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Advanced Medium-Voltage Bidirectional DC-DC Conversion Systems for Future Electric Energy Delivery and Management Systems

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ADVANCED MEDIUM-VOLTAGE BIDIRECTIONAL DC-DC CONVERSION SYSTEMS
FOR FUTURE ELECTRIC ENERGY DELIVERY AND MANAGEMENT SYSTEMS

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I dedicated this to my parents Changqin Fan and Yonggui Song

and

my wife Lina Zhao.
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TABLE OF CONTENTS

List of Tables ................................................................................................................................ vii
List of Figures .................................................................................................................................. viii
Abstract .......................................................................................................................................... xii

1. INTRODUCTION .........................................................................................................................1
   1.1 Research Background .............................................................................................................1
       1.1.1 Solid State Transformer for Future AC Grid Systems .........................................................1
       1.1.2 Medium Voltage DC Grid Systems ....................................................................................4
       1.1.3 Next Generation Electric Ship Power Systems .................................................................6
   1.2 Dissertation Proposal and Research Objective ......................................................................8

2. STATE-OF-THE-ART MV DC-DC CONVERTER TOPOLOGY ..................................................10
   2.1 New Semiconductor Technology Based MV DC-DC Converter .........................................10
   2.2 Multi-Level MV DC-DC Converter .......................................................................................12
   2.3 Input-Series-Output-Parallel Modular DC-DC Converter .....................................................15
       2.3.1 Topology of the Constituent Module of ISOP Modular DC-DC Converter ....................16
       2.3.2 Control Strategy of ISOP Modular DC-DC Converters .....................................................19
   2.4 Summary ...............................................................................................................................20

3. THE MODULAR MV BIDIRECTIONAL DC-DC CONVERSION SYSTEMS .........................21
   3.1 Systems Description ..............................................................................................................21
   3.2 High-Frequency High Efficiency DC-DC Converter Module Design ...................................24
       3.2.1 Topology Selection ............................................................................................................24
       3.2.2 High Frequency Transformer Design ..................................................................................28
       3.2.3 Power Loss Analysis ..........................................................................................................32
   3.3 Experimental Results ............................................................................................................34
   3.4 Summary ...............................................................................................................................37

4. THE PHASE-SHIFT DHB CONVERTER WITH HIGH EFFICIENCY OVER WIDE LOAD RANGE .................................................................38
   4.1 Introduction ...........................................................................................................................38
   4.2 Operation Principle of the DHB Converter with an Adaptive Inductor ............................39
   4.3 Adaptive Inductor Design ......................................................................................................43
   4.4 Improvement of the Phase-Shift DHB Converter with an Adaptive Inductor ..................44
   4.5 Experimental Results ............................................................................................................46
   4.6 Summary ...............................................................................................................................51

5. THE IDENTICALLY DISTRIBUTED CONTROL OF ISOP MODULAR DHB CONVERTER .................................................................................................................................52
   5.1 Introduction ...........................................................................................................................53
   5.2 The ISOP DHB Converter with Identically Distributed Control ........................................53
LIST OF TABLES

3.1 Operational Conditions Comparison between DHB and DAB .................................................27
3.2 Key Specifications and Circuit Parameters of DHB Converter .................................................34
4.1 Key Specifications and Circuit Parameters of Two DHB Converters ..........................................46
5.1 Commutation Inductance of the IPOP System ........................................................................71
LIST OF FIGURES

1.1 The Electric Grid Diagram of the FREEDM System ..............................................................2
1.2 Transformer Core Size with Respect to Frequency ...............................................................3
1.3 The Block Diagram of 3-Stage 20 kVA Single Phase SST .....................................................3
1.4 Offshore Wind Farms ............................................................................................................4
1.5 MV DC-DC Converter in the Wind Turbine Nacelle ...........................................................5
1.6 MVDC Collector Grid for Offshore Wind Farms ....................................................................5
1.7 Configuration of MVDC Grid ...............................................................................................6
1.8 NGIPS Power Generation Roadmap (NAVSEA 2007) ..........................................................6
1.9 Concept of a MVDC Power Distribution System ..................................................................7
2.1 Simple Model for Three Series-Connected IGBTs .................................................................11
2.2 Voltage Distribution across Three Series-Connected IGBTs ..............................................11
2.3 Multi-Level, NPC Converters ...............................................................................................13
2.4 Multi-Level, FC Converters .................................................................................................14
2.5 Three-Level NPC Isolated DC-DC Converter .......................................................................15
2.6 Block Diagram of an ISOP Modular DC-DC converter ........................................................16
2.7 Phase-Shifted Full-Bridge Converters as Constituent Modules of ISOP Modular DC-DC Converter .........................................................................................................................17
2.8 Forward DC-DC Converter ..................................................................................................17
2.9 TTF Converters as Constituent Modules of ISOP Modular DC-DC Converter ....................18
2.10 Push-Pull DC-DC Converter .............................................................................................18
3.1 The Single Phase SST in the FREEDM System at One Residential Home ......................22
3.2 Modular Dual-Half-Bridge Bidirectional DC-DC Converter ...............................................22
3.3 3-D Structure of the Proposed MDHB Converter ...............................................................23
3.4 Schematics of Phase-Shift Dual-Bridge Converter ...............................................................24
3.5 Phase-Shift Bidirectional DC-DC Converter ...................................................................25
3.6 Output Power with Respect to Phase-Shift Angle .................................................................26
3.7 Comparison between DAB and DHB ............................................................................27
3.8 The Planar Transformer .................................................................................................29
3.9 Transformer Core Loss with Respect to Frequency and Primary Voltage .......................29
3.10 Cross Section of Transformer Winding and Corresponding Plot of $J$ and $H$ Distribution 30
3.11 AC Resistance with Respect to Frequency ..................................................................31
3.12 Operation and Waveforms of $S_1$ ...............................................................................32
3.13 Power Loss Breakdown ...............................................................................................33
3.14 Photo of the Prototype ...............................................................................................34
3.15 Photo of the Experimental Setup ................................................................................35
3.16 Key Waveforms of Single DHB Module at 1 kW ..........................................................35
3.17 Key Waveforms of Single DHB Module at 300 W .......................................................36
3.18 Measured Efficiency with Respect to Output Power .....................................................36
4.1 The Conventional DHB Converter ................................................................................39
4.2 The Proposed DHB Converter with an Adaptive Inductor .........................................40
4.3 Key Ideal Waveforms of the Phase-Shift DHB Converter ..........................................40
4.4 The Circulating Energy of the Conventional DHB Converter with an Optimized Fixed Inductor ..................................................................................................................42
4.5 The Adaptive Inductor ...............................................................................................44
4.6 Inductance as Function of DC Bias Current ...................................................................45
4.7 Phase Shift with Respect to Output Power .................................................................45
4.8 The Current Stress of the DHB Converters ...............................................................46
4.9 The Circulating Energy of the DHB Converters .......................................................46
4.10 The Photos of Prototype .........................................................................................47
4.11 Key Waveforms of Two DHB Converters at 300 W ..............................................49
4.12 Key Waveforms of Two DHB Converters at 1 kW ...............................................50
4.13 Measured Efficiency of Two DHB Converters ......................................................51

5.1 Block Diagram of the Proposed ISOP Converter with Identically Distributed Control .....53
5.2 The Schematic of the Constituent DHB Module ......................................................54
5.3 The Primary-Referred Equivalent Circuit of the Individual DHB Converter ..........54
5.4 The Ideal Operation Waveforms of the DHB Converter .......................................55
5.5 Large-Signal Average Model of the DHB Converter .............................................58
5.6 The Large-Signal Average Model of the ISOP DHB Converter ................................59
5.7 The Proposed ISOP DHB Converter with Identically Distributed Control ............61
5.8 IPOP Conversion System Consisting of Three DHB Converters ..............................62
5.9 The Output Currents of Three DHB Converters Connected in IPOP without Additional Control ...........................................................................................................63
5.10 IPOP Conversion System Employing ISOP DHB Converter as Constituent Subsystem ....63
5.11 Large-Signal Average Model of ISOP DHB Converter Adopting Common-Duty-Ratio Control ..............................................................................................................65
5.12 Simulation Waveforms of the ISOP DHB Converter Adopting Common-Duty-Ratio Control ..............................................................................................................66
5.13 Large Signal Average Model IPOP DHB Converter with Common-Duty-Control......67
5.14 Simulation Waveforms of Output Voltage and Output Currents of IPOP DHB Converter with Common-Duty-Control ........................................................................68
5.15 Input Voltage and Output Current Sharing of ISOP DHB Converter during Start Up ........69
5.16 Response to Load Transient of ISOP DHB Converter ..........................................................70
5.17 IPOP Conversion System Consisting of Three ISOP DHB Converters ..................71
5.18 Total Output Currents of Three ISOP Converters of the IPOP Conversion System ........72
5.19 Current Sharing inside Each ISOP DHB Converter of IPOP Conversion System ........72
5.20 Voltage Sharing inside Each ISOP DHB Converter of IPOP Conversion System ........73
5.21 Input-Voltage and Output-Current Sharing during Start Up .................................................74
5.22 Transformer Currents under Steady State .............................................................................74
5.23 Input Voltages during Steady State .......................................................................................75
5.24 Photo of the ISOP DHB Converter Consisting of Three DHB Modules .........................75
ABSTRACT

The distributed renewable energy generation and utilization are constantly growing, and are expected to be integrated with the conventional grid. The growing pressure for innovative solutions will demand power electronics to take an even larger role in future electric energy delivery and management systems, since power electronics are required for the conversion and control of electric energy by most dispersed generation systems. Furthermore, power electronics systems can provide additional intelligent energy management, grid stability, and power quality capabilities. Medium-voltage isolated dc-dc converter will become one of the key interfaces for grid components with moderate power ratings.

To address the demand of medium voltage (MV) and high power capability for future electric energy delivery and management systems, the power electronics community and industry have been reacting in two different ways: developing semiconductor technology or directly connecting devices in series/parallel to reach higher nominal voltages and currents while maintaining conventional converter topologies; and by developing new converter topologies with traditional semiconductor technology, known as multilevel converters or modular converters.

The modular approach uses the well-known, mature, and cheaper power semiconductor devices by adopting new converter topologies. The main advantages of the modular approach include: significant improvement in reliability by introducing desired level of redundancy; standardization of components leading to reduction in manufacturing cost and time; power systems can be easily reconfigured to support varying input-output specifications; and possibly higher efficiency and power density of the overall system. Input-series output-parallel (ISOP) modular configuration is a good choice to realize MV to low voltage (LV) conversion for utility application. However, challenges still remain.

First of all, for the high-frequency MV utility application, the low switching loss and conduction loss are must-haves for high efficiency, while bidirectional power flow capability is a must for power management requirement. To address the demand, the phase-shift dual-half-bridge (DHB) is proposed as the constituent module of ISOP configuration for MV application. The proposed ISOP DHB converter employs zero-voltage-switching (ZVS) technique combined with LV MOSFETs to achieve low switching and conduction losses under high frequency
operation, and therefore high efficiency and high power density, and bidirectional power flow as well.

Secondly, a large load range of high efficiency is desired rather than only a specific load point due to the continuous operation and large load variation range of utility application, which is of high importance because of the rising energy cost. This work proposes a novel DHB converter with an adaptive commutation inductor. By utilizing an adaptive inductor as the main energy transfer element, the output power can be controlled by not only the phase shift but also the commutation inductance, which allows the circulating energy to be optimized for different load conditions to maintain ZVS under light load conditions and minimize additional conduction losses under heavy load conditions as well. As a result, the efficiency at both light and heavy load can be significantly improved compared with the conventional DHB converter, and therefore extended high-efficiency range can be achieved. In addition, current stress of switch devices can be reduced. The theoretical analysis is presented and validated by the experimental results on a 50 kHz, 1 kW dc-dc converter module.

Thirdly, input-voltage sharing and output-current sharing are critical to assure the advantages of the ISOP modular configuration. To solve this issue, an identically distributed control scheme is proposed in this work. The proposed control scheme, using only one distributed voltage loop to realize both input-voltage and output-current sharing, provides plug-and-play capability, possible high-level fault tolerance, and easy implementation. Another unique advantage of the proposed ISOP DHB converter is the power rating can be easily extended further by directly connecting multiple ISOP DHB converters in input-parallel-output-parallel (IPOP) while no additional control is needed. The proposed control scheme is elaborated using the large-signal average model. Further, the stability of the control schemes is analyzed in terms of the constituent modules’ topology as well as the configuration, and then an important fact that the stability of control scheme depends on not only the configuration but also the constituent module topology is first revealed in this work. Finally, the simulation and experimental results of an ISOP DHB converter consisting of three modules are presented to verify the proposed control scheme and the high frequency high efficiency operation.
CHAPTER ONE

INTRODUCTION

1.1 Research Background

In order to reduce the CO\textsubscript{2} emissions and the resulting impact of climate change and address the energy crisis, the worldwide interest in renewable energy sources has risen significantly over the past decades. As a result, the distributed renewable energy generation and utilization are constantly growing, and are expected to be integrated with the electric energy delivery and management systems. Consequently, the share of decentralized power systems in the electricity infrastructure has increased considerably. Most dispersed generation systems require power electronics for the conversion and control of electric energy. Furthermore, power electronics systems can provide additional intelligent energy management, grid stability and power quality capabilities. Thus, power electronics will play a key role in this paradigm shift to more renewable electric energy and more intelligence in future electricity distribution and management systems.

1.1.1 Solid State Transformer for Future AC Grid Systems

To develop technology to revolutionize the nation's power grid and speed renewable electric-energy technologies into every home and business, the Future Renewable Electric Energy Delivery and Management (FREEDM) Systems Center was established in 2008 by the national science foundation (NSF). Fig. 1.1 is the electric grid diagram showing the key elements of the FREEDM System. Other than today's electric energy system based on one-hundred year old centralized generation, transmission and distribution model, the FREEDM System will provide the users the ability to plug-and-generate, plug-and-store energies at home and in factories, as well as the ability to manage the energy consumption.
High-voltage high-efficiency power electronics based solid state transformer (SST) is one of the key enabling technologies to achieve the mentioned paradigm shift in energy delivery and management system. SST is intended to replace the conventional line-frequency (50/60Hz) transformer based on iron/steel cores and copper/aluminum coil [1-6]. SST can achieve high power density, low weight, and low volume with good power quality. Fig. 1.2 shows the Transformer core size with respect to frequency, and it can be seen that the transformer core size can be dramatically reduced when increasing the frequency from line frequency level to kHz order.

Various configurations for SST were reported in [1-6], of which the ac-dc-dc-ac configuration is more promising due to its advantages to provide power factor correction, reactive power compensation, and an additional dc bus. Fig.1.3 shows the block diagram of 3-stage 20 kVA single phase SST consisting of an ac-dc rectifier, an isolated dc-dc converter, and a dc-ac inverter. The ac-dc rectifier interfacing with the 7.2 kV electric utility grid is to provide power factor correction function while converting 7.2 kV ac to 12 kV dc. The dc-dc converter provides high frequency galvanic isolation and converts 12 kV high voltage dc to 400 V low
voltage dc as well. The 400 V dc is then converted to 120/240 V low-voltage ac for end-use application through a dc-ac inverter.

![Core Size vs Frequency Graph](image)

**Fig. 1.2.** Transformer core size with respect to frequency.

The dc-dc converter is the key element of SST, since it not only provides high frequency galvanic isolation, but also determines the overall efficiency and power density of SST. Some of the major challenges of the isolated bidirectional dc-dc converter are how to deal with MV and high power, and how to achieve high efficiency under high frequency operation.

![Block Diagram of 3-stage 20 kVA single phase SST](image)

**Fig. 1.3.** The block diagram of 3-stage 20 kVA single phase SST.
1.1.2 Medium Voltage DC Grid Systems

The most important options for the future are the renewable energies to generate clean electric power. Especially the utilization of wind power has proven to be already a mature and a global option. In general it can be expected, that the investment in wind power generation plants will continue to grow considerably and that by 2030 worldwide at least 20% of the power will be generated by wind power. [7]. Compared with onshore wind farms, the offshore wind farm shown in Fig. 1.4 provides more advantages such as less visual impact, higher and more constant wind velocity, and closeness to population centers. [8]

Fig. 1.4. Offshore wind farms.

It has been proposed to use MVDC connections rather than conventional AC links in order to obtain weight savings, higher efficiency, better behavior of the wind farm during grid faults and perturbations, better wind turbine control, and the possibility of implementing reactive power control. [9-11]. Fig. 1.5 and Fig. 1.6 show wind turbine nacelle with generator and power electronics converters and a possible configuration of MVDC collector grid for offshore wind farms, respectively, where the MV DC-DC converter is the key component [12-13].
Medium- and low voltage DC technology has the potential to substantially improve future power distribution systems [14-19]. In the future, a paradigm shift to an MVDC infrastructure might become the overall more economical and ecological response to the changing demands of modern power generation, storage and consumption. Fig. 1.7 shows possible configuration of future MVDC grid [17], where the MV high power DC-DC converter is vital to its success.
1.1.3 Next Generation Electric Ship Power Systems

Increased demand for electric power in projected future naval combatants with electric drive, advanced sensors, and electrically actuated weapons and launch systems motivates the consideration of new power systems architectures. This consideration includes potentially extensive use of DC power distribution, i.e., a MVDC power system [20]. Development of MVDC power systems for naval combatants is posited to provide many operational benefits, and will entail technological risk. In 2007, the Navy produced a Next Generation Integrated Power System (NGIPS) Technology Development Roadmap shown in Fig. 1.8.
The long term goal is to establish the Navy’s goal of incorporating a MVDC Integrated Power System (IPS) in future surface combatants and submarines. The application of MVDC however, requires the development and standardization of new ways to manage power, assure system stability, detect faults, and isolate faults. New equipment must also be developed and qualified for ship use. Once developed, MVDC promises affordable high power dense power generation systems.

Fig. 1.9 shows the concept of future MVDC power distribution system [21], where the DC-DC converters with high frequency transformer will be used to connect MVDC bus with energy storage systems such as batteries and fuel cells. The isolated DC-DC converters should have MV high power capability and bidirectional power flow capability, and are expected to achieve high efficiency and high power density.
1.2 Dissertation Proposal and Research Objective

As discussed above, the growing pressure for innovative solutions will demand power electronics to take an even larger role in future electric energy delivery and management systems. MV DC-DC converters will become key interfaces for grid components with moderate power ratings. Direct connection to medium voltage levels requires the intelligent use of advanced topologies combined with proper power semiconductor devices.

The main objective of this research is to develop an advanced DC-DC conversion system based on current commercially available power semiconductor devices to fulfill the requirements of future electric energy delivery and management systems as follows.

- Medium voltage capability
- High power capability
- High efficiency and high power density
- High frequency transformer galvanic isolation
- Bidirectional power flow
- Fault tolerance

In order to achieve the outlined goals, the dissertation proposes the following approach:

1. Proposal of input-series output-parallel modular structure combined with phase-shift ZVS technology to obtain
   a. Low-voltage MOSFET to achieve low conduction losses
   b. ZVS technology to achieve low switching losses with high frequency operation
   c. Planar transformer design to achieve low transformer loss as well as solid isolation and therefore enhanced reliability
   d. Phase-shift dual bridge topology to achieve seamless bidirectional power flow control

2. Proposal of novel adaptive commutation inductor as energy transfer element of phase-shift dual-bridge converter to achieve
   a. Reduced conduction losses due to circulating energy at heavy load
   b. Extended ZVS operation range and therefore reduced switching losses at light load
c. Minimized current stress of power devices
d. And therefore high efficiency over extended load range

3. Development of large-signal average model and proposal of identically distributed control strategy for ISOP phase-shift dual-bridge converter to realize
   a. Evenly shared currents and voltages among multiple constitute modules
   b. Possible fault tolerance by redundancy design
   c. Extendable power rating by connecting ISOP converter in IPOP without additional control

Though the proposed technology, as well as the analysis presented here, is valid for a general MV electric power system, to be more specific, the elaboration, design, and simulation and experimental verification will be presented based on SST application.
CHAPTER TWO

STATE-OF-THE-ART OF MV DC-DC CONVERTER TECHNOLOGY

To address the demand of medium voltage and high power capability for future electric energy delivery and management systems, the power electronics community and industry have been reacting in two different ways: developing semiconductor technology or directly connecting devices in series/parallel to reach higher nominal voltages and currents [22-26] while maintaining conventional converter topologies (mainly two-level converters); and by developing new converter topologies, with traditional semiconductor technology, known as multilevel converters [27-33], or modular converters [39-50].

2.1 New Semiconductor Technology Based MV DC-DC Converter

The MV DC-DC converter based on new power semiconductor device with higher nominal voltage and conventional converter topologies inherits the benefit of well-known circuit structures and control methods. However, the new power semiconductor devices are more expensive. In addition, the power semiconductor devices with higher voltage rating usually have very high switching losses when operated with the switching frequency of more than 20 kHz, and therefore fail to achieve high efficiency and high power density. Moreover, the semiconductor device voltage rating required by future electric energy transmission and distribution systems is much higher than what is currently commercially available.

Direct series connection of devices such as MOSFETs and IGBTs for MV applications has also been investigated [25-26]. Fig. 2.1 shows that due to real unequal and variable IGBT parameters (i.e. collector-emitter-capacitance, leakage current, switching delays and storage-times) and with real gate-drive-circuits for each single element (with unmatched delay-times and supply-voltages) the voltage-sharing of the whole switch is not symmetrical over all the elements (see Fig 2.2). The transient (during turn-on and turn-off, phases 1 and 3 in Fig 2.2) and static (in off-state, phase 4) divergent collector-emitter-voltages introduce a risk of destruction of the elements due to voltage- and power-dissipation-stress. Additional voltage-balancing control is
Fig. 2.1. Simple model for three series-connected IGBTs.

Fig. 2.2. Voltage distribution across three series connected IGBTs.
essentially needed to ensure that each element overtakes the correct amount of collector-emitter voltage in the transient and static phases. It is important to point out that the failure of each element will cause the failure of the entire converter, and consequently the failure rate of the converter will be increased due to the increased amount of power semiconductor devices connected in series.

2.2 Multi-Level MV DC-DC Converter

Multi-level MV DC-DC converter uses the well-known and cheaper power semiconductor devices, but the more complex circuit structures come along with several challenges for implementation and control. Multilevel converters are power-conversion systems composed by an array of power semiconductor devices and capacitive voltage sources that, when properly connected and controlled, can generate a multiple-step voltage waveform. The number of levels of a converter can be defined as the number of steps or constant voltage values that can be generated by the converter between the output terminal and any arbitrary internal reference node within the converter. Several multilevel converter circuit topologies have been proposed to overcome voltage sharing problem for series connected power semiconductor devices [27-32]. Among them, the neutral-point clamped (NPC), flying capacitor (FC) are the most studied. Fig. 2.3 and Fig. 2.4 show single-phase examples of NPC multilevel converters and FC multilevel converters, respectively. As shown in Fig. 2.5, the MV to LV isolated DC-DC converter can be formed by connecting point a and n of these multilevel converter to rectifier on low voltage side by transformer [33].

Multilevel converter technology can overcome voltage sharing problem for series connected devices. However, for those over than three-level converters, the NPC multilevel converter requires additional balance circuits, which usually is very complicated, to control the clamping capacitor voltage. Moreover, the number of diodes required will make the system impractical to implement. If the converter runs under PWM, the diode reverse recovery of these clamping diodes becomes the major design challenge in high-voltage, high-power applications. In contrast, the FC multilevel converter requires a large number of bulk capacitors to clamp the voltage. In addition, the energy stored in the clamping capacitors need be carefully addressed. Consequently, the system reliability cannot be guaranteed for a large quantity of diodes or flying capacitors [34-35].
Fig. 2.3. Multi-Level, NPC Converters: (a) Three-Level; (b) Five-Level.
Fig. 2.4. Multi-Level, FC Converters: (a) Three-Level; (b) Five-Level.
2.3 Input-Series-Output-Parallel Modular DC-DC Converter

Similar to multi-level converter, the modular approach uses the well-known, mature, and cheaper power semiconductor devices by adopting new converter topologies. The main advantages of the modular approach include: significant improvement in reliability by introducing desired level of redundancy [36-38]; standardization of components leading to reduction in manufacturing cost and time; power systems can be easily reconfigured to support varying input-output specifications; and possibly higher efficiency and power density of the overall system.

As shown in Fig. 2.6, the input-series-output-parallel (ISOP) modular DC-DC converter consists of multiple constituent DC-DC converter modules connected in series at the input and in parallel at the output, where each individual DC-DC converter module only need to handle a fraction of total input voltage at the input side and a fraction of total output current at output side, and therefore low voltage power semiconductor devices with low conduction loss and high switching speed can be used to obtain high power density and high conversion efficiency[48]. For this reason, the ISOP modular DC-DC converter is the most promising candidate for high power medium voltage to low voltage conversion system for future electric energy delivery and management systems.
2.3.1 Topology of the Constituent Module of ISOP Modular DC-DC Converter

Several ISOP modular DC-DC converters have been reported in [39-50], and much of previous work has focused on developing control techniques to achieve input voltage and output current sharing, stable operation, and better dynamic performance in terms of ISOP configuration. However, all the previous work paid little attention to the topology of the constituent module of ISOP modular DC-DC converter, and none of them has set MV power system as the target application. Although various topologies have been employed for the constituent modules of ISOP modular DC-DC converter in previous researches, unfortunately, none of them can meet the requirements of the future MV electric energy delivery and management systems. Phase-shift full-bridge converter presented in [40, 42, 44, 45], as shown in Fig. 2.7, is attractive for high power application, and can achieve ZVS operation for primary switches, but fails to provide bidirectional power flow path due to the rectifying diodes on secondary side. The forward converter in [41, 39] shown in Fig. 2.8 and two-transistor forward (TTF) converter in [50] shown in Fig. 2.9 uses less active switching devices, but cannot provide
high power capability and bidirectional power flow. The push-pull converter shown in Fig. 2.10 proposed as constituent module as ISOP modular converter for information technology equipment application in [48] succeeds to provide bidirectional power flow, and achieve high efficiency with high frequency operation with each module operated with less 10 volts input. However, the power loss will increase dramatically due to the hard switching and the resulted the switching loss when the input voltage is shifted to several hundred volts or even higher. In addition, the voltage stress of primary side switches in push-pull converter twice the input voltage. For these reasons, the push-pull converter is not a good choice for the MV power system application.

Fig. 2.7. Phase-shifted full-bridge converters as constituent modules of ISOP modular DC-DC converter.

Fig. 2.8. Forward DC-DC converter.
However, none of previous work has investigated the phase-shift dual-bridge bidirectional DC-DC converter as the constituent module of ISOP converter, although it appears to be the most promising candidate for future MV power system application due to its capability to achieve high-frequency transformer isolation, bidirectional power flow, and zero-voltage switching (ZVS) for all switching devices without auxiliary switch devices.
2.3.2 Control Strategy of ISOP Modular DC-DC Converters

For an ISOP modular DC-DC converter, voltage balance among the individual series inputs and current balance among parallel outputs are critical to ensure the advantages and stable operation of ISOP configuration as well. Several control schemes have been proposed to achieve the objective. A complicated charge control method with an input voltage feedforward for an ISOP system employing phase-shift full bridge as constituent module was proposed in [39]. The ISOP configuration in [41] uses three control loops with a common output voltage loop shared by all constituent modules. The approach in [44] resembles [41], with voltage mode controllers in place of current mode controllers. Uniform voltage distribution in ISOP converters is realized in [43] through input voltage distribution control. A cross feedback output current sharing control scheme for ISOP converters without any input voltage control loops was proposed in [50]. Common duty ratio scheme utilizing the inherent self-correcting mechanism of ISOP configuration when the modules operate at identical duty ratios is investigated in [30]. Similar to common duty ratio control, [48] presents a sensorless scheme for ISOP converters with well matched constituent modules. Component tolerances such as mismatches in transformers and inductors will distort the sharing slightly. A masterless ISOP scheme based on common duty ratio control was presented to achieve fault tolerance as well as stable operation in [49].

In most of the previous researches, the ISOP converters employ buck-derived converters as constituent modules, such as phase-shift full-bridge converter [40, 42, 44, 45], push-pull converter [48], forward converter [41, 39], and TTF converter, and based on which control schemes were developed to obtain input-voltage and output-current sharing. In spite of several advantages, such as high-frequency transformer isolation, bidirectional power flow, and ZVS for all switching devices without auxiliary switch devices, of the phase-shift dual-bridge converter [51-54], none of the previous work has investigated the ISOP phase-shift dual-bridge converter. In addition, no paper has clarified the effect of the constituent module topology on the control scheme stability while most paper discussed the stability of the control scheme only in terms of the converter configuration. Moreover, the reported control schemes developed for buck-derived converters may be invalid for the ISOP phase-shift dual-bridge converters, and even result in unstable operation.
2.4 Summary

In this chapter, two different ways to address the challenges of high power MV DC-DC converter were reviewed. One is to develop new power semiconductor devices while keeping traditional converter topology, the other is to develop new converter topologies while maintaining currently commercially available switching devices. Among these approaches, ISOP modular DC-DC converter seems to be the most promising one due to its unique advantages such as possibly high efficiency and high power density, enhanced reliability, reduced manufacturing cost; and the flexibility to be reconfigured to support varying input-output specifications.

Since most of previous work has focused on the control schemes in terms of the ISOP configuration itself only, more research efforts are needed to investigate the appropriate topology of the constituent module aiming at future high power MV electric energy conversion system, control scheme to achieve both fault-tolerance capability and current/voltage balance, and control stability in terms of the topology of the constituent module as well as the converter configuration.
CHAPTER THREE
THE MODULAR MV BIDIRECTIONAL DC-DC CONVERSION SYSTEMS

This chapter introduces the proposed MV bidirectional DC-DC conversion system, i.e., ISOP modular dual-half-bridge (MDHB) bidirectional dc-dc converter. Though the proposed technology, as well as the analysis presented here, is valid for a general MVDC utility application, the elaboration, design, simulation and experimental verification will be presented based on the application of SST, the key enabling technology to integrate highly distributed and scalable alternative generating sources in FREEDM systems. This chapter focuses on the detailed design of high-frequency, high-efficiency, and high-power-density constituent DC-DC converter modules of the ISOP modular converter. Topologies of the constituent module have been investigated for this application. A planar transformer adopting interleaved printed-circuit-board (PCB) winding is designed to realize reduced ac resistance, high voltage solid isolation, and identical parameters in multiple modules. The power loss of each main component for MDHB converter has been analyzed. Finally, the experimental results are presented to validate the analysis. The control scheme of the proposed MV bidirectional DC-DC conversion system will be presented in chapter 5.

3.1 Systems Description

Various configurations for SST were reported in [1-5], of which the ac-dc-dc-ac configuration has the advantage of providing power factor correction, reactive power, and an additional dc bus. The dc-dc conversion is a key element in the ac-dc-dc-ac SST configuration. Fig. 3.1 shows the block diagram of 3-stage 20 kVA single phase SST consisting of an ac-dc rectifier, an isolated dc-dc converter, and a dc-ac inverter. The ac-dc rectifier interfacing with the 7.2 kV electric utility grid is to provide power factor correction function while converting 7.2 kV ac to 12 kV dc. The dc-dc converter, the key stage of SST, provides high frequency galvanic...
isolation and converts 12 kV high voltage dc to 400 V low voltage dc as well. The 400 V dc is then converted to 120/240 V low-voltage ac for end-use application through a dc-ac inverter.

Fig. 3.1. The single phase SST in the FREEDM system at one residential home.

Fig. 3.2. Modular dual-half-bridge bidirectional dc-dc converter.
As shown in Fig. 3.2, the proposed MDHB converter for the dc-dc conversion stage consists of multiple low-voltage bidirectional dc-dc converter modules connected in input-series and output-parallel mode. The input and output voltage of each module are chosen as 500 V and 400 V respectively. Thus, the low-voltage commercial silicon MOSFETs with low conduction losses and high switching speed can be selected as the switching device. In order to interface with the 12 kV dc voltage from rectification stage, an 8-layer structure with 3 modules on each layer is shown in Fig. 3.3. Each module is a bidirectional dc-dc converter, which adopts phase-shift technique to realize ZVS operation mode for all switching devices without auxiliary switch devices in either direction of power flow [51, 52, and 54], and therefore enables the high switching frequency operation with low switching losses. Although a total of 24 modules will be used to interface high voltage, the utilization of low-voltage devices along with ZVS operation...
mode results in high efficiency, high frequency, good thermal performance and eventually high power density of the dc-dc conversion stage. As a result, the SST can achieve much smaller size than conventional line frequency (50/60 Hz) transformer by adopting the proposed high frequency high efficiency MDHB design.

3.2 High-Frequency High-Efficiency DC-DC Converter Module Design

3.2.1 Topology Selection

In spite of many advantages of SST, the basic function of SST is to realize high frequency galvanic isolation and bidirectional power flow to act as a real “transformer”, which is therefore the requirement for the dc-dc converter used for the isolated dc-dc stage of the SST. Phase-shift ZVS bidirectional dc-dc converters, such as dual-half-bridge (DHB) and dual-active-bridge (DAB) shown in Fig. 3.4, are attractive for this application.

(a)

(b)

Fig. 3.4. Schematics of phase-shift dual-bridge converter: (a) DHB, and (b) DAB.
Commutation inductor $L_1$, the sum of linkage inductance of the transformer and the external auxiliary inductance, is the main energy transfer element. The two half-bridges on both sides of $L_1$ can be simplified as two square-wave voltage sources with a phase shift angle $\varphi$, and
the principle waveforms are shown in Fig. 3.5. The energy will transferred from leading square-wave voltage source to the lagging one. Fig. 3.6 shows the relationship between the delivered power and the phase shift angle usually ranging from \(-\pi/2\) to \(\pi/2\). The energy is transferred from \(V_1\) to \(V_2\) when \(\varphi\) is between 0 and \(\pi/2\), and the power flow direction reverses when \(\varphi\) crosses zero and enters the range of 0 to \(-\pi/2\), so the direction change of power flow is a smooth process. In addition, the phase shift ZVS technique can realize ZVS for all switches without extra switch, which enables the high frequency operation and results in low transformer size.

![Output power with respect to phase-shift angle](image)

**Fig. 3.6. Output power with respect to phase-shift angle**

DAB and DHB are two popular topologies among phase-shift ZVS bidirectional dc-dc converters. Fig. 3.7 and Table 3-1 compare the operational conditions of DAB and DHB converters. Transformer flux swing of DHB topology is only half of DAB’s when same effective cross sectional area of the transformer are adopted at same switching frequency. The DHB converter achieves smaller transformer core loss. This will be described in detail in the following transformer design section. Moreover, A DHB’s use of half the number of switching devices as DAB, results in a more economical implementation especially in this multiple modules structure. The phase-shift DHB is therefore selected for the dc-dc converter module.
The output power of phase-shift DHB can be expressed as:

\[
P_{out1} = NV_{in1}V_{out} \phi (\pi - |\phi_1|) / 8\pi^2 fL_1
\]

where \(V_{in1}\) is input voltage, \(V_{out}\) is output voltage, \(f\) is the switching frequency, \(N\) is transformer turn ratio, \(L_1\) is the commutation inductance, and \(\phi_1\) is phase shift angle. Then, the output current can be given by

\[
i_{out1} = P_{out1} / V_{out} = NV_{in1} \phi (\pi - |\phi_1|) / 8\pi^2 fL_1
\]
From (3-2), it can be seen that the output current $I_{\text{out}}$ is independent of the output voltage $V_{\text{out}}$. This characteristic is similar to that of a current-source converter, which is an important advantage for this modular structure with parallel connection on output side. This feature will help the control design of the cascaded MDHB converter modules, and will be analyzed in detail in chapter 5.

### 3.2.2 High Frequency Transformer Design

Planar transformer with coils encapsulated within multi-layer PCB can achieve lower profile and higher power density than conventional wire-wound transformer especially for the multiple modules system MDHB. In addition, the windings of transformer are etched within the PCB and thus are completely repeatable; this can make the windings of the transformer identical in multiple modules and contribute to the balance among these modules. Furthermore, the planar Transformer utilizes solid insulation excluding air from the construction to minimize corona and partial discharge and therefore enhance reliability of SST. However, it is difficult to find a planar core suitable for this high voltage application requiring large cross sectional area. In this paper, a pair of PC40 PQ107/87/70 ferrite cores is modified to much lower profile while keeping the desired cross sectional area. After modification, the total window height of the transformer is reduced from 56 mm to 4.55 mm. The final transformer prototype is shown in Fig. 3.8, the primary to secondary turn ratio is 15:12, and the core loss can be calculated by the following empirical formula:

$$P_{\text{cl}} = V_e C_m f^x B_{\text{ac}}^y$$  \hfill (3-3)

where $V_e$ is effective core volume of transformer, $C_m$, $x$, and $y$ are coefficient related to core material, $B_{\text{ac}}$ is maximum flux density and can be expressed as:

$$B_{\text{ac}} = \frac{V_T D}{2 N_p A_f}$$  \hfill (3-4)

where $N_p$ is the primary number of turns, and $V_T$ is the applied voltage on the primary side of transformer. As shown in Fig. 3.7, $V_T$ equals to $V_m$ for DAB and $V_m/2$ for DHB. For 80°C, $C_m = 2.0$, $x=1.46$, $y = 2.57$. Fig. 3.9 shows the transformer core loss with respect to $V_T$ and $f$. The higher the frequency $f$ and the lower $V_T$, the lower the core loss is. For 50 kHz operation, core
loss of DAB with $V_T = 500\ \text{V}$ is 10.56 W, while the core loss of DHB with $V_T = 250$ is only 1.778 W, which verifies the analysis that DHB has much lower core loss than DAB.

Fig. 3.8. The planar transformer.

Fig. 3.9. Transformer core loss with respect to frequency and primary voltage.
Fig. 3.10. Cross section of transformer winding and corresponding plot of $J$ and $H$ distribution: (a) non-interleaved winding arrangement, (b) interleaved winding arrangement.
Both skin effect and proximity effect will increase high frequency copper losses in transformer winding [57-58], and therefore these effects must be taken into account when designing the transformer winding. PCB winding offers the flexibility to achieve the winding structure as desired. In this paper, 10-layer PCB with 2 oz copper is adopted for the transformer winding. Fig. 3.10 shows two different winding arrangements and their corresponding simulation results of current and magnetic field strength distribution in each layer. Other than the non-interleaved winding shown in Fig. 3.10 (a), a triple interleaved winding arrangement is utilized to optimize the magnetic field strength and therefore reduce ac resistance in this work. The maximum magnetic field strength of the interleaved winding is only 1000 A/m, while that of non-interleaved winding is up to 3200 A/m. Fig. 3.11 shows the total ac resistance of the transformer winding with respect to switching frequency. Compared with non-interleaved winding, the triple interleaved winding presented in this paper can achieve much lower high frequency ac resistance and therefore lower winding loss.

![AC resistance with respect to frequency.](Fig. 3.11)
3.2.3 Power Loss Analysis

Since the Phase-shift DHB converter realizes zero-voltage turn on for all switching devices, the turn-off losses and conduction losses of the switching devices are considered as the only power loss for the devices. Fig. 3.12 shows steady state waveforms of one switching cycle and zoomed switching waveforms of $S_1$.

![Waveform Diagram](Image)

The current of $S_1$ of one switching cycle can be expressed as follows:

\[
\begin{align*}
{i_S}_1 &= -\varphi V_{in} / 4\pi fL_u + (V_{in} / L_u) t & 0 \leq t < T_s (\varphi / 2 \pi) \\
{i_S}_1 &= \varphi V_{in} / 4\pi fL_u & T_s (\varphi / 2 \pi) \leq t \leq T_s / 2 \\
{i_S}_1 &= 0 & T_s / 2 < t < T_s
\end{align*}
\]

(3-5)

where $T_s$ is switching period, and the rms value of current through $S_1$ can be given as:

\[
I_{rms} = \sqrt{\frac{1}{T_s} \int_{0}^{T_s} (i_S)_1^2 \, dt} = (\varphi V_{in} / 4\pi fL_u) \sqrt{(3\pi - 2\varphi) / 6\pi}
\]

(3-6)

The conduction loss of $S_1$ can be calculated as:
\[ P_{\text{conduction}} = R_{\text{on}} (I_{\text{rms}})^2 = R_{\text{on}} (\varphi V_{\text{in}} / 4\pi fL_a)^2 (3\pi - 2\varphi) / 6\pi \] (3-7)

where \( R_{\text{on}} \) is on state resistance of \( S_1 \).

As shown in Fig. 3.12, at the moment \( S_1 \) is turned off, \( i_{L1} \), the current of \( L_a \), can be approximately considered as constant during the turn-off interval and will charge \( C_{o1} \), the output capacitor of \( S_1 \), and discharge \( C_{o2} \), the output capacitor of \( S_2 \). The turn off loss of \( S_1 \) therefore can be obtained as:

\[ P_{\text{off}} = \int_0^{t_f} v_{ds} i_{s1} \]
\[ = \int \left[i_{s1}(T_z/2)\right]^2 t_f^2 / 48c_{o1} = \frac{(\varphi V_{\text{in}} t_f)^2}{768\pi fL_a c_{o1}} \] (3-8)

where \( t_f \) is the fall time of \( S_1 \).

Power loss breakdown (W)

<table>
<thead>
<tr>
<th>Loss Type</th>
<th>Power Loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switching devices conduction loss</td>
<td>0.979</td>
</tr>
<tr>
<td>Switching devices turn-off loss</td>
<td>0.928</td>
</tr>
<tr>
<td>Transformer core loss</td>
<td>2</td>
</tr>
<tr>
<td>Transformer copper loss</td>
<td>7.482</td>
</tr>
<tr>
<td>Inductor core loss</td>
<td>14.397</td>
</tr>
<tr>
<td>Inductor copper loss</td>
<td>2.661</td>
</tr>
<tr>
<td>Others</td>
<td>1.992</td>
</tr>
</tbody>
</table>

Fig. 3.13. Power loss breakdown.

The loss calculation method of \( S_1 \) can apply to \( S_2 \), \( S_3 \), and \( S_4 \) since DHB is symmetrical. Core loss and copper loss of inductor \( L_a \) can be calculated using the method introduced in transformer design section. The loss breakdown for the 1kW dc-dc converter module is shown in Fig. 3.13.
3.3 Experimental Results

A 1 kW dc-dc converter module shown in Fig. 3.14 has been built and tested to verify the high frequency and high efficiency operation. Fig. 3.15 shows the experimental setup. The specifications and circuit parameters of a single dc-dc converter module are shown in Table 3-2.

<table>
<thead>
<tr>
<th>$V_{in}$ (V)</th>
<th>$V_{out}$ (V)</th>
<th>Rated Power (W)</th>
<th>Commutation Inductance (µH)</th>
<th>Primary MOSFET</th>
<th>Secondary MOSFET</th>
</tr>
</thead>
<tbody>
<tr>
<td>500</td>
<td>400</td>
<td>1000</td>
<td>90</td>
<td>IXFT24N80P</td>
<td>IXFT36N60P</td>
</tr>
</tbody>
</table>

Fig. 3.16 and Fig. 3.17 show the experimental waveforms of transformer current, drain-source voltage ($V_{ds2}$), and gate driver signal ($V_{gs2}$) of the low side MOSFET $S_2$ on the primary side of the DHB converter module with 1 kW and 300 W output, respectively. It can be seen from Fig. 3.16 that the ZVS operation can be achieved with 1 kW output power. With the lighter load such as 300 W, the ZVS condition, however, cannot be maintained and the MOSFET will work in hard switching mode as shown in Fig. 3.17. Consequently, the switching losses will be increased. Fig. 3.18 shows the measured efficiency with respect to the output power of the DHB converter module, where the high efficiency up to 97% can be achieved.

Fig. 3.14. Photo of the prototype.
Fig. 3.15.  Photo of the experimental setup.

Fig. 3.16.  Key waveforms of single DHB module at 1 kW.
Fig. 3.17. Key waveforms of single DHB module at 300 W.

Fig. 3.18. Measured efficiency with respect to output power.
3.4 Summary

High frequency bidirectional dc-dc converter module based on low voltage switching device has been proposed for 20 kVA single phase solid state transformer. Phase-shift ZVS technique along with low voltage switching device enables the high frequency operation of the dc-dc converter module while keeping low switching loss and conduction loss. In addition, planar transformer with interleaved windings has been designed and implemented to obtain low core loss, optimized high frequency copper loss, low profile, and solid insulation which results in enhanced reliability. As a result, the high frequency, high efficiency, and high power density can be achieved. A 50 kHz bidirectional dc-dc converter module have been built successfully and the measured efficiency is up to 97%.

Although ZVS operation can be achieved at heavy load, it will lose when the load power is much lower, and therefore leads to increased switching loss and reduced efficiency at light load. The drawback of the conventional DHB converter will be investigated and addressed in chapter 4.
CHAPTER FOUR

THE PHASE-SHIFT DHB CONVERTER WITH HIGH
EFFICIENCY OVER WIDE LOAD RANGE

4.1 Introduction

The efficiency of the isolated bidirectional dc-dc converter for SST application is of high importance because of the rising energy cost and its continuous operation. In order to achieve high efficiency under high frequency operation, a modular strategy using low-voltage rating power device is proposed and designed in chapter 3, where the dc-dc converter, as shown in Fig. 3.2, consists of multiple low-voltage modules connected in input-series and output-parallel mode so that low-voltage commercial silicon MOSFETs, which usually have low conduction losses and high switching speed, can be adopted. The phase-shift dual-bridge dc-dc converter appears to be the most promising candidate for individual dc-dc converter module since it can realize high-frequency transformer isolation, bidirectional power flow, and ZVS for all switching devices without auxiliary switch devices [52, 53, 54, 59, 60, and 61].

The Phase-shift dual-bridge dc-dc converters have been reported in [52, 54, 59, and 60]. However, they can operate in the ZVS mode only within a limited region restricted by the converter voltage ratio of input to output and the load condition, and suffer additional conduction losses due to the circulating energy at heavy load. Consequently, high efficiency can be achieved only within a limited load range. A few new control methods were proposed in [53 and 61] to handle the loss of ZVS due to input voltage variations. However, none of these previous attempts addressed both the loss of ZVS at light load and the additional conduction losses due to circulating energy at heavy load.

This chapter proposes a novel phase-shift DHB converter with an adaptive inductor. By utilizing an adaptive inductor as the main energy transfer element, the output power can be controlled by not only the phase shift but also the commutation inductance, which allows the circulating energy to be optimized for different load conditions to maintain ZVS under light load conditions and minimize additional conduction losses under heavy load conditions as well. As a
result, the efficiency at both light and heavy load can be significantly improved compared with
the conventional DHB converter, and therefore extended high-efficiency range can be achieved.
In addition, current stress of switch devices can be reduced. The theoretical analysis is presented
and validated by the experimental results on a 50 kHz, 1 kW dc-dc converter module.

4.2 Operation Principle of the DHB Converter with an Adaptive Inductor

Fig. 4.1 and Fig. 4.2 show the circuit diagram of the conventional DHB converter and
that of the proposed DHB converter with an adaptive inductor, respectively. The key ideal
steady-state waveforms of the two converters are similar and are shown in Fig. 4.3. The operation
of the conventional DHB converter is fully described in [52, 54, 59, and 60]. In the proposed
new DHB converter, an auxiliary adaptive inductor $L$ is used as the commutation inductor and
can be controlled to adapt to the output power by utilizing the output current $I_{out}$ as the bias
current $I_{BIAS}$, while the commutation inductance of the conventional DHB converter is fixed
regardless of the output power. This feature enables the new DHB converter to be operated with
optimized commutation inductance $L$ and phase shift $\varphi$ over the entire load range, and therefore
maintain ZVS operation at light load and reduce circulating energy and resulted additional
conduction losses at heavy load as well. As a result, the efficiency at both light and heavy load
conditions, compared with the conventional DHB converter, can be improved.

Fig. 4.1. The conventional DHB converter.
Fig. 4.2. The proposed DHB converter with an adaptive inductor.

Fig. 4.3. Key ideal waveforms of the phase-shift DHB converter.
To simplify the analysis, the converter voltage ratio of the input voltage $V_{in}$ to output voltage $V_{out}$ is assumed to be 1 (i.e., $V_{out} = N_s V_{in} / N_p$, where $N_p$ and $N_s$ are the number of primary and secondary turns in the transformer, respectively.). The output power can be given by:

$$P_{out} = \frac{1}{t_f - t_0} \int_{t_0}^{t_f} V_{out}^2 i_L = V_{in}^2 \varphi (\pi - |\varphi|) / 8\pi^2 f L$$

(4-1)

where $f$ is the switching frequency, $i_L$ is the instantaneous current of commutation inductor $L$, and $\varphi$ is the phase shift usually ranging from $-\pi/2$ to $\pi/2$. The output power of the proposed converter can be controlled by $\varphi$ and commutation inductance $L$ as well, while the output power of the conventional DHB converter can only be controlled by $\varphi$.

The instantaneous current of commutation inductor at $t_2$ can be given by:

$$i_{L(t_2)} = -V_{in} \varphi / 4\pi f L$$

(4-2)

By combing (4-1) and (4-2), the current stress of $S_2$ can be given by:

$$\left| i_{L(t_2)} \right| = 2\pi P_{out} / V_{in} (\pi - |\varphi|)$$

(4-3)

The delivered power during $(t_0-t_1)$ and $(t_2-t_6)$ can be obtained by:

$$\begin{align*}
P_{d01} &= \frac{1}{t_1 - t_0} \int_{t_0}^{t_1} V_{out}^2 i_L \\
P_{d26} &= \frac{1}{t_6 - t_2} \int_{t_2}^{t_6} V_{out}^2 i_L
\end{align*}$$

(4-4)

Combining (4-4) and the boundary conditions $i_{L(t_1)} = -i_{L(t_0)}$ and $i_{L(t_2)} = -i_{L(t_6)}$ yields:

$$P_{d01} = P_{d26} = 0$$

(4-5)

which means the average power during both $(t_0-t_1)$ and $(t_2-t_6)$ are equal to zero, but the energy stored in the commutation inductor will circulate in the circuit; the total circulating energy during one switching period $(t_0-t_7)$ can be given by:

$$E_C = 4\int_{t_2}^{t_6} V_{out}^2 i_L = V_{in}^2 \varphi^2 / 8\pi^2 f L = \varphi P_{out} / 2 f (\pi - |\varphi|)$$

(4-6)

$E_C$ is the sum of the energy stored in the commutation inductor during $(t_0-t_1)$ and $(t_2-t_6)$, and the energy stored in the commutation inductor during each transition period is half of $E_C$, and can be expressed as:
\[ E_L = \frac{E_C}{2} = \frac{\varphi P_{out}}{4f} \left( \pi - |\varphi| \right) \]  

ZVS operation of the switching devices is achieved by utilizing the circulating energy \( E_C \) to discharge the output capacitance \( C_o \) of the switching devices, and the circulating energy \( E_C \) should be no less than total capacitor energy to maintain ZVS operation, i.e.,

\[ E_C \geq E_{th} \]

Where \( E_{th} \) is the energy of ZVS threshold and insufficient energy \( E_C \) will result in the loss of ZVS of the switching devices. On the other hand, large circulating energy will cause large additional conduction losses. In order to achieve high efficiency over a wide load range, \( E_C \) is expected to be large enough to maintain ZVS to reduce switching losses at light load, but to be minimized to reduce additional conduction losses at heavy load.

Fig. 4.4. The circulating energy of the conventional DHB converter with an optimized fixed inductor.

In the conventional phase-shift DHB converter, commutation inductance \( L \) is fixed and the output power is controlled only by \( \varphi \). According to power equation (4-1), phase shift \( \varphi \), if ranging from \(-\pi/2\) to \(\pi/2\), is proportional to the output power \( P_{out} \). Combined with equation (4-6), it can be seen that the circulating energy \( E_C \) is proportional to \( P_{out} \), and therefore \( E_C \) at heavy load
is always much larger than that at light load due to larger $\phi$ and $P_{\text{out}}$. Consequently, $L$ can only be optimized for a limited load range, and so does the high efficiency. Fig. 4.4 shows the circulating energy with respect to the output power of the conventional DHB converter with an optimized fixed inductance. It can be seen that the circulating energy is too small to maintain ZVS at light load conditions while it is too large and will cause large additional conduction losses at heavy load conditions.

The proposed phase-shift DHB converter with an adaptive inductor in this work can overcome the above mentioned drawbacks by introducing another control variable, commutation inductance $L$. At light load, $L$ is controlled to be relatively large to obtain large $\phi$ and therefore sufficient $E_C$ to maintain ZVS of switching devices. On the other hand, $L$ at heavy load is controlled to be much smaller than that at light load to reduce $\phi$ and therefore reduce $E_C$. As a result, the circulating energy and the resulting additional conduction losses at heavy load can be minimized without losing ZVS at light load. Compared with conventional phase-shift DHB converter, the efficiency of the proposed dc-dc converter can be significantly improved under both light and heavy load conditions, and high competence can be achieved over wide load range.

4.3 Adaptive Inductor Design

The adaptive inductor can be implemented using a double RM core, as explained in [62-63]. The schematic and the photo are shown in Fig. 4.5. The main inductance $L$ is wound around the center leg whereas the symmetrical bias winding is wound around the two side arms. The side arms windings are serially connected in opposite polarity to cancel out the ac voltages induced by the center leg. The inductance $L$ can be controlled by the bias current $I_{\text{BIAS}}$. The output current $I_{\text{out}}$ is used as bias current in this paper so that $L$ and therefore circulating energy can be optimized according to the output power automatically to maintain ZVS operation at light load conditions and minimize the conduction loss caused by the circulating energy at heavy load conditions. As a result, the efficiency of the new DHB converter with an adaptive inductor, comparing to the conventional DHB converter with a fixed inductor, can be improved at both light load and heavy load conditions.
4.4 Improvement of the Phase-Shift DHB Converter with an Adaptive Inductor

The numerical analysis is presented on one dc-dc converter module in this section. The specifications are: $f = 50$ kHz, rated power $P_{\text{out\_rated}} = 1$ kW, $V_{\text{in}} = 500$ V, and $V_{\text{out}} = 400$ V. For the comparison purpose, an optimized fixed 90 µH inductor $L$ is chosen for the conventional phase-shift DHB converter, while the adaptive inductance as shown in Fig. 4.6 is adopted for the proposed DHB converter.
By substituting $L$ into (4-1), $\varphi$ as a function of output power can be obtained as shown in Fig. 4.7. Then according to (4-3) and (4-6), the current stress of $S_2$ and the circulating energy $E_C$ can be obtained as shown in Fig. 4.8 and Fig. 4.9, respectively. Compared with the conventional DHB converter, the proposed DHB converter with an adaptive inductor can not only achieve much higher circulating energy at light load to maintain ZVS to reduce switching losses, but also realize much lower circulating energy at heavy load to reduce the additional conduction losses. Therefore, high efficiency can be achieved over wide load range.
Fig. 4.8. The Current stress of the DHB converters.

Fig. 4.9. The circulating energy of the DHB converters.

4.5 Experimental Results

| TABLE 4-1. KEY SPECIFICATIONS AND CIRCUIT PARAMETERS OF TWO DHB CONVERTERS |
|-----------------|-----------------|-----------------|
| $V_{in}$ (V)    | $V_{out}$ (V)   | Transformer Core |
| $V_{in}$ (V)    | $V_{out}$ (V)   | Commutation      |
| $V_{in}$ (V)    | $V_{out}$ (V)   | Primary          |
| $V_{in}$ (V)    | $V_{out}$ (V)   | Secondary        |
| 500             | 400             | PC40-PQ107/87/70 |
|                 |                 | N87-RM10         |
|                 |                 | IXFT24N80P       |
|                 |                 | IXFT36N60P       |

A 50 kHz, 1 kW new DHB converter module with an adaptive inductor as well as a conventional DHB converter module with a fixed inductor, as shown in Fig. 4.10, was built in
the lab and tested to verify the high frequency and high efficiency operation. The specifications and circuit parameters of the individual converter module are shown in Table 4-1.
The conventional DHB converter with fixed inductor was also tested and compared with the proposed new DHB converter with an adaptive inductor to validate the extended high-efficiency range of the latter. The adaptive inductance shown in Fig. 4.6 is adopted for the new DHB converter while an optimized fixed 90 μH is chosen for the conventional DHB converter.

Fig. 4.11 shows the key switching waveforms of two DHB converters at light load condition. ZVS of $S_2$ of the DHB converter with an adaptive inductor can be maintained in (a), while $S_2$ of the conventional DHB converter loses ZVS in (b). Fig. 4.12 shows the switching waveforms of the two DHB converters at heavy load condition. The DHB converter with an adaptive inductor can achieve lower current stress and smaller phase shift which means smaller circulating energy and resulted additional conduction losses. Fig. 4.13 shows comparison of the measured efficiency of two DHB converters. Both DHB converters can realize high efficiency during mid-range load conditions. However, the DHB converter with an adaptive inductor can achieve higher efficiency at both light and heavy load conditions.
Fig. 4.11. Key waveforms of two DHB converters at 300 W: (a) new DHB converter with an adaptive inductor, (b) conventional DHB converter with a fixed inductor.
Fig. 4.12. Key waveforms of two DHB converters at 1 kW: (a) new DHB converter with an adaptive inductor, (b) conventional DHB converter with a fixed inductor.
A novel phase-shift dual-half-bridge converter with an adaptive inductor has been proposed. By utilizing an adaptive inductor as the commutation inductor, the proposed converter can control not only the phase shift but also the commutation inductance so that circulating energy can be optimized for both light and heavy load, and therefore extended ZVS operation range, reduced current stress, and additional conduction losses due to circulating energy at heavy load can be achieved. Hence, the efficiency under both light load conditions and heavy load conditions can be significantly improved, and high efficiency can be achieved within an extended load range.

Fig. 4.13. Measured efficiency of two DHB converters.

4.6 Summary

A novel phase-shift dual-half-bridge converter with an adaptive inductor has been proposed. By utilizing an adaptive inductor as the commutation inductor, the proposed converter can control not only the phase shift but also the commutation inductance so that circulating energy can be optimized for both light and heavy load, and therefore extended ZVS operation range, reduced current stress, and additional conduction losses due to circulating energy at heavy load can be achieved. Hence, the efficiency under both light load conditions and heavy load conditions can be significantly improved, and high efficiency can be achieved within an extended load range.
CHAPTER FIVE

THE IDENTICALLY DISTRIBUTED CONTROL OF ISOP MODULAR DHB CONVERTER

5.1 Introduction

Due to the constant growth of the distributed renewable energy generation and utilization, future electric energy delivery and management systems will demand power electronics to take an even larger role in the conversion and control of electric energy generated by highly distributed renewable energy source. Medium-voltage isolated bidirectional dc-dc converter will become one of the key interfaces for grid components with moderate power ratings.

ISOP modular dc-dc converter is one of the most promising candidates to address the demand of medium-voltage and high-power capability of future electric energy delivery and management systems, since the modular approach, that utilizes the well-known, mature, and more economic power semiconductor devices to handle MV power conversion can achieve several advantages include: significant improvement in reliability by introducing desired level of redundancy; standardization of components leading to reduction in manufacturing cost and time; power systems can be easily reconfigured to support varying input-output specifications, and possible higher efficiency and power density of the overall system. Input-voltage sharing and output-current sharing are critical to assure the advantages of the ISOP modular configuration. Various control schemes were developed to achieve the goal for ISOP buck-derived converters. However, none of the previous work has investigated the ISOP phase-shift dual-bridge converters such as ISOP DHB and ISOP DAB converter which have different characteristics from the buck-derived converters. In addition, no paper has clarified the effect of the constituent module topology on the control scheme stability while most paper discussed the stability of the control scheme only in terms of the converter configuration. Moreover, the reported control schemes developed for buck-derived converters may be invalid for the ISOP phase-shift dual-bridge converters, and even result in unstable operation.
This chapter proposes an identically distributed control scheme for ISOP DHB converter. The proposed control scheme, using only one distributed voltage loop to realize both input-voltage and output-current sharing, provides plug-and-play capability, possible high-level fault tolerance, and easy implementation. Another unique advantage of the proposed ISOP DHB converter is the power rating can be easily extended further by directly connecting multiple ISOP DHB converters in input-parallel-out-parallel (IPOP) while no additional control is needed. The proposed control scheme is elaborated using the large-signal average model. Further, the stability of the control schemes is analyzed in terms of the constituent modules’ topology as well as the configuration, and then an important fact that the stability of control scheme depends on not only the configuration but also the constituent module topology is first revealed in this work. Finally, the simulation and experiment are conducted to evaluate both static and dynamic performances of the proposed dc-dc converter, and the simulation and experimental results are presented to verify the proposed control scheme.

5.2 The ISOP DHB Converter with Identically Distributed Control

Fig. 5.1. Block diagram of the proposed ISOP converter with identically distributed control.

Fig. 5.1 shows the block diagram of an ISOP DHB converter with identically distributed control. Each constituent module is a phase-shift DHB converter with the schematic shown in...
Fig. 5.2. In the proposed identically distributed control scheme, all control circuits are distributed evenly in the constituent modules, and only a common dc bus voltage, $V_{bus}$, is shared as a dynamic reference. Hence, the individual modules are identical and self-contained, and therefore plug-and-play capability and possible high-level fault tolerance can be achieved. The detailed operation principle and stability analysis of the control scheme will be presented in section 5.4.

![Fig. 5.2. The schematic of the constituent DHB module.]

### 5.3 Large Signal Average Model

Large signal average models of DC-DC converters offer several advantages over the switching models such as the analysis of the relationship between the input and output without being obscured by the switching-frequency ripple, faster simulation of transient response to large-signal changes, and allowing general-purpose simulators to linearize the converters for the feedback controller design [64]. In this paper, the average model is developed to elaborate the control scheme and evaluate both static and dynamic performance of the proposed dc-dc converter.

![Fig. 5.3. The primary-referred equivalent circuit of the individual DHB converter.]

54
Fig. 5.3 and Fig. 5.4 show the primary-referred equivalent circuit of the single DHB converter module and its ideal operation waveforms, respectively. According to Fig. 5.4, the instantaneous inductor current during one switching cycle can be expressed by:
The boundary conditions can be given by:

\[
\begin{align*}
    i_L(0) &= -i_L(\pi) \\
    i_L(\varphi) &= -i_L(\pi + \varphi)
\end{align*}
\]  

(5-2)

By substituting (5-1) into (5-2), the instantaneous inductor current at boundary conditions can be obtained and given by:

\[
\begin{align*}
    i_L(0) &= -\frac{(v_i - v_s)(\pi - \varphi) - (v_i + v_s)\varphi}{2\omega L} \\
    i_L(\varphi) &= -\frac{(v_i - v_s)(\pi - \varphi) + (v_i + v_s)\varphi}{2\omega L} \\
    i_L(\pi) &= \frac{(v_i - v_s)(\pi - \varphi) + (v_i + v_s)\varphi}{2\omega L} \\
    i_L(\pi + \varphi) &= \frac{(v_i - v_s)(\pi - \varphi) - (v_i + v_s)\varphi}{2\omega L}
\end{align*}
\]  

(5-3)

Then state equations of split capacitors on secondary can be expressed as:

\[
\begin{align*}
    C_s \frac{dv_3}{dt} &= [i_L(\alpha) - C_o \frac{dv_{out}}{dt}]G_3 - C_o \frac{dv_{out}}{dt} (1 - G_3) \\
    C_s \frac{dv_4}{dt} &= [-i_L(\alpha) - C_o \frac{dv_{out}}{dt}]G_3 - C_o \frac{dv_{out}}{dt} (1 - G_3)
\end{align*}
\]  

(5-4)

To simply the analysis, the parameters are assumed as:

\[
C_s = C_4 = C_3
\]  

(5-5)

Since the output voltage equals to the sum of voltage of capacitor \(C_3\) and \(C_4\), i.e.

\[
v_{out} = v_3 + v_4
\]  

(5-6)

The state equation of output voltage can be expressed as:
\[
\frac{C_s + C_o}{2} \frac{dv_{\text{out}}}{dt} = \frac{i_s(\alpha)G_i - i_i(\alpha)(1 - G_i)}{4} \tag{5-7}
\]

Where
\[
\begin{aligned}
i_s(\alpha)G_i &= \frac{[i_s(\varphi) + i_s(\pi)](\pi - \varphi) + [i_i(\varphi + \pi) + i_i(\pi)]\varphi}{2\pi} \\
i_i(\alpha)(1 - G_i) &= -i_i(\alpha)G_i
\end{aligned} \tag{5-8}
\]

Substituting (3) into (8) yields:
\[
\begin{aligned}
i_s(\alpha)G_i &= \frac{\varphi(\pi - \varphi)(2v_i + v_i - v_s)}{2\pi o L} \\
i_i(\alpha)(1 - G_i) &= -\frac{\varphi(\pi - \varphi)(2v_i + v_i - v_s)}{2\pi o L}
\end{aligned} \tag{5-9}
\]

Generally, DHB operates with duty cycle equals to 0.5, which suggests:
\[
\begin{aligned}
v_1 = v_4 &= \frac{1}{2}v_{\text{out}} \\
v_i = v_2 &= \frac{1}{2}v_{\text{in}}
\end{aligned} \tag{5-10}
\]

Combining (5-7), (5-9), and (5-10), the state equation of the output voltage can be expressed by
\[
C_{\text{out}} \frac{dv_{\text{out}}}{dt} = \frac{\varphi(\pi - \varphi)}{4\pi o L} v_{\text{in}} \tag{5-11}
\]

where \(C_{\text{out}}\) is the equivalent output capacitance with following expression
\[
C_{\text{out}} = \frac{C_s + C_o}{2} \tag{5-12}
\]

Due to symmetrical structure, the state equation of input voltage can be given by
\[
C_{\text{in}} \frac{dv_{\text{in}}}{dt} = \frac{\varphi(\pi - \varphi)}{4\pi o L} v_{\text{out}} \tag{5-13}
\]

where \(C_{\text{in}}\) is the equivalent input capacitance of the DHB converter. Based on the state equation, the large-signal average model of individual DHB converter and the proposed ISOP DHB
converter, as shown in Fig. 5.5 and Fig. 5.6, respectively, can be obtained, and will be used to analyze the control scheme.

![Diagram of DHB converter](image)

Fig. 5.5. Large-signal average model of the DHB converter.

For DHB module 1, the following equations can be given:

\[
i_t = g_{m1} V_{out} \tag{5-14}
\]

\[
i_{out1} = g_{m1} V_{in1} \tag{5-15}
\]

where \( g_{m1} \) is the transfer conductance of DHB module 1, and can be expressed as:

\[
g_{m1} = N \phi_1 (\pi - \phi_1) / 8 \pi^2 f \mu_t \tag{5-16}
\]

These analyses are also valid for other DHB modules constituting the ISOP DHB converter.
5.4 The Control Scheme for ISOP DHB Converter

As there is no dc path through the series inputs, the currents flowing in the input capacitors need be controlled to be equal to balance the input voltages among multiple modules.
\[ i_1 = i_2 = \ldots = i_m \] (5-17)

Since the input current \( i_m \) for all constituent modules is same due to the series connection on input side, the converter currents must fulfill:

\[ i_1 = i_2 = \ldots = i_n \] (5-18)

By substituting (5-14) into (5-18) and taking into account that the output voltage \( V_{out} \) of all DHB modules is same due to the parallel connection, yields

\[ g_{m1} = g_{m2} = \ldots = g_{mn} \] (5-19)

It indicates that the transfer conductance of the DHB modules must be equal to each other to achieve the input voltage sharing. Assuming the series inputs are controlled to share voltage evenly, then the DHB modules will have same input voltages and transfer conductance. According to (5-15), equal input voltages and transfer conductance will result in equal output currents

\[ i_{out1} = i_{out2} = \ldots = i_{outm} \] (5-20)

It means that the output current sharing can be obtained automatically once the input voltage sharing is achieved.

Based on the above analysis, a simple identically distributed control scheme is proposed and implemented in this work. As shown in Fig. 5.7, each individual DHB module has an input voltage loop to realize both input voltage sharing and output current sharing, while a local output voltage loop is used to regulate the output voltage. These distributed input voltage loops share a common dc bus voltage \( V_{bus} \) as the reference, which can be generated by averaging the series input voltages, and can be expressed as:

\[ V_{bus} = \left( \frac{\sum_{j=1}^{n} V_{aj}}{n} \right) \] (5-21)

Since the implementation circuits of \( V_{bus} \) are distributed evenly in constituent modules, the modules are identical. If there is only one module, i.e., \( n=1 \), \( V_{bus} \) is always equal to the input voltage according to (5-21), and the output voltage loop will not be modified by input voltage loop. Thus, the individual module can operate independently with full output voltage regulation function, which suggests that these individual modules of the ISOP DHB converter are self-contained. Thanks to the identically distributed control scheme, the individual module features
plug-and-play capability and fault-tolerance through proper redundancy design. Although the novel control scheme is developed based the ISOP DHB converter, it can also valid for other ISOP phase-shift dual-bridge converters.

Fig. 5.7. The proposed ISOP DHB converter with identically distributed control.
5.5 The Expanded Power Capability of the ISOP DHB Converter

Taking DHB module 1 as an example, the output power can be expressed as

\[ P_{out1} = NV_{in1}V_{out1}\varphi(\pi - |\varphi_1|)/8\pi^2fL_1 \]  

(5-22)

where \( V_{in1} \) is input voltage, \( V_{out1} \) is output voltage, \( f \) is the switching frequency, \( N \) is transformer turn ratio, \( L_1 \) is the commutation inductance, and \( \varphi_1 \) is phase shift angle. Then, the output current can be given by

\[ i_{out1} = P_{out1}/V_{out1} = NV_{in1}\varphi_1(\pi - |\varphi_1|)/8\pi^2fL_1 \]  

(5-23)

The output current expression suggests that the output current \( i_{out1} \) is independent from the output voltage \( V_{out1} \). This unique current-source like characteristics distinguishes DHB from the buck-derived dc-dc converters, and enables inherent stable operation when they are connected in input-parallel-out-parallel (IPOP). Fig. 5.8 shows the IPOP conversion system consisting of three DHB converters as shown in Fig. 5.2, and no any additional balance control is added. In order to verify the stable operation without additional control, the commutation inductances of these DHB converters are set as different values on purpose, namely \( L_1=90\mu H \), \( L_2=70\mu H \), and \( L_3=50\mu H \). The simulation results are shown in Fig. 5.9, where it can be seen that the stable operation can be achieved although the output currents of each DHB converter are different from each other due to different commutation inductance among these DHB converters.

![Diagram of IPOP conversion system consisting of three DHB converters.](image)

Fig. 5.8. IPOP conversion system consisting of three DHB converters.

62
Fig. 5.9. The output currents of three DHB converters connected in IPOP without additional control.

Fig. 5.10. IPOP conversion system employing ISOP DHB converter as constituent subsystem.
Benefiting from the current-source-like characteristic and the proposed distributed control, multiple ISOP DHB converters can also be directly connected in IPOP to extend the power rating while no additional control is needed. This is an important advantage of the proposed ISOP DHB converter for the medium-voltage high-power application. Fig. 5.10 shows an IPOP conversion system, of which the constituent subsystems are ISOP DHB converters shown in Fig. 5.1. The operation will be verified in section 5.7.

5.6 Discussion of Control Stability

Much of the previous work discussed the stability of the control schemes only in terms of the converter configuration. No paper, however, has been published to study the effect of the constituent module topology on the control scheme stability. An important fact is first revealed in this paper that besides the converter configuration, the topology of the constituent modules should be taken into account as well when it comes to the stability of control schemes. Otherwise, it may lead to incorrect conclusion. For instance, common-duty-ratio control proposed for ISOP forward converter in [39] and sensorless current mode control proposed for ISOP push-pull converter in [48] have been validated by simulation and experimental results. These two control schemes can obtain stable operation for ISOP buck-derived converters due to the so-called self-correcting mechanism of buck-derived converters connected in ISOP configuration. Since the two control methods are based on a similar principle, only common-duty-ratio control will be taken as an example for the detailed analysis and simulation verification in this paper.

As discussed in [39], employing forward converter as the constituent modules, the common-duty-ratio control can realize stable operation for ISOP configuration, but results in unstable operation for IPOP configuration. The contrary conclusion, however, will be drawn as follows when employing DHB converter as the constituent modules. Common-duty-ratio control will lead to unstable operation for ISOP configuration, but stable operation for IPOP configuration due to the unique current-source like characteristic distinguishing phase-shift dual-bridge converter from buck-derived converter.

Fig. 5.11 shows the large-signal average model of ISOP DHB converter adopting common duty ratio control. The commutation inductances of two DHB converter modules are set as different values on purpose, namely 90 µH and 88 µH, to test the control scheme. According
to (5-14) and (5-16), the current $i_2$ of DHB module 2 is larger than the current $i_1$ of DHB module 1 due to the smaller commutation inductance $L_2$ of DHB module 2, which suggests that the charging current into $C_{in2}$, $i_c2$, will be smaller than that into $C_{in1}$, $i_c1$. Consequently, input voltage of DHB module 1 will be driven to a higher level than the input voltage of DHB module 2. Unlike the buck-derived converter, the increased input voltage of DHB converter $V_{in1}$, however, will not increase $i_1$ since it is independent of $V_{in1}$. Hence, $i_c1$ is always higher than $i_c2$, which keeps driving $V_{in1}$ to higher and higher level, and eventually results in a runaway condition. Fig. 5.12 shows the simulation waveforms of ISOP DHB converter adopting common-duty-ratio control.

**Fig. 5.11.** Large-signal average model of ISOP DHB converter adopting common-duty-ratio control.
Fig. 5.12. Simulation waveforms of the ISOP DHB converter adopting common-duty-ratio control: (a) output voltage and input voltages, and (b) output voltage and output currents.
Fig. 5.13 and Fig. 5.14 show the large-signal average model of IPOP DHB converter adopting common-duty-ratio control and simulation results with same parameters as the ISOP DHB converter discussed above, respectively. Although there is slight difference between currents of the two constituent DHB modules, stable operation can be achieved.

For the same converter configuration, the same control scheme may lead to completely contrary conclusion when employing different topology as constituent module of the modular converter. Hence, the topology of individual constituent module should be taken into account as well as the converter configuration when it comes to the stability of the modular converter.

Fig. 5.13. Large signal average model of IPOP DHB converter with common-duty-control.
5.7 Simulation and Experimental Results

An ISOP DHB converter consisting of three DHB modules with identically distributed control has been designed and tested by simulation to verify the stable and accurate input voltage and current sharing as well as the high frequency operation. The power rating of the ISOP DHB converter is 3 kW, and its input voltage and output voltage are 1500 V and 400 V, respectively. Accordingly, each individual DHB module is rated for 1 kW and 500 V for output power and input voltage, respectively.

In the simulation, big differences among the commutation inductances of three constituent DHB modules are set on purpose to verify the proposed control scheme, where \( L_1=90\mu\text{H} \), \( L_2=70\mu\text{H} \), and \( L_3=50\mu\text{H} \). Fig. 5.15 and Fig. 5.16 show the input voltage and output
current sharing during start up and the 75%-100%-75% load transient with constant-resistance load.

Fig. 5.15. Input voltage and output current sharing of ISOP DHB converter during start up.
Further, as shown in Fig. 5.17, three of these ISOP DHB converters are used as subsystems to build an IPOP conversion system without any additional current-sharing control. The rated output power of the IPOP system is three times large as the ISOP converter while keep the same input voltage. The commutation inductances are also set as different values on purpose and listed in Table 5-1.
The IPOP conversion system consisting of three ISOP DHB converters is illustrated in Fig. 5.17.

**TABLE 5-1. COMMUTATION INDUCTANCE OF THE IPOP SYSTEM**

<table>
<thead>
<tr>
<th>Inductance (µH)</th>
<th>Module 1</th>
<th>Module 2</th>
<th>Module 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>ISOP DHB converter A</td>
<td>90</td>
<td>70</td>
<td>50</td>
</tr>
<tr>
<td>ISOP DHB converter B</td>
<td>88</td>
<td>68</td>
<td>48</td>
</tr>
<tr>
<td>ISOP DHB converter C</td>
<td>86</td>
<td>66</td>
<td>46</td>
</tr>
</tbody>
</table>

Fig. 5.18 shows the total output current of each ISOP subsystem; the stable operation can be achieved without any additional current-sharing control, although there is difference among these currents due to unmatched parameters. Fig. 5.19 and Fig. 5.20 show current sharing and voltage sharing inside each ISOP subsystem, respectively. It can see that both input voltage and output current are shared equally inside every ISOP subsystem.
Fig. 5.18. Total output currents of three ISOP converters of the IPOP conversion system.

Fig. 5.19. Current sharing inside each ISOP DHB converter of IPOP conversion system.
The prototype of an ISOP DHB converter consisting of three DHB modules with identically distributed control has been designed, built in the lab, and tested to verify the stable and accurate input voltage and current sharing as well as the high frequency and high efficiency operation. The power rating of the ISOP DHB converter is 3 kW, and its input voltage and output voltage are 1500 V and 400 V, respectively. Accordingly, each individual DHB module is rated for 1 kW and 500 V for output power and input voltage, respectively. The experiments of the ISOP DHB converter was conducted with down-scaled input voltage of 750 V and output voltage of 200 V. Fig. 5.21 shows the waveforms during start up. Fig. 5.22 and Fig. 5.23 show transformer currents and input voltages of three constituent DHB modules at steady state, respectively. Fig 5.24 shows the prototype built in lab.
Fig. 5.21. Input-voltage and output-current sharing during start up.

Fig. 5.22. Transformer currents under steady state.
Fig. 5.23. Input voltages during steady state.

Fig. 5.24. Photo of the ISOP DHB converter consisting of three DHB modules.
5.8 Summary

An identically distributed control has been proposed for ISOP modular DHB converter. Benefiting from the proposed identically distributed control, the constituent modules of the ISOP DHB converter are identical and self-contained, thus the plug-and-play capability and high-level fault-tolerance can be obtained by proper redundancy design. The current-source like characteristics, combined identically distributed control, provides the ISOP DHB converter the capability to extend the power rating to even higher level by directly connecting ISOP DHB converter in IPOP configuration without any additional control. In addition, an important fact that the stability of the control schemes depends not only the converter configuration but also the topology of the constituent modules is first revealed and explained in this paper. The control scheme is also valid for other phase-shift dual-bridge converters connecting in ISOP configuration. A 50 kHz ISOP DHB converter consisting of three DHB modules will be built with the proposed identically distributed control scheme to verify the theoretical analysis.

An ISOP DHB converter with identically distributed control has been proposed for medium-voltage SST. The ISOP configuration enables the using of low-voltage switching devices featuring low conduction losses for medium voltage application. The DHB module utilizes phase-shift ZVS technique to achieve high frequency operation with low switching losses. Therefore, high frequency, high efficiency, and high power density can be achieved. Benefiting from the proposed identically distributed control, the constituent modules of the ISOP DHB converter are identical and self-contained, thus the plug-and-play capability and high-level fault-tolerance can be obtained by proper redundancy design. The current-source like characteristics, combined identically distributed control, provides the ISOP DHB converter the capability to extend the power rating to even higher level by directly connecting ISOP DHB converter in IPOP configuration without any additional control. In addition, an important fact that the stability of the control schemes depends not only the converter configuration but also the topology of the constituent modules is first revealed and explained in this paper. The control scheme is also valid for other phase-shift dual-bridge converters connecting in ISOP configuration. A 50 kHz ISOP DHB converter consisting of three DHB modules has been built successfully with the proposed identically distributed control scheme. The constituent modules
can share the input voltage and the output current evenly under both transient and steady state operation modes and 97% can be achieved at rated output power.
CHAPTER SIX

CONCLUSIONS AND FUTURE WORK

6.1 Conclusions

Due to the constant growth of the distributed renewable energy generation and utilization, future electric energy delivery and management systems will demand power electronics to take an even larger role in the conversion and control of electric energy by highly distributed renewable energy source. Medium-voltage isolated bidirectional dc-dc converter will become one of the key interfaces for grid components with moderate power ratings.

In this work, input-series output-parallel (ISOP) modular DHB converter based on traditional low-voltage power semiconductor devices is proposed to address the demand of medium-voltage and high-power capability for future electric energy delivery and management systems. The proposed ISOP modular DHB converter employs zero-voltage-switching technique combined with low-voltage MOSFETs to achieve low switching and conduction losses under high frequency operation, and therefore high efficiency and high power density, and bidirectional power flow as well. Besides topology investigation, planar transformer with interleaved windings has been designed and implemented to obtain low core loss, optimized high frequency copper loss, low profile, and solid insulation which results in enhanced reliability. A 50 kHz bidirectional dc-dc converter module have been built successfully and the measured efficiency is up to 97%.

Another contribution of this work is the proposed novel DHB converter with an adaptive commutation inductor, which enables an extended load range of high efficiency. Other than the fixed commutation inductance of the conventional DHB converter, the newly proposed DHB converter utilizes an adaptive inductor as the main energy transfer element, and therefore the output power can be controlled by not only the phase shift but also the commutation inductance, which allows the circulating energy to be optimized for different load conditions to maintain ZVS under light load conditions and minimize additional conduction losses under heavy load conditions as well. As a result, the efficiency at both light and heavy load can be significantly
improved compared with the conventional DHB converter, and therefore extended high-efficiency range can be achieved. In addition, current stress of switch devices can be reduced.

In addition, an identically distributed control scheme is proposed in this work to assure the input-voltage sharing and output-current sharing and resulted advantages of the ISOP modular configuration. The proposed control scheme, using only one distributed voltage loop to realize both input-voltage and output-current sharing, provides plug-and-play capability, possible high-level fault tolerance, and easy implementation. Another unique advantage of the proposed ISOP DHB converter is the power rating can be easily extended further by directly connecting multiple ISOP DHB converters in input-parallel-out-parallel (IPOP) while no additional control is needed.

Moreover, the stability of the control schemes is analyzed in terms of the constituent modules’ topology as well as the configuration, and then an important fact that the stability of control scheme depends on not only the configuration but also the constituent module topology is first revealed in this work.

6.2 Scope of Future Work

The ISOP DHB converter proposed in this work utilizes low-voltage rated MOSFETs and ZVS technique to achieve high-frequency high-efficiency operation for medium voltage application. The possible improvements on performance and reliability can be expected by further investigation.

First of all, the conventional silicon MOSFETs are used because of their low conduction loss and high switching speed. The breakdown voltages of the MOSFETs, however, are usually low and therefore large amount of constituent modules and MOSFETs are required to address the medium voltage. Consequently, circuit complexity and failure rate are increased. One possible solution is to replace silicon MOSFETs with post-silicon devices such as silicon carbide (SiC) and gallium nitride (GaN). A lot of efforts have been put into the development of the next-generation, ultra low-loss, high-speed, high breakdown voltage post-silicon switching devices, and their effectiveness in improving the overall performance and reliability of MV dc-dc converter should be investigated.

In addition, the economic ferrite core is adopted for the transformer and inductor to achieve low core loss. However, the saturation flux density of ferrite material is relatively low,
usually around 0.4 Tesla at room temperature, and may result in large core size when higher voltage rating is demanded. The core material with higher saturation flux density, such as nanocrystalline core and amorphous can be investigated for the transformer design to reduce the transformer size for higher voltage application.

Moreover, the proposed ISOP modular configuration features possible high-level fault tolerance. However, the appropriate fault tolerant circuit is needed to assure the advantage. Fault tolerance means that in the event of failure of one or more modules the system should still be able to supply the load without interruption. This is normally done by providing some redundancy in the system, i.e., by incorporating more modules than is required for supplying the load. The fault tolerant circuit, which is the key to realize the improved reliability, should be investigated.
REFERENCES


[21] IEEE P1709™/ Draft D06.1, “Recommended Practice for 1 to 35 kV Medium Voltage DC Power Systems on Ships”.


BIOGRAPHICAL SKETCH

Haifeng Fan received his B.S. degree from Huazhong University of Science & Technology, Wuhan, China in 2001 and M.S. degree from Zhejiang University, Hangzhou, China in 2004, both in Electrical Engineering. From 2004 to 2005, he worked as an Electrical Engineer at Celestica Shanghai R&D Center in China. From 2005 to 2008, he worked as a Supervisor of DC-DC Converter Design with Delta Electronics Shanghai Design Center in China.

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