected at the DC bus and it is maintained at a heavy load condition of 160W. At 2.1s, when the output voltage reference value decreases from 50V to 45V, its real value quickly tracks to the reference value. The inductor current has a transient dip due to the voltage change and it is stabilized at a constant value at steady state. As can be observed, the stable operation of the system is guaranteed under the output voltage variation.

The impact of the input voltage variation is shown in Fig. 5.15. The load

Figure 5.13: Experiment results with the variation of large CPL under the proposed controller

Figure 5.14: Experiment results with output voltage variation under the proposed controller

maintains at 50W. At 1.1s, the input voltage decreases from 25V to 20V, the inductor current increases consequently whereas the output voltage maintains at 50V with negligible disturbances. Thus, the proposed method ensures stable operation under the input voltage variation.

Based on the above results, the proposed controller ensures fast dynamic per-

Figure 5.15: Experiment results with input voltage variation under the proposed controller

formance, accurate voltage tracking and guaranteed system stability, being consistent with the theoretical results in Theorem 1 and simulation verification in Section 5.4.2. There exist some minor discrepancies between simulation and experimental results. The dynamic response for the experimental case is a bit slower than the corresponding simulation result. This is acceptable because in the simulation studies, the ideal constant power load with the step change is adopted, whereas in the experimental tests, the variation of CPL (Chroma load in CPL mode) is restricted by its programmable controller bandwidth. Moreover, the experimental results suffer from real-life system conditions such as EMI, noise, ESR, delays and various uncertainties/disturbances, which are not considered in the simulations.

5.5.2 Comparison with conventional PI controller

To further illustrate the advantages of the proposed controller, the conventional double-loop PI controller is implemented for the boost converter with an inner current loop and an outer voltage loop [112].

As PI parameters have an impact on system stability and dynamic response, faster dynamics will lead to less stability margin [112]. To have a fair comparison, the PI controller parameters are tuned to have similar dynamic response with the proposed controller. In this case, the inner and outer loop PI gains k_{cp} , k_{ci} , k_{vp}

Figure 5.16: Experiment results with both resistive load and CPL under conventional PI controller

Figure 5.17: Experiment results with variation of large CPL under conventional PI controller

and k_{vi} are set at 0.0628, 39.4, 0.8 and 37.1, respectively. Other parameters are the same as that in Table 5.1. Fig. 5.16 shows the dynamic response under the conventional PI controller with the same loading profile in Fig. 5.12. It reveals that with load changes, the voltage tracking is achieved with settling time around 40ms, which is similar to that under the proposed strategy. Then the stability margin of the two methods are compared.

Figure 5.18: Experiment results with output voltage variation under conventional PI controller

To compare the stability margin, the case of large CPL variation in Fig. 5.13 is applied for the system under the PI controller. When the CPL steps from 120W to 160W, a large oscillation occurs and increases until the inductor current i_L reaches its upper limit of 10A, then the protection system is triggered and the power source is turned down automatically, as is shown in Fig. 5.17. But for the proposed controller, the experiment result in Fig. 5.13 shows accurate voltage tracking under stable operation. With the same loading profile, the proposed strategy achieves the control objective and maintains the system stability, whereas the conventional PI controller fails to do so.

Similarly, the impact of the output voltage variation implemented in Fig. 5.14 is investigated for the system with the PI controller in Fig. 5.18. When the output voltage reference value steps down from 50V to 46V with CPL fixed at 150W, a large oscillation occurs at the inductor current i_L , as is shown in Fig. 5.18. But the dynamic response in Fig. 5.14 illustrates that stable operation with accurate tracking is still maintained with the proposed controller under the same condition.

Therefore, based on the above comparisons, with similar dynamic performance, the proposed controller can guarantee accurate voltage tracking with the large CPL variation and output voltage variation, which cannot be done by the conventional PI controller.

5.6 Conclusion

This chapter proposes a composite nonlinear controller for the DC/DC boost converter feeding CPLs based on NDO and backstepping algorithms. The system model is first converted to the canonical form. The NDO technique is employed to estimate load power within fast dynamic response to ensure accurate voltage tracking. Then by following a recursive design procedure, the proposed controller is designed with guaranteed large signal stability. Both simulations and experiments are conducted to verify the effectiveness of the proposed controller. The advantages of the proposed controller are further illustrated by its comparison with the conventional PI controller. The proposed controller design procedure is also applicable for stabilization of other DC/DC converters.

Chapter 6

Conclusion and Future Works

6.1 Conclusion

In this thesis, decentralized control strategies and stability analysis tools are developed to deal with the challenges in isolated MGs, i.e. the critical demand-supply balance under the intermittent renewable generators and the stability issues emerged from the high penetration of constant power loads. The main results are summarized as follows:

Chapter 2 and Chapter 3 consider the power balance issue. A decentralized power sharing strategy is proposed in Chapter 2 for the hybrid energy storage system to compensate the imbalanced power caused by the RESs. It consists of a virtual capacitance droop controller for the SC converter and a virtual resistance droop controller for battery regulation. Using this strategy, the autonomous dynamic power sharing is achieved that SC smooths high frequency fluctuations during transient and the battery compensates steady state power mismatch. The effectiveness of the proposed strategy is verified by the simulation and experimental results.

As an improvement for the work in Chapter 2, Chapter 3 proposes a decentralized power management strategy with the consideration of the bus voltage restoration and SC state-of-charge recovery. A virtual high-pass filter based droop controller is proposed to regulate the battery converter and a virtual capacitance droop controller is implemented for the SC converter. The proposed strategy can achieve the dynamic power sharing, bus voltage restoration and SC state-of-charge recovery in a fully decentralized manner. Simulations and experiments are conducted to validate the proposed approach. As no communication net-

works and central controllers are required, the proposed approaches in Chapter 2 and Chapter 3 can achieve high reliability, scalability and plug-and-play features. They provide effective power management schemes for systems with hybrid energy sources, such as the battery/SC system and the fuel cell/battery system. The proposed strategies can be applied in many applications including electric vehicular systems, DC MGs, hybrid AC/DC MGs, wind power plant, smart buildings and many other systems with the integration of multiple energy sources. The corresponding works have been published in [J1] and [J2] in Author's Publications.

Chapter 4 and Chapter 5 deal with the stability issue. A modular approach is proposed in Chapter 4 for effective system modeling and stability analysis. By this means, a complex system matrix can be easily integrated based on system configuration and the overall system stability can be evaluated. The hybrid AC/DC more electric aircraft power system adopted by Boeing 787 is analyzed using the proposed technique as an application example. The proposed approach provides an effective and flexible approach for modeling and stability analysis of complex MGs with system extension, reconfiguration and plug-and-play requirements. The corresponding work has been published in [J3] in Author's Publications.

Chapter 5 proposes a composite nonlinear controller for stabilization of constant power loads. The nonlinear disturbance observer technique is adopted to estimate the load power variation within a fast dynamic response, serving as a feedforward compensation to increase the accuracy of output voltage regulation. Then a nonlinear controller is developed by following the step-by-step backstepping algorithm with a strictly guaranteed large signal stability. The proposed controller not only ensures global stability under large variations of the constant power loads, but also features a fast dynamic response with accurate tracking over a wide operating range. Both simulations and experiments are conducted to verify the proposed controller. It provides a simple and effective solution for the stable operation of DC MGs. The corresponding work has been published in [J4] in Author's Publications.

6.2 Future works

Besides the work presented in this thesis, there are some interesting research topics that are worthy to be explored in the future study, as listed below:

- 1) *Nonlinear techniques for decentralized power sharing and stabilization*

In Chapter 2-3, the droop based control strategies are implemented for de-

centralized power sharing in isolated MGs. But the inner voltage and current control of the converters are based on the conventional PI control, which cannot ensure large signal stability. In Chapter 5, a nonlinear controller based on adaptive backstepping algorithm is proposed for stabilizing DC/DC converter feeding CPL. But they are only applicable for a single converter system with CPL to the current stage, and will be very complicated for a system with multiple converters. However, for a MG in real applications, there are multiple parallel connected power converters; a prominent feature of the MG is that it has open boundary and allows distributed generators to plug and play flexibly, especially with the increasing penetration of distributed generators and electric vehicles in the future. Thus the structure is not fixed and it will be difficult to perform the overall system stability analysis. To solve this problem, each converter should be designed without system information whereas the stability of the whole microgrid system is guaranteed. Therefore, in the future, nonlinear controllers should be developed to achieve decentralized power sharing with the guaranteed large signal stability.

2) Large signal stability analysis

The proposed modular approach in Chapter 4 is a small signal technique and could not ensure system stability at large disturbances, such as faults, generator start up, sudden load connections or disconnections, etc. Due to the nonlinearity of the power electronic loads and the complexity of system, it is significant to perform large signal analysis. The stability issue is especially important for on-board power systems of vehicular systems, such as more electric aircrafts, electric vehicles and all electric ships, where there is a high penetration of constant power loads. Currently, some analytical methods to investigate large signal stability have been proposed in literatures and there is a detailed review on these methods in [76]. However, these large signal analysis methods are only reported for simple systems, such as inverter-fed motor drive systems with DC power sources. They are not able to handle the complex MGs with multiple power converters at the current stage of development. Large signal analysis should be carried out in the near future, hoping to get the boundaries of the asymptotic stability region and understand the transient behavior of the on-board power systems through such analytical method.

3) Reliability analysis

Power semiconductor devices and electrolytic capacitors are reported to be the most vulnerable links in power converters [113]. Oscillation and transient dynamics will pose stress on these sensitive devices, affect their life cycles and

finally impact system reliability. During stability analysis in Chapter 4, it can also be observed from simulation results that, there are transient dynamics and oscillations with the variation of loads. As there is widespread utilization of power electronic converters in MGs, the system reliability is significant, especially considering these vulnerable power electronic devices. System reliability considering the transients and oscillations can be evaluated for MGs in the future to ensure reliable operation and control.

4) Multiple interconnected MGs

Energy sharing among multiple interconnected MGs is a promising solution for efficient, reliable and economic operation of MGs. It enables MGs with energy surplus to share energy to those with energy deficit and allows exploiting the averaging properties of stochastic processes such as renewable power generation, which makes less stringent the requirements on the energy storage control. Moreover, as the distance among the MGs in a distribution network is normally smaller than that between MGs and the utility grid, system transmission losses can be reduced. There are many issues that should be considered for the interconnected multiple MGs, including modeling, control, power management and energy management. The research on the interconnected-MGs system is a very promising area that should be explored in the future.

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Figure A6: Line model.

$$\begin{aligned}\dot{X}_{vq} &= V_{inq}^* - V_{inq} \\ \dot{X}_{id} &= k_{vdp}(V_{ind}^* - V_{ind}) + k_{vdi}X_{vd} - I_{ind} \\ \dot{X}_{iq} &= k_{vqp}(V_{inq}^* - V_{inq}) + k_{vqi}X_{vq} - I_{inq}\end{aligned}$$

A.1.6 Line model

$$\begin{aligned}\dot{I}_{fld} &= -\frac{R_{fl}}{L_{fl}}I_{fld} + \omega I_{flq} - \frac{1}{L_{fl}}V_{flod} + \frac{1}{L_{fl}}V_{flind} \\ \dot{I}_{flq} &= -\omega I_{fld} - \frac{R_{fl}}{L_{fl}}I_{flq} - \frac{1}{L_{fl}}V_{floq} + \frac{1}{L_{fl}}V_{flinq} \\ \dot{V}_{flod} &= \frac{1}{C_{fl}}I_{fld} + \omega V_{flq} - \frac{1}{C_{fl}}I_{flod} \\ \dot{V}_{flq} &= \frac{1}{C_{fl}}I_{flq} - \omega V_{fld} - \frac{1}{C_{fl}}I_{floq}\end{aligned}$$

A.2 Submatrices

$$H_0 = \begin{bmatrix} -(L_{ls} + L_{md}) & L_{md} & L_{md} & 0 & 0 & 0 & 0 & 0 & 0 \\ -L_{md} & L_{lF} + L_{md} & L_{md} & 0 & 0 & 0 & -k_{pgi} & 0 & 0 \\ -L_{md} & L_{md} & L_{lk d} + L_{md} & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -(L_{ls} + L_{mq}) & L_{mq} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & -L_{mq} & L_{lk q} + L_{mq} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & -k_{pgv} & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}_{9 \times 9}$$

$$A_{sg} = \begin{bmatrix} R_{gs} & 0 & 0 & -\omega(L_{ls} + L_{mq}) & \omega L_{mq} & 0 & 0 & 1 & 0 \\ 0 & -R_F & 0 & 0 & 0 & 0 & k_{igi} & 0 & 0 \\ 0 & 0 & -R_{kd} & 0 & 0 & 0 & 0 & 0 & 0 \\ \omega(L_{ls} + L_{md}) & -\omega L_{md} & -\omega L_{md} & R_{gs} & 0 & 0 & 0 & 0 & 1 \\ 0 & 0 & 0 & 0 & -R_{kq} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & -1 & 0 & 0 & 0 & k_{igv} & 0 & 0 & 0 \\ -C_{eq}^{-1} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & \omega \\ 0 & 0 & 0 & -C_{eq}^{-1} & 0 & 0 & 0 & -\omega & 0 \end{bmatrix}_{9 \times 9}$$

$$B_{sg,sgD} = H_0^{-1} \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & -C_{eq}^{-1} & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -C_{eq}^{-1} \end{bmatrix}_{2 \times 9}^T$$

$$C_{sgd} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}_{2 \times 9}$$

$$A_a = \begin{bmatrix} -\frac{R_{eq}}{L_{eq}} & \omega & -\frac{1}{L_{eq}} & 0 & 0 & 0 \\ -\omega & -\frac{R_{eq}}{L_{eq}} & 0 & -\frac{1}{L_{eq}} & 0 & 0 \\ \frac{1}{C_{eq2}} & 0 & 0 & \omega & -\frac{\sqrt{12}\sin(\delta_0+\varphi_0)}{\pi C_{eq2}} & 0 \\ 0 & \frac{1}{C_{eq1}} & -\omega & 0 & -\frac{\sqrt{12}\cos(\delta_0+\varphi_0)}{\pi C_{eq1}} & 0 \\ 0 & 0 & \frac{\sqrt{27}\sin(\delta_0+\varphi_0)}{\pi L_f} & \frac{\sqrt{27}\cos(\delta_0+\varphi_0)}{\pi L_f} & -\frac{r_f+r_\mu}{L_f} & -\frac{1}{L_f} \\ 0 & 0 & 0 & 0 & \frac{1}{C_f} & 0 \end{bmatrix}_{6 \times 6}$$

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$$B_{a,aD} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & -C_f^{-1} \end{bmatrix}_{1 \times 6}^T B_{a,aU} = \begin{bmatrix} \frac{1}{L_{eq}} & 0 & 0 & 0 & 0 & 0 \\ 0 & \frac{1}{L_{eq}} & 0 & 0 & 0 & 0 \end{bmatrix}_{1 \times 6}^T$$

$$C_{au} = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 \end{bmatrix}_{2 \times 6} C_{ad} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}_{1 \times 6}$$

$$B_{r,rU} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & a_{r7} \\ 0 & 0 & 0 & 0 & 0 & a_{r8} \end{bmatrix}_{2 \times 6}^T B_{r,rD} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & -\frac{1}{C_{rf}} \end{bmatrix}_{1 \times 6}^T$$

$$C_{ru} = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 \end{bmatrix}_{2 \times 6} C_{rd} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}_{1 \times 6}$$

$$A_r = \begin{bmatrix} -\frac{R_{fc}+k_{rdp}}{L_{fc}} & 0 & \frac{k_{rei}k_{rdp}}{L_{fc}} & \frac{k_{rdi}}{L_{fc}} & 0 & -\frac{k_{rep}k_{rdp}}{L_{fc}} \\ 0 & -\frac{R_{fc}+k_{rqp}}{L_{fc}} & 0 & 0 & \frac{k_{rqi}}{L_{fc}} & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 \\ -1 & 0 & k_{rei} & 0 & 0 & -k_{rep} \\ 0 & -1 & 0 & 0 & 0 & 0 \\ a_{r1} & a_{r2} & a_{r3} & a_{r4} & a_{r5} & a_{r6} \end{bmatrix}_{6 \times 6}$$

$$a_{r1} = \frac{3}{2} \frac{V_{sd0} - k_{rei}k_{rdp}X_{re0} + 2k_{rdp}I_{rd0} - k_{rdi}X_{rd0}}{C_{rf}V_{dc10}}; a_{r2} = \frac{3}{2} \frac{V_{sq0} + k_{pcq}I_{rq0} - k_{rqi}X_{rq0}}{C_{rf}V_{dc10}};$$

$$a_{r3} = -\frac{3}{2} \frac{k_{rei}k_{rdp}I_{rd0}}{C_{rf}V_{dc10}}; a_{r4} = -\frac{3}{2} \frac{k_{rdi}I_{rd0}}{C_{rf}V_{dc10}}; a_{r5} = -\frac{3}{2} \frac{k_{rqi}I_{rq0}}{C_{rf}V_{dc10}};$$

$$a_{r6} = -\frac{3(V_{rd}^*I_{rd0} + V_{rq}^*I_{rq0})}{2C_{rf}V_{dc10}^2}; a_{r7} = \frac{3}{2} \frac{I_{rd0}}{C_{rf}V_{dc10}}; a_{r8} = \frac{3}{2} \frac{I_{rq0}}{C_{rf}V_{dc10}};$$

$$A_m = \begin{bmatrix} 0 & \frac{K_T}{J_m} & 0 & 0 & 0 \\ -\frac{pK_f}{2\tau_s R_{sm}} & -\frac{1}{\tau_s} - \frac{K_f}{\tau_s R_{sm} \tau_r I_{sdm0}} & -\frac{V_{sqm0}}{\tau_s R_{sm} V_{f0}} & \frac{1}{\tau_s R_{sm}} & 0 \\ 0 & 0 & -\frac{1}{\tau_f} & 0 & 0 \\ -K_{pim}K_{i\omega} + \frac{pK_{pim}K_f}{2\tau_s R_{sm}} & a_{m42} & \frac{K_{pim}V_{sqm0}}{\tau_s R_{sm} V_{f0}} & -\frac{K_{pim}}{\tau_s R_{sm}} & K_{iim} \\ -K_{i\omega} & -\frac{K_{p\omega}K_T}{J_m} & 0 & 0 & 0 \end{bmatrix}_{5 \times 5}$$

$$B_{m,mU} = \begin{bmatrix} 0 & \frac{M_{q0}}{2\tau_s R_{sm}} & \frac{1}{2\tau_f} & -\frac{K_{pim}M_{q0}}{2\tau_s R_{sm}} & 0 \end{bmatrix}_{1 \times 5}^T$$

$$C_{mu} = \begin{bmatrix} \frac{3p\sigma L_{sm}I_{sqm0}I_{sdm0}}{8C_f V_{f0}} & a_{m2} & a_{m3} & -\frac{3I_{sqm0}}{4C_f V_{f0}} & 0 \end{bmatrix}_{1 \times 5}$$

$$a_{m42} = -K_{iim} - \frac{K_{pim}K_{p\omega}K_T}{J_m} + \frac{K_{pim}}{\tau_s} + \frac{K_{pim}K_f}{\tau_s R_{sm} \tau_r I_{sdm0}};$$

$$a_{m2} = -\frac{3M_{q0}}{4C_f} + \frac{3\sigma L_{sm}I_{sqm0}}{4\tau_r C_f V_{f0}} + \frac{3\sigma L_{sm}I_{sdm0}\omega_{e0}}{4C_f V_{f0}}; a_{m3} = \frac{3(I_{sdm0}V_{sdm0} + I_{sqm0}V_{sqm0})}{4C_f V_{f0}^2};$$

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$$C_{invu} = \begin{bmatrix} a_{in1} & a_{in2} & a_{in3} & a_{in4} & a_{in5} & a_{in6} & a_{in7} & a_{in8} & a_{in9} \end{bmatrix}_{1 \times 9}$$

$$B_{inv,invU} = \begin{bmatrix} 0 & 0 & 0 & 0 & \frac{1}{2\tau_{inf}} & 0 & 0 & 0 & 0 \end{bmatrix}_{1 \times 9}^T$$

$$a_{in1} = -\frac{3(V_{indm0} - k_{idp}I_{ind0} - \omega L_{inf}I_{inq0})}{4C_{inf}V_{inf0}}; a_{in2} = -\frac{3(V_{inqm0} - k_{iqp}I_{inq0} - \omega L_{inf}I_{ind0})}{4C_{inf}V_{inf0}};$$

$$a_{in3} = -\frac{3I_{ind0}(1 - k_{idp}k_{vdp})}{4C_{inf}V_{inf0}}; a_{in4} = -\frac{3I_{inq0}(1 - k_{iqp}k_{vqp})}{4C_{inf}V_{inf0}};$$

$$a_{in5} = \frac{3(V_{indm0}I_{ind0} + V_{inqm0}I_{inq0})}{4C_{inf}V_{inf0}^2}; a_{in6} = -\frac{3I_{ind0}k_{idp}k_{vdi}}{4C_{inf}V_{inf0}};$$

$$a_{in7} = -\frac{3I_{ind0}k_{idi}}{4C_{inf}V_{inf0}}; a_{in8} = -\frac{3I_{inq0}k_{iqp}k_{vqi}}{4C_{inf}V_{inf0}}; a_{in9} = -\frac{3I_{inq0}k_{iqi}}{4C_{inf}V_{inf0}}$$

$$A_{inv} = \begin{bmatrix} -\frac{R_{inf}}{L_{inf}} & \omega & -\frac{1+k_{idp}k_{vdp}}{L_{inf}} & 0 & 0 & k_{idp}k_{vdi} & k_{idi} & 0 & 0 \\ -\omega & -\frac{R_{inf}}{L_{inf}} & 0 & -\frac{1+k_{iqp}k_{vqp}}{L_{inf}} & 0 & 0 & 0 & k_{iqp}k_{vqi} & k_{iqi} \\ \frac{1}{C_{inf}} & 0 & -\frac{1}{C_{inf}R_{LVAC}} & \omega & 0 & 0 & 0 & 0 & 0 \\ 0 & \frac{1}{C_{inf}} & -\omega & -\frac{1}{C_{inf}R_{LVAC}} & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & -\frac{1}{\tau_{inf}} & 0 & 0 & 0 & 0 \\ 0 & 0 & -1 & 0 & 0 & 0 & 0 & 0 & 0 \\ -1 & 0 & -k_{vdp} & 0 & 0 & k_{vdi} & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 0 & 0 & 0 & 0 \\ 0 & -1 & 0 & -k_{vqp} & 0 & 0 & 0 & k_{vqi} & 0 \end{bmatrix}_{9 \times 9}$$

$$A_{fl} = \begin{bmatrix} -\frac{R_{fl}}{L_{fl}} & \omega & -\frac{1}{L_{fl}} & 0 \\ -\omega & -\frac{R_{fl}}{L_{fl}} & 0 & -\frac{1}{L_{fl}} \\ \frac{1}{C_{fl}} & 0 & 0 & \omega \\ 0 & \frac{1}{C_{fl}} & -\omega & 0 \end{bmatrix}_{6 \times 6}$$

$$B_{fl,flU} = \begin{bmatrix} \frac{1}{L_{lf}} & 0 & 0 & 0 \\ 0 & \frac{1}{L_{lf}} & 0 & 0 \end{bmatrix}_{2 \times 4}^T; B_{fl,flD} = \begin{bmatrix} 0 & 0 & -\frac{1}{C_{fl}} & 0 \\ 0 & 0 & 0 & -\frac{1}{C_{fl}} \end{bmatrix}_{2 \times 4}^T;$$

$$C_{fltu} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}_{2 \times 4}; C_{fld} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}_{2 \times 4}$$

A.3 Parameters

1) Generator

$V_s=230\text{V}$, $f=400\text{Hz}$, $R_{gs}=0.0044\Omega$, $L_{ls}=19.89\mu\text{H}$, $L_{md}=220.164\mu\text{H}$, $L_{mq}=161.807\mu\text{H}$,
 $R_F=0.068884\Omega$, $L_{lf}=32.83\mu\text{H}$, $R_{kd}=0.142\Omega$, $L_{lkd}=34.079\mu\text{H}$, $R_{kq}=0.003095\Omega$,
 $L_{lkq}=144.274\mu\text{H}$, $k_{pgi}=0.0487$, $k_{igi}=99.88$, $k_{pgv}=1.78$, $k_{igv}=227.02$

2) ATRU/TRU:

$r_p=1\Omega$, $r_s=0.36\Omega$, $l_p=2.55\text{mH}$, $l_s=61.6\mu\text{H}$, $n=6.464$, $C_{eq1}=2\text{nF}$, $C_{eq2}=2\text{nF}$,
 $r_f=0.01\Omega$, $L_f=2\text{mH}$, $C_f=500\mu\text{F}$.

3) PWM rectifier:

$V_s=230\text{V}$, $\omega_e=400*2\pi(\text{rad/s})$, $R_{rf}=0.1\Omega$, $L_{rf}=1\text{mH}$, $C_{rf}=2.4\text{mF}$, $k_{rep}=0.0541$,
 $k_{rei}=2.43$, $k_{rdp}=0.403$, $k_{rdi}=986.96$, $k_{rqp}=0.403$, $k_{rqi}=986.96$, $V_{dc1*}=800\text{V}$, $\tau_{rf}=0.001$

4) ECS/EMA:

$\omega_r^*=975\text{rpm}$, $R_{sm}=0.6\Omega$, $L_{sm}=30.39\mu\text{H}$, $R_r=0.159\Omega$, $L_r=30.39\mu\text{H}$, $L_m=29.03\mu\text{H}$,
 $P=6$, $J_m=0.281\text{kgm}^2$, $I_{sdm}=34.115\text{A}$, $\tau_{f1}=0.02\text{s}$, $k_{pw}=13.27$, $k_{iw}=1042.33$, $k_{pim}=4.75$,
 $k_{iim}=4199.14$

5) Inverter:

$\tau_{f2}=0.05\text{s}$, $k_{pvd}=0.5$, $k_{ivd}=1000$, $k_{pvq}=0.5$, $k_{ivq}=1000$, $k_{pid}=10$, $k_{iid}=500$, $k_{piq}=10$,
 $k_{iiq}=500$, $r_{f1}=0.2\Omega$, $L_{f1}=2\text{mH}$, $C_{f1}=50\mu\text{F}$

6) Transmission lines:

$R_{fl}=0.02\Omega$, $L_{fl}=100\mu\text{H}$, $C_{fl}=2\text{nF}$

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Appendix B. Preliminaries for Global Stabilization

Preliminaries utilized in the Chapter 6 during the derivative procedure in Section 5.3.3 are stated as follows:

Consider the following system

$$\dot{x} = f(t, x, u) \quad (\text{B.1})$$

where $f : [0, \infty) \times \mathbb{R}^n \times \mathbb{R}^m \rightarrow \mathbb{R}^n$ is a continuous function with $f(0) = 0$.

Definition 1: A continuous function $\alpha : [0, a) \rightarrow [0, \infty)$ belongs to class function if it is strictly increasing and $\alpha(0) = 0$. It belongs to class K_∞ functions if $\alpha = \infty$ and $\alpha(r) \rightarrow \infty$.

Lemma 1 [114]: Let $V : [0, \infty) \times \mathbb{R}^n \rightarrow \mathbb{R}$ be a continuous differentiable function such that

$$\alpha_1(\|x\|) \leq V(t, x) \leq \alpha_2(\|x\|) \quad (\text{B.2})$$

$$\frac{\partial V}{\partial t} + \frac{\partial V}{\partial x} f(t, x, u) \leq -\Gamma(x), \forall \|x\| \geq \rho(\|u\|) > 0 \quad (\text{B.3})$$

$\forall (t, x, u) \in [0, \infty) \times \mathbb{R}^n \times \mathbb{R}^m$, where α_1, α_2 are class K_∞ functions, ρ is a class K function, and $\Gamma(x)$ is a continuous positive definite function on $\mathbb{R}^n$. Then the system is globally input-to-state stable (ISS).

Lemma 2 [114]: If the system (B.1) is globally input-to-state stable and $\lim_{t \rightarrow \infty} u = 0$, then the system is also globally asymptotically stable.

Lemma 3 [115]: Let c, d be positive constants. Given any positive constant γ , the following inequality holds

$$|x|^c |y|^d \leq \frac{c}{c+d} \gamma^{c/d} |x|^{c+d} + \frac{d}{c+d} \gamma^{-c/d} |y|^{c+d} \quad (\text{B.4})$$

Author's Publications

Journal

- [J1] **Q. Xu**, J. Xiao, X. Hu, P. Wang, and M. Y. Lee, "A decentralized power management strategy for hybrid energy storage system with autonomous bus voltage restoration and state-of-charge recovery," *IEEE Transactions on Industrial Electronics*, vol. 64, no. 9, pp. 7098–7108, Sept 2017.
- [J2] **Q. Xu**, X. Hu, P. Wang, J. Xiao, P. Tu, C. Wen, and M. Y. Lee, "A decentralized dynamic power sharing strategy for hybrid energy storage system in autonomous dc microgrid," *IEEE Transactions on Industrial Electronics*, vol. 64, no. 7, pp. 5930–5941, July 2017.
- [J3] **Q. Xu**, P. Wang, J. Chen, C. Wen, and M. Y. Lee, "A module-based approach for stability analysis of complex more electric aircraft power system," *IEEE Transactions on Transportation Electrification*, vol. 3, no. 4, pp. 901–919, Dec. 2017.
- [J4] **Q. Xu**, C. Zhang, C. Wen, and P. Wang, "A novel composite nonlinear controller for stabilization of constant power load in dc microgrid," *IEEE Transactions on Smart Grid*, vol. PP, no. 99, pp. 1–1, 2017.
- [J5] **Q. Xu**, J. Xiao, P. Wang, X. Pan, and C. Wen, "A decentralized control strategy for autonomous transient power sharing and state-of-charge recovery in hybrid energy storage systems," *IEEE Transactions on Sustainable Energy*, vol. 8, no. 4, pp. 1443–1452, Oct. 2017.
- [J6] **Q. Xu**, J. Xiao, P. Wang, and C. Wen, "A decentralized control strategy for economic operation of autonomous ac, dc and hybrid ac/dc microgrids," *IEEE Transactions on Energy Conversion*, vol. 32, no. 4, pp. 1345–1355, Dec. 2017.

- [J7] X. Pan, A. Ghoshal, Y. Liu, **Q. Xu**, and A. K. Rathore, "Hybrid modulation based bidirectional electrolytic capacitor-less three-phase inverter for fuel cell vehicles: Analysis, design, and experimental results," *IEEE Transactions on Power Electronics*, vol. PP, no. 99, pp. 1–1, 2017.
- [J8] C. Dong, H. Jia, **Q. Xu**, J. Xiao, Y. Xu, P. Tu, P. Lin, X. Li, and P. Wang, "Time-delay stability analysis for hybrid energy storage system with hierarchical control in dc microgrids," *IEEE Transactions on Smart Grid*, vol. PP, no. 99, pp. 1–1, 2017.
- [J9] J. Xiao, P. Wang, L. Setyawan, and **Q. Xu**, "Multi-level energy management system for real-time scheduling of dc microgrids with multiple slack terminals," *IEEE Transactions on Energy Conversion*, vol. 31, no. 1, pp. 392–400, March 2016.

Conference

- [C1] **Q. Xu**, P. Wang, and T. Zhao, "Optimal Operation of Battery Swapping-charging Systems Considering Quality-of-service Constraints," in *2017 IEEE PES general meeting(PESGM)*, July 2017, pp. 1–5.
- [C2] **Q. Xu**, C. Zhang, C. Wen, P. Wang, and L. M. Yeong, "A novel adaptive backstepping controller for stabilization of dc/dc converter feeding constant power load," in *2017 IEEE 26th International Symposium on Industrial Electronics (ISIE)*, June 2017, pp. 570–575.
- [C3] **Q. Xu**, P. Tu, C. Dong, C. Qi, S. Yang, S. Yao, and P. Wang, "Pof based reliability prediction for cascaded h-bridge converter in drive application," in *2017 IEEE 3rd International Future Energy Electronics Conference and ECCE Asia (IFEEEC 2017 - ECCE Asia)*, June 2017, pp. 1759–1764.
- [C4] **Q. Xu**, P. Wang, Y. Zhang, C. Wen, and J. Xiao, "A decentralized control strategy for economic operation of autonomous ac microgrids," in *IECON 2016 - 42nd Annual Conference of the IEEE Industrial Electronics Society*, Oct 2016, pp. 4020–4024.
- [C5] **Q. Xu**, X. Hu, P. Wang, J. Xiao, L. Setyawan, C. Wen, and L. M. Yeong, "Design and stability analysis for an autonomous dc microgrid with constant power load," in *2016 IEEE Applied Power Electronics Conference and Exposition (APEC)*, March 2016, pp. 3409–3415.

- [C6] **Q. Xu**, J. Chen, P. Wang, C. Wen, L. M. Yeong, and R. Simanjorang, "Investigation of the future mea power system architecture-from the perspective of system stability," in *8th IET International Conference on Power Electronics, Machines and Drives (PEMD 2016)*, April 2016, pp. 1–6.
- [C7] **Q. Xu**, P. Wang, J. Xiao, C. Wen, and L. M. Yeong, "Modeling and stability analysis of hybrid energy storage system under hierarchical control," in *2015 IEEE PES Asia-Pacific Power and Energy Engineering Conference (APPEEC)*, Nov 2015, pp. 1–5.
- [C8] S. Adhikari, **Q. Xu**, Y. Tang, and P. Wang, "Decentralized control of dc microgrid clusters," in *2017 IEEE 3rd International Future Energy Electronics Conference and ECCE Asia (IFEEC 2017 - ECCE Asia)*, June 2017, pp. 567–572.
- [C9] X. Pan, **Q. Xu**, T. Zhao, F. Yang, and P. Wang, "Design and implementation of communication network and system management for hybrid ac/dc microgrid module," in *2017 IEEE 3rd International Future Energy Electronics Conference and ECCE Asia (IFEEC 2017 - ECCE Asia)*, June 2017, pp. 1260–1265.
- [C10] P. Lin, P. Wang, **Q. Xu**, J. Xiao, I. U. Nutkani, and C. F. Hoong, "An integral-droop based dynamic power sharing control for hybrid energy storage system in dc microgrid," in *2017 IEEE 3rd International Future Energy Electronics Conference and ECCE Asia (IFEEC 2017 - ECCE Asia)*, June 2017, pp. 338–343.
- [C11] P. Tu, S. Yang, P. Wang, **Q. Xu**, and C. Qi, "Analytical averaged loss model of half-bridge modular multilevel converters," in *2017 IEEE 3rd International Future Energy Electronics Conference and ECCE Asia (IFEEC 2017 - ECCE Asia)*, June 2017, pp. 1965–1970.
- [C12] J. Xiao, L. Zhang, P. Wang, and **Q. Xu**, "Distributed coordination for economical operation of power park with multiple microgrids," in *2016 Asian Conference on Energy, Power and Transportation Electrification (ACEPT)*, Oct 2016, pp. 1–6.
- [C13] P. Lin, **Q. Xu**, P. Wang, J. Xiao, and H. Li, "Improved virtual capacitive droop control for hybridization of energy storages in dc microgrid,"

AUTHOR'S PUBLICATIONS

in *2016 IEEE 2nd Annual Southern Power Electronics Conference (SPEC)*,
Dec 2016, pp. 1–5.

Bibliography

- [1] “International Energy Outlook 2016,” [Online], Available: <https://www.eia.gov/outlooks/ieo/>.
- [2] “Renewables 2017 Global Status Report,” [Online], Available: http://www.ren21.net/wp-content/uploads/2017/06/17-8399_GSR_2017_Full_Report_0621_Opt.pdf.
- [3] “Sustainable energy for all- global tracking framework 2017: Progress toward sustainable energy ,” [Online], Available: <http://www.se4all.org/sites/default/files/GTF%20Executive%20Summary%202017.pdf>.
- [4] “Climate Vulnerable Forum: World’s vulnerable open gateway to climate safe future at Paris,” [Online], Available: <http://www.thecvf.org/wp-content/uploads/2015/11/HighLevel-Meeting-1.pdf>.
- [5] “European Commission, Energy roadmap 2050,” [Online], Available: https://ec.europa.eu/energy/sites/ener/files/documents/2012_energy_roadmap_2050_en_0.pdf.
- [6] “The National Renewable Energy Laboratory’s (NREL) Renewable Electricity Futures Study (RE Futures),” [Online], Available: https://www.nrel.gov/analysis/re_futures/.
- [7] A. M. Annaswamy, *IEEE Vision for Smart Grid Controls: 2030 and Beyond: Roadmap*. IEEE CSS, 2013.
- [8] M. H. Bollen, “The smart grid: Adapting the power system to new challenges,” *Synthesis Lectures on Power Electronics*, vol. 2, no. 1, pp. 1–180, 2011.
- [9] N. Hatziaargyriou *et al.*, *Microgrids: architectures and control*. Wiley Online Library, 2014.

- [10] R. Lasseter, A. Akhil, C. Marnay, J. Stephens, J. Dagle, R. Guttromson, A. Meliopoulos, R. Yinger, and J. Eto, "The certs microgrid concept," *White paper for Transmission Reliability Program, Office of Power Technologies, US Department of Energy*, vol. 2, no. 3, p. 30, 2002.
- [11] M. Sánchez, "Overview of microgrid research and development activities in the eu," in *Proc. the 2006 Symposium on Microgrids*, 2006.
- [12] X. Wu, X. Yin, Q. Wei, Y. Jia, and J. Wang, "Research on microgrid and its application in china," *Energy and Power Engineering*, vol. 5, no. 04, p. 171, 2013.
- [13] "Pulau Ubin Micro-grid Test-Bed," [Online], Available: https://www.ema.gov.sg/Pulau_Ubin_Micro-grid_Test_Bed.aspx.
- [14] X. Liu, P. Wang, and P. C. Loh, "A hybrid ac/dc microgrid and its co-ordination control," *IEEE Transactions on Smart Grid*, vol. 2, no. 2, pp. 278–286, 2011.
- [15] J. M. Guerrero, P. C. Loh, T.-L. Lee, and M. Chandorkar, "Advanced control architectures for intelligent microgrids-part ii: Power quality, energy storage, and ac/dc microgrids," *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1263–1270, 2013.
- [16] J. Rocabert, A. Luna, F. Blaabjerg, and P. Rodríguez, "Control of Power Converters in AC Microgrids," *IEEE Transactions on Power Electronics*, vol. 27, no. 11, pp. 4734–4749, nov 2012. [Online]. Available: <http://ieeexplore.ieee.org/document/6200347/>
- [17] J. C. Vasquez, J. M. Guerrero, M. Savaghebi, J. Eloy-Garcia, and R. Teodorescu, "Modeling, Analysis, and Design of Stationary-Reference-Frame Droop-Controlled Parallel Three-Phase Voltage Source Inverters," *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1271–1280, apr 2013. [Online]. Available: <http://ieeexplore.ieee.org/document/6184306/>
- [18] F. Blaabjerg, R. Teodorescu, M. Liserre, and A. Timbus, "Overview of Control and Grid Synchronization for Distributed Power Generation Systems," *IEEE Transactions on Industrial Electronics*, vol. 53, no. 5, pp. 1398–1409, oct 2006. [Online]. Available:

<http://www.scopus.com/inward/record.url?eid=2-s2.0-33750109723{&}partnerID=40{&}md5=e3aa66bfcf33dec3da2c465d97cb873chttp://ieeexplore.ieee.org/document/1705631/>

- [19] H. Han, X. Hou, J. Yang, J. Wu, M. Su, and J. M. Guerrero, "Review of Power Sharing Control Strategies for Islanding Operation of AC Microgrids," *IEEE Transactions on Smart Grid*, vol. 7, no. 1, pp. 200–215, jan 2016. [Online]. Available: <http://ieeexplore.ieee.org/document/7121016/>
- [20] J. Xiao, P. Wang, and L. Setyawan, "Hierarchical control of hybrid energy storage system in dc microgrids," *IEEE Transactions on Industrial Electronics*, vol. 62, no. 8, pp. 4915–4924, 2015.
- [21] J. J. Justo, F. Mwasilu, J. Lee, and J.-W. Jung, "Ac-microgrids versus dc-microgrids with distributed energy resources: A review," *Renewable and Sustainable Energy Reviews*, vol. 24, pp. 387–405, 2013.
- [22] T. Dragičević, X. Lu, J. C. Vasquez, and J. M. Guerrero, "Dc microgrids-part i: A review of control strategies and stabilization techniques," *IEEE Transactions on power electronics*, vol. 31, no. 7, pp. 4876–4891, 2016.
- [23] —, "Dc microgrids-part ii: A review of power architectures, applications, and standardization issues," *IEEE transactions on power electronics*, vol. 31, no. 5, pp. 3528–3549, 2016.
- [24] P. C. Loh, D. Li, Y. K. Chai, and F. Blaabjerg, "Autonomous control of interlinking converter with energy storage in hybrid ac–dc microgrid," *IEEE Transactions on Industry Applications*, vol. 49, no. 3, pp. 1374–1382, 2013.
- [25] E. Unamuno and J. A. Barrena, "Hybrid ac/dc microgrids-part i: Review and classification of topologies," *Renewable and Sustainable Energy Reviews*, vol. 52, pp. 1251–1259, 2015.
- [26] N. Eghtedarpour and E. Farjah, "Power control and management in a hybrid ac/dc microgrid," *IEEE transactions on smart grid*, vol. 5, no. 3, pp. 1494–1505, 2014.
- [27] 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.
- [28] S. Parhizi, H. Lotfi, A. Khodaei, and S. Bahramirad, “State of the art in research on microgrids: A review,” *IEEE Access*, vol. 3, pp. 890–925, 2015.
- [29] S. Vazquez, S. M. Lukic, E. Galvan, L. G. Franquelo, and J. M. Carrasco, “Energy storage systems for transport and grid applications,” *IEEE Transactions on Industrial Electronics*, vol. 57, no. 12, pp. 3881–3895, 2010.
- [30] M. Nehrir, C. Wang, K. Strunz, H. Aki, R. Ramakumar, J. Bing, Z. Miao, and Z. Salameh, “A review of hybrid renewable/alternative energy systems for electric power generation: Configurations, control, and applications,” *IEEE Transactions on Sustainable Energy*, vol. 2, no. 4, pp. 392–403, 2011.
- [31] W. Li, G. Joós, and J. Bélanger, “Real-time simulation of a wind turbine generator coupled with a battery supercapacitor energy storage system,” *IEEE Transactions on Industrial Electronics*, vol. 57, no. 4, pp. 1137–1145, 2010.
- [32] J. Shen and A. Khaligh, “A supervisory energy management control strategy in a battery/ultracapacitor hybrid energy storage system,” *IEEE Transactions on Transportation Electrification*, vol. 1, no. 3, pp. 223–231, 2015.
- [33] S. Dusmez and A. Khaligh, “A supervisory power-splitting approach for a new ultracapacitor–battery vehicle deploying two propulsion machines,” *IEEE Transactions on Industrial Informatics*, vol. 10, no. 3, pp. 1960–1971, 2014.
- [34] J. Deng, J. Shi, Y. Liu, and Y. Tang, “Application of a hybrid energy storage system in the fast charging station of electric vehicles,” *IET Generation, Transmission & Distribution*, vol. 10, no. 4, pp. 1092–1097, 2016.
- [35] H. Zhou, T. Bhattacharya, D. Tran, T. S. T. Siew, and A. M. Khambadkone, “Composite energy storage system involving battery and ultracapacitor with dynamic energy management in microgrid applications,” *IEEE transactions on power electronics*, vol. 26, no. 3, pp. 923–930, 2011.
- [36] S. Wang, Y. Tang, J. Shi, K. Gong, Y. Liu, L. Ren, and J. Li, “Design and advanced control strategies of a hybrid energy storage system for the grid

- integration of wind power generations,” *IET Renewable Power Generation*, vol. 9, no. 2, pp. 89–98, 2014.
- [37] R. A. Dougal, S. Liu, and R. E. White, “Power and life extension of battery-ultracapacitor hybrids,” *IEEE Transactions on components and packaging technologies*, vol. 25, no. 1, pp. 120–131, 2002.
- [38] A. Esmaili, B. Novakovic, A. Nasiri, and O. Abdel-Baqi, “A hybrid system of li-ion capacitors and flow battery for dynamic wind energy support,” *IEEE Transactions on Industry applications*, vol. 49, no. 4, pp. 1649–1657, 2013.
- [39] N. R. Tummuru, M. K. Mishra, and S. Srinivas, “Dynamic energy management of renewable grid integrated hybrid energy storage system,” *IEEE Transactions on Industrial Electronics*, vol. 62, no. 12, pp. 7728–7737, 2015.
- [40] Z. Chen, C. C. Mi, J. Xu, X. Gong, and C. You, “Energy management for a power-split plug-in hybrid electric vehicle based on dynamic programming and neural networks,” *IEEE Transactions on Vehicular Technology*, vol. 63, no. 4, pp. 1567–1580, 2014.
- [41] S. M. Lukic, J. Cao, R. C. Bansal, F. Rodriguez, and A. Emadi, “Energy storage systems for automotive applications,” *IEEE Transactions on industrial electronics*, vol. 55, no. 6, pp. 2258–2267, 2008.
- [42] F. Blaabjerg, Z. Chen, and S. B. Kjaer, “Power electronics as efficient interface in dispersed power generation systems,” *IEEE transactions on power electronics*, vol. 19, no. 5, pp. 1184–1194, 2004.
- [43] M. Wu and D. D.-C. Lu, “A novel stabilization method of lc input filter with constant power loads without load performance compromise in dc microgrids,” *IEEE Transactions on Industrial Electronics*, vol. 62, no. 7, pp. 4552–4562, 2015.
- [44] A. Emadi, A. Khaligh, C. H. Rivetta, and G. A. Williamson, “Constant power loads and negative impedance instability in automotive systems: definition, modeling, stability, and control of power electronic converters and motor drives,” *IEEE Transactions on Vehicular Technology*, vol. 55, no. 4, pp. 1112–1125, 2006.

- [45] A. Riccobono and E. Santi, "Comprehensive review of stability criteria for dc power distribution systems," *IEEE Transactions on Industry Applications*, vol. 50, no. 5, pp. 3525–3535, 2014.
- [46] X. Zhang, Q.-C. Zhong, and W.-L. Ming, "Stabilization of a cascaded dc converter system via adding a virtual adaptive parallel impedance to the input of the load converter," *IEEE Transactions on Power Electronics*, vol. 31, no. 3, pp. 1826–1832, 2016.
- [47] J. M. Guerrero, M. Chandorkar, T.-L. Lee, and P. C. Loh, "Advanced control architectures for intelligent microgrids-part i: Decentralized and hierarchical control," *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1254–1262, 2013.
- [48] J. M. Guerrero, J. C. Vasquez, J. Matas, L. G. de Vicuna, and M. Castilla, "Hierarchical Control of Droop-Controlled AC and DC Microgrids: A General Approach Toward Standardization," *IEEE Transactions on Industrial Electronics*, vol. 58, no. 1, pp. 158–172, 2011. [Online]. Available: <http://ieeexplore.ieee.org/lpdocs/epic03/wrapper.htm?arnumber=5546958>
- [49] A. Santucci, A. Sornioti, and C. Lekakou, "Power split strategies for hybrid energy storage systems for vehicular applications," *Journal of Power Sources*, vol. 258, pp. 395–407, 2014.
- [50] B. Hredzak, V. G. Agelidis, and M. Jang, "A model predictive control system for a hybrid battery-ultracapacitor power source," *IEEE Transactions on Power Electronics*, vol. 29, no. 3, pp. 1469–1479, 2014.
- [51] A. Mohamed, V. Salehi, and O. Mohammed, "Real-time energy management algorithm for mitigation of pulse loads in hybrid microgrids," *IEEE Transactions on Smart Grid*, vol. 3, no. 4, pp. 1911–1922, 2012.
- [52] O. Madani, A. Bhattacharjee, and T. Das, "Decentralized power management in a hybrid fuel cell ultracapacitor system," *IEEE Transactions on Control Systems Technology*, vol. 24, no. 3, pp. 765–778, 2016.
- [53] Y. Gu, W. Li, and X. He, "Frequency-coordinating virtual impedance for autonomous power management of dc microgrid," *IEEE Transactions on Power Electronics*, vol. 30, no. 4, pp. 2328–2337, 2015.

- [54] A. P. N. Tahim, D. J. Pagano, E. Lenz, and V. Stramosk, "Modeling and stability analysis of islanded dc microgrids under droop control," *IEEE Transactions on Power Electronics*, vol. 30, no. 8, pp. 4597–4607, 2015.
- [55] S. Augustine, M. K. Mishra, and N. Lakshminarasamma, "Adaptive droop control strategy for load sharing and circulating current minimization in low-voltage standalone dc microgrid," *IEEE Transactions on Sustainable Energy*, vol. 6, no. 1, pp. 132–141, 2015.
- [56] X. Lu, J. M. Guerrero, K. Sun, and J. C. Vasquez, "An improved droop control method for dc microgrids based on low bandwidth communication with dc bus voltage restoration and enhanced current sharing accuracy," *IEEE Transactions on Power Electronics*, vol. 29, no. 4, pp. 1800–1812, 2014.
- [57] X. Lu, K. Sun, J. M. Guerrero, J. C. Vasquez, and L. Huang, "State-of-charge balance using adaptive droop control for distributed energy storage systems in dc microgrid applications," *IEEE Transactions on Industrial electronics*, vol. 61, no. 6, pp. 2804–2815, 2014.
- [58] A. M. Rahimi and A. Emadi, "Active damping in dc/dc power electronic converters: A novel method to overcome the problems of constant power loads," *IEEE Transactions on Industrial Electronics*, vol. 56, no. 5, pp. 1428–1439, 2009.
- [59] Y. Zhang and Y. W. Li, "Energy management strategy for supercapacitor in droop-controlled dc microgrid using virtual impedance," *IEEE Transactions on Power Electronics*, vol. 32, no. 4, pp. 2704–2716, 2017.
- [60] J. C. V. N. L. Diaz, T. Dragicevic and J. M. Guerrero, "Intelligent distributed generation and storage units for dc microgrids - a new concept on cooperative control without communications beyond droop control," *IEEE Trans. Smart Grid*, vol. 5, no. 5, p. 2476–2485, 2014.
- [61] T. Dragičević, J. M. Guerrero, J. C. Vasquez, and D. Škrlec, "Supervisory control of an adaptive-droop regulated dc microgrid with battery management capability," *IEEE Transactions on Power Electronics*, vol. 29, no. 2, pp. 695–706, 2014.

- [62] Y. Gu, X. Xiang, W. Li, and X. He, "Mode-adaptive decentralized control for renewable dc microgrid with enhanced reliability and flexibility," *IEEE Transactions on Power Electronics*, vol. 29, no. 9, pp. 5072–5080, 2014.
- [63] D. Chen and L. Xu, "Autonomous dc voltage control of a dc microgrid with multiple slack terminals," *IEEE Transactions on Power Systems*, vol. 27, no. 4, pp. 1897–1905, 2012.
- [64] G. Xu, L. Xu, D. J. Morrow, and D. Chen, "Coordinated dc voltage control of wind turbine with embedded energy storage system," *IEEE Transactions on Energy Conversion*, vol. 27, no. 4, pp. 1036–1045, 2012.
- [65] F. Golnaraghi and B. Kuo, "Automatic control systems," *Complex Variables*, vol. 2, pp. 1–1, 2010.
- [66] 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.
- [67] A. Griffio and J. Wang, "Modeling and stability analysis of hybrid power systems for the more electric aircraft," *Electric Power Systems Research*, vol. 82, no. 1, pp. 59–67, 2012.
- [68] K. Areerak, S. Bozhko, G. Asher, L. De Lillo, and D. Thomas, "Stability study for a hybrid ac-dc more-electric aircraft power system," *IEEE Transactions on Aerospace and Electronic Systems*, vol. 48, no. 1, pp. 329–347, 2012.
- [69] H. Ebrahimi and H. El-Kishky, "A novel generalized state-space averaging (gssa) model for advanced aircraft electric power systems," *Energy Conversion and Management*, vol. 89, pp. 507–524, 2015.
- [70] X. Feng, J. Liu, and F. C. Lee, "Impedance specifications for stable dc distributed power systems," *IEEE Transactions on Power Electronics*, vol. 17, no. 2, pp. 157–162, 2002.
- [71] K. Rajashekara, "More electric aircraft trends [technology leaders]," *IEEE Electrification Magazine*, vol. 2, no. 4, pp. 4–39, 2014.
- [72] P. Wheeler and S. Bozhko, "The more electric aircraft: Technology and challenges." *IEEE Electrification Magazine*, vol. 2, no. 4, pp. 6–12, 2014.

- [73] B. Sarlioglu and C. T. Morris, "More electric aircraft: Review, challenges, and opportunities for commercial transport aircraft," *IEEE Transactions on Transportation Electrification*, vol. 1, no. 1, pp. 54–64, 2015.
- [74] A. Griffo and J. Wang, "Large signal stability analysis of 'more electric' aircraft power systems with constant power loads," *IEEE Transactions on Aerospace and Electronic Systems*, vol. 48, no. 1, pp. 477–489, 2012.
- [75] S. Even, *Graph algorithms*. Cambridge University Press, 2011.
- [76] D. Marx, P. Magne, B. Nahid-Mobarakeh, S. Pierfederici, and B. Davat, "Large signal stability analysis tools in dc power systems with constant power loads and variable power loads-a review," *IEEE Transactions on Power Electronics*, vol. 27, no. 4, pp. 1773–1787, 2012.
- [77] A. Abdel-Hafez, "Power generation and distribution system for a more electric aircraft-a review," in *Recent advances in aircraft technology*. InTech, 2012.
- [78] J. Chen, X. Zhang, and C. Wen, "Harmonics attenuation and power factor correction of a more electric aircraft power grid using active power filter," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 12, pp. 7310–7319, 2016.
- [79] M. Terorde and D. Schulz, "New real-time heuristics for electrical load rebalancing in aircraft," *IEEE Transactions on Aerospace and Electronic Systems*, vol. 52, no. 3, pp. 1120–1131, 2016.
- [80] M. K. Zadeh, R. Gavagsaz-Ghoachani, B. Nahid-Mobarakeh, S. Pierfederici, and M. Molinas, "Stability analysis of hybrid ac/dc power systems for more electric aircraft," in *Applied Power Electronics Conference and Exposition (APEC), 2016 IEEE*. IEEE, 2016, pp. 446–452.
- [81] F. Gao, X. Zheng, S. Bozhko, C. I. Hill, and G. Asher, "Modal analysis of a pmsg-based dc electrical power system in the more electric aircraft using eigenvalues sensitivity," *IEEE Transactions on Transportation Electrification*, vol. 1, no. 1, pp. 65–76, 2015.
- [82] Y. S. Shen, L. Xia, S. Bozhko, and G. Asher, "Stability analysis of aircraft electrical power system with active front rectifier system in generation

- channel,” in *Power Engineering Conference (UPEC), 2013 48th International Universities’*. IEEE, 2013, pp. 1–6.
- [83] M. K. Zadeh, R. Gavagsaz-Ghoachani, S. Pierfederici, B. Nahid-Mobarakeh, and M. Molinas, “Stability analysis and dynamic performance evaluation of a power electronics-based dc distribution system with active stabilizer,” *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 4, no. 1, pp. 93–102, 2016.
 - [84] R. Burgos, S. Rosado, F. Wang, D. Boroyevich, Z. Lewis, and K. Karimi, “Modeling considerations and stability analysis of aerospace power systems with hybrid ac/dc distribution,” SAE Technical Paper, Tech. Rep., 2006.
 - [85] T. Wu, S. Bozhko, G. Asher, and D. P. Thomas, “A fast dynamic phasor model of autotransformer rectifier unit for more electric aircraft,” in *Industrial Electronics, 2009. IECON’09. 35th Annual Conference of IEEE*. IEEE, 2009, pp. 2531–2536.
 - [86] K.-N. Areerak, S. Bozhko, G. Asher, and D. Thomas, “Stability analysis and modelling of ac-dc system with mixed load using dq-transformation method,” in *Industrial Electronics, 2008. ISIE 2008. IEEE International Symposium on*. IEEE, 2008, pp. 19–24.
 - [87] W. Ke-Xin and W. Shui-Ming, “Modeling and simulation of three-phase voltage source pwm rectifier,” in *Advanced Computer Theory and Engineering, 2008. ICACTE’08. International Conference on*. IEEE, 2008, pp. 982–986.
 - [88] K. Ause, P. Arek, S. D. Sudhoff, O. Wasynczuk *et al.*, “Analysis of electric machinery and drive systems,” 2013.
 - [89] T. Wu, S. Bozhko, G. Asher, P. Wheeler, and D. Thomas, “Fast reduced functional models of electromechanical actuators for more-electric aircraft power system study,” SAE Technical Paper, Tech. Rep., 2008.
 - [90] Y. Zhang, Z. Jiang, and X. Yu, “Small-signal modeling and analysis of parallel-connected voltage source inverters,” in *Power Electronics and Motion Control Conference, 2009. IPEMC’09. IEEE 6th International*. IEEE, 2009, pp. 377–383.

- [91] G. A. Whyatt and L. A. Chick, "Electrical generation for more-electric aircraft using solid oxide fuel cells," *Pacific Northwest National Laboratory, Richland, Washington, Tech. Rep. PNNL-21382*, 2012.
- [92] T. Dragičević, X. Lu, J. C. Vasquez, and J. M. Guerrero, "Dc microgrids-part i: A review of control strategies and stabilization techniques," *IEEE Transactions on Power Electronics*, vol. 31, no. 7, pp. 4876–4891, 2016.
- [93] J. Xiao, P. Wang, L. Setyawati, and Q. Xu, "Multi-level energy management system for real-time scheduling of dc microgrids with multiple slack terminals," *IEEE Transactions on Energy Conversion*, vol. 31, no. 1, pp. 392–400, 2016.
- [94] N. Bottrell, M. Prodanovic, and T. C. Green, "Dynamic stability of a microgrid with an active load," *IEEE Transactions on Power Electronics*, vol. 28, no. 11, pp. 5107–5119, 2013.
- [95] A. Kwasinski and C. N. Onwuchekwa, "Dynamic behavior and stabilization of dc microgrids with instantaneous constant-power loads," *IEEE Transactions on Power Electronics*, vol. 26, no. 3, pp. 822–834, 2011.
- [96] M. Cespedes, L. Xing, and J. Sun, "Constant-power load system stabilization by passive damping," *IEEE Transactions on Power Electronics*, vol. 26, no. 7, pp. 1832–1836, 2011.
- [97] P. Magne, D. Marx, B. Nahid-Mobarakeh, and S. Pierfederici, "Large-signal stabilization of a dc-link supplying a constant power load using a virtual capacitor: impact on the domain of attraction," *IEEE Transactions on Industry Applications*, vol. 48, no. 3, pp. 878–887, 2012.
- [98] A. M. Rahimi and A. Emadi, "Active damping in dc/dc power electronic converters: A novel method to overcome the problems of constant power loads," *IEEE Transactions on Industrial Electronics*, vol. 56, no. 5, pp. 1428–1439, 2009.
- [99] X. Zhang, X. Ruan, and Q.-C. Zhong, "Improving the stability of cascaded dc/dc converter systems via shaping the input impedance of the load converter with a parallel or series virtual impedance," *IEEE Transactions on Industrial Electronics*, vol. 62, no. 12, pp. 7499–7512, 2015.

- [100] J. Neely, S. Pekarek, R. DeCarlo, and N. Vaks, "Real-time hybrid model predictive control of a boost converter with constant power load," in *Applied Power Electronics Conference and Exposition (APEC), 2010 Twenty-Fifth Annual IEEE*. IEEE, 2010, pp. 480–490.
- [101] A. P. Tahim, D. J. Pagano, and E. Ponce, "Nonlinear control of dc-dc bidirectional converters in stand-alone dc microgrids," in *2012 IEEE 51st IEEE Conference on Decision and Control (CDC)*. IEEE, 2012, pp. 3068–3073.
- [102] Y. Zhao, W. Qiao, and D. Ha, "A sliding-mode duty-ratio controller for dc/dc buck converters with constant power loads," *IEEE transactions on industry applications*, vol. 50, no. 2, pp. 1448–1458, 2014.
- [103] J. Zeng, Z. Zhang, and W. Qiao, "An interconnection and damping assignment passivity-based controller for a dc–dc boost converter with a constant power load," *IEEE Transactions on Industry Applications*, vol. 50, no. 4, pp. 2314–2322, 2014.
- [104] J. Zhou and C. Wen, *Adaptive backstepping control of uncertain systems: Nonsmooth nonlinearities, interactions or time-variations*. Springer, 2008.
- [105] S. M. Ashabani and Y. A.-R. I. Mohamed, "A flexible control strategy for grid-connected and islanded microgrids with enhanced stability using nonlinear microgrid stabilizer," *IEEE Transactions on Smart Grid*, vol. 3, no. 3, pp. 1291–1301, 2012.
- [106] T.-S. Lee and M.-L. Chen, "Applied nonlinear control with adaptive backstepping technique for a three-phase ac/dc boost converter," in *Industrial Electronics, 2008. IECON 2008. 34th Annual Conference of IEEE*. IEEE, 2008, pp. 567–571.
- [107] H. El Fadil and F. Giri, "Backstepping based control of pwm dc-dc boost power converters," in *2007 IEEE International Symposium on Industrial Electronics*. IEEE, 2007, pp. 395–400.
- [108] W.-H. Chen, J. Yang, L. Guo, and S. Li, "Disturbance-observer-based control and related methods-an overview," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 2, pp. 1083–1095, 2016.

- [109] C. Zhang, J. Wang, S. Li, B. Wu, and C. Qian, "Robust control for pwm-based dc-dc buck power converters with uncertainty via sampled-data output feedback," *IEEE Transactions on Power Electronics*, vol. 30, no. 1, pp. 504–515, 2015.
- [110] S. Li, J. Yang, W.-H. Chen, and X. Chen, *Disturbance observer-based control: methods and applications*. CRC press, 2014.
- [111] H. SIRA-RAMIREZ and M. Ilic-Spong, "Exact linearization in switched-mode dc-to-dc power converters," *International journal of control*, vol. 50, no. 2, pp. 511–524, 1989.
- [112] S. Bacha, I. Munteanu, A. I. Bratcu *et al.*, "Power electronic converters modeling and control," *Advanced Textbooks in Control and Signal Processing.-London: Springer*, vol. 454, 2014.
- [113] Y. Song and B. Wang, "Survey on reliability of power electronic systems," *IEEE Transactions on Power Electronics*, vol. 28, no. 1, pp. 591–604, 2013.
- [114] H. K. Khalil and J. Grizzle, *Nonlinear systems*. Prentice hall New Jersey, 1996, vol. 3.
- [115] C. Qian and W. Lin, "A continuous feedback approach to global strong stabilization of nonlinear systems," *IEEE Transactions on Automatic Control*, vol. 46, no. 7, pp. 1061–1079, 2001.