Charging and discharging methods for modularized battery energy storage systems (MBESS)

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CHARGING AND DISCHARGING METHODS FOR MODULARIZED BATTERY ENERGY STORAGE SYSTEMS (MBESS)

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Abstract

Electric vehicles (EV) promise to be a panacea to address the pressing problem of global warming from greenhouse gas emissions. EV possesses the potential to radically transform the personal transportation sector in the near future. The research studies related to EV are multi-disciplinary, comprising of material science, electro-chemistry, power electronics, mechanical, embedded systems etc. This research primarily focuses on the study of operation and control of power converters that are suitable for EV application. Modularization and swapping of battery system are taken as major approaches in this research study to improve the reliability of EVs from the energy source standpoint. Conventionally, the battery pack and hence the power converters employed for charging and driving the motor in an EV are monolithically constructed. Consequently, the operation of EV completely depends on a single energy source. Alternatively, modularizing the energy source could greatly increase the operational reliability of an EV and thus the necessity of modularizing the power converters arises. The charging time required by the EVs is significantly longer when compared with the time required to refuel a gasoline vehicle. This is considered as the debilitating disadvantage for the electric vehicle commercialization. Hence, instead of charging the battery in the EV, swapping the depleted battery with another charged battery could ameliorate the aforementioned disadvantage. The novel inductive power transfer (IPT) concept is considered for implementation to improve the performance of this battery swapping process. Two systems incorporating IPT are proposed and analysed as a part of this research study.

The Cascaded H-bridge (CHB) multi-level converter is generally considered an apt choice when multiple DC sources are available as in the case of modularized battery energy storage system (MBESS). This work investigates a battery module/micro-pack level SOC balancing control in a CHB based MBESS using Multi-dimensional pulse width modulation (MD-PWM). MD-PWM is a generalized modulation strategy for CHB converters, and a review of which is conducted. Modular battery pack system or micro-pack concept for CHB is introduced to increase the availability of the EV in the event of fault. A control strategy is proposed for the SOC balancing between the battery micro-packs in the CHB. This proposed control strategy aims to deliver a semblance of uniform operation of the whole battery system irrespective of the current SOC of each micro-pack. The realization of fault tolerant operation using MD-PWM is also proposed. The

advantage of the proposed methods is that it can be conveniently integrated with the MD-PWM algorithm, eliminating any external circuit. The results obtained from the laboratory setup of a five-level three-phase CHB with six battery modules of 15 V 6.2 Ah is presented for RL load. The results obtained from the laboratory setup of a five-level three-phase CHB driving an induction motor with six battery modules of 52.8 V 60 Ah is also presented. The results verify that the fault tolerant and SOC balancing operation can be achieved with CHB using MD-PWM. The results demonstrates the need of SOC balancing, and efficacy of the proposed methods in achieving fault tolerant as well as SOC balanced operation to prolong the system operation.

A modularized battery system with Double Star Chopper Cell (DSCC) based modular multilevel converter is proposed for a battery operated electric vehicle (BEV). A conceptual design of lithium ion based battery micro-packs for DSCC is described. MD-PWM with integrated inter-module SOC balancing and fault tolerant control for DSCC is proposed and explained. The methods of operation of the DSCC as an inverter and as a synchronous rectifier using MD-PWM with integrated inter-module SOC balancing and fault tolerant control is discussed. The performances of the DSCC for SOC balancing and fault tolerant operations are verified through simulations of a 28 kWh BESS with DSCC and the results are discussed.

Another objective of reducing the down-time of an EV is achieved by implementing battery swapping regime. To realize a replaceable and modularized energy storage system with wireless interface, two topologies are proposed for a battery operated EV. The proposed systems are analysed from the power converters operation and the Inductive Power Transfer (IPT) perspective. The proposed systems would have a casing (hollow compartment fitted to the vehicle chassis) that can accommodate certain number of modules (box like structure). The casing side is deployed with IPT secondary side which is connected to the DC link of the inverter driving the motor. The module (box like structure) side is deployed with the battery micro-pack and IPT primary side. First topology is comprised of multiple and independent module-casing combination, whereas the second topology is comprised of multiple modules inside a single casing.

The operation of the first topology is analysed with an equivalent circuit and an averaged state-space model. A feedback linearization technique to establish the DC link voltage regulation based on the phase shift ratio between the primary side and the

secondary side voltages is implemented. The power flow is controlled using the module side converter to maintain the DC link voltage at the three-phase inverter driving the three-phase load. The second topology is analysed with an equivalent electric circuit and the expressions for power flow are derived using steady state analysis of the equivalent circuit. The operation and control of the proposed systems are described, presenting steady state analysis and simulation results of an 80 kW system. Laboratory prototypes of 1.5 kW systems for each topology with IPT interfaces are built, and the results from both simulations and the laboratory prototypes are presented and discussed. The results show that the power flow from the battery modules can be controlled independently. Thus the operational reliability and the availability of an EV could be potentially enhanced by employing the proposed system which offers modularity and would make the battery swapping process more safe, convenient and reliable.

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List of Abbreviations

BESS	Battery Energy Storage System
BEV	Battery Electric Vehicle
CHB	Cascaded H-bridge
DSCC	Double Star Chopper Cell
EV	Electric Vehicle
ESS	Energy Storage System
MD-PWM	Multi-dimensional Pulse Width Modulation
PWM	Pulse Width Modulation
RMS	Root Mean Square
SOC	State of Charge
THD	Total Harmonic Distortion

Chapter 1 Introduction

1.1 Background and Motivation

Electrification of personal transportation has been under investigation by various organizations and vehicle manufacturers. The replacement of gasoline fuelled vehicles with electric vehicles (EVs) in mega cities has been considered by various governments with the motive of reducing the greenhouse gases emission. Various levels of electrification of vehicle power including hybrid electric vehicles, electric vehicles fully operated with batteries, fuel cells, photo-voltaic, super-capacitor or ultra-capacitor based energy sources etc., have been thoroughly investigated in the recent past. The results of which have been translated by some manufacturers including Toyota, Mitsubishi etc. The electrification of vehicles is not only adopted for personal transportation, but also have been investigated for mass rapid transit systems like trains, planes and ships. A brief list and specifications available online of some commercially available electric cars for personal transportation are compiled in Table 1.1. It can be observed that almost all cars have similar architecture having a single battery pack driving a three-phase AC motor through a three-phase two-level inverter.

The EVs in production have been well received in the market, but need further improvement to make them more attractive option for potential consumers. The battery operated electric vehicles (BEVs) are under further research and development, primarily to increase the performance of the EVs. This is important given that the EV's performance lags on certain parameters when compared with commonly used gasoline fuelled vehicles. The robustness, safety concern, range anxiety, waiting time to charge the EV battery pack are such parameters which are commonly considered weakening disadvantages of EV. Apart from developing a more efficient EV, the successful commercialization of EVs requires significant gestation time and effort in altering the environments and infrastructures of gasoline vehicles to accommodate the EVs. Hence the research in EV is the multi-disciplinary subsuming the combined efforts of various engineering and science disciplines: from the electrochemists researching into novel materials and new more efficient batteries for EV to the mechanical engineers researching into the design of vehicles that accommodates the batteries and electric motor.

Table 1.1 List of commercial electric cars [Source: Online]

Car	Year	Battery Specifications	Motor Specifications
Volkswagen e-Golf	2015	Li ion 26.5 kWh	3 phase AC synchronous 85 kW
Tesla Model S	2015	Li ion 85 kWh 375 V	3 phase AC Induction Motor 310 kW
BMW i3	2014	Li ion 22 kWh	3 phase synchronous motor 130 kW
Chevrolet spark EV	2014	Li ion 21.3 kWh 330 V	3 phase permanent magnet motor 85 kW
iMiEV	2013	Li ion 16 kWh 330 V	3 phase permanent magnet motor 47 kW
Toyota RAV4 second generation	2013	Li ion 41.8 kWh 386 V	3 phase AC Induction Motor 115 kW
Nissan Leaf	2013	Li ion 24 kWh	3 phase synchronous motor 80 kW
Hyundai Blue ON	2009	Li polymer 16.4 kWh	3 phase synchronous motor 61 kW
REVAi	2008	Lead Acid 9.6 kWh 48 V	3 phase AC Induction Motor 13 kW
GMEV1	1999	NiMH 26.4 kWh 312 V	3 phase AC Induction Motor 116.4 kW

Given the specific need to improve the performance of EVs for personal transport, this study principally focuses on some novel architectures of the battery and drive system of EVs that are of potential commercial interest for such consumers. The performance of any EV possesses a high dependency on the battery electrochemistry. In the nascent stages of the development of EV, batteries like lead acid and Nickel metal hydride were commonly employed. Further development in the electrochemistry field had rendered

such battery systems obsolete. The specific requirement of any EV is a battery pack with the lowest possible volume and weight matched with the highest possible power and energy. With the development of lithium based cells, a battery system that promises high energy density per volume and weight, the EV industry has undergone a shift. Most of the EV manufacturers now prefer to use Lithium based electrochemical cells in the battery pack as it can be observed from Table 1.1.

Power electronics is considered as one of the important discipline associated with the development of EVs. The various roles of power electronics in the transportation electrification especially in battery charging, electric motor drive etc. are detailed in [1]. The trends and functions of various configurations of power electronic subsystems in an EV are reviewed in [2]. A conventional configuration of battery pack with inverter and drive is shown conceptually in Fig. 1.1. As portrayed, power electronics plays an important role in the appellations of on-board charger, DC/DC converter, DC/AC converter in a typical EV power train.

Fig. 1.1 A typical power train configuration in a BEV

The battery pack has to be recharged to regain its depleted energy in order to run the vehicle again. It is analogous to refuelling a typical gasoline vehicle. Since the batteries employed in EVs are of high power rating and energy, recharging them requires a considerable amount of downtime for the EV. The downtime which refers to the time taken to refuel a vehicle is often considered as the crippling impediment in choosing an EV over the conventional fossil fuel powered vehicle. Consumers who may shrink from

using EVs due to this major disadvantage represent a loss of vital clientele for the electrified transportation sector. Typical downtimes required by the fossil fuel powered vehicle, an EV with the level 1 charging, an EV with the level 2 charging and an EV with a battery swapping feature is contrasted in Fig. 1.2.

Fig. 1.2 Vehicle downtime comparison chart

Usually, the existing conventional gasoline vehicles have a downtime of less than 5 minutes. Contrastingly, the electrochemical battery based EVs require a downtime in the order of hours. As an improvement over the level 1 slow charging which needed a downtime of 8-12 hours, level 2 fast charging was developed which reduced the downtime to approximately 2-3 hours [3]. Battery swapping significantly reduces the downtime. Adopting this solution would not diminish the convenience currently indulged by the users of fossil fuel powered vehicles for they would now have comparable downtimes [4, 5]. Therefore, this solution which overcomes the very significant disadvantage of longer downtime periods, promises to revolutionize the EV transportation. Usually, the EVs have the battery packs attached permanently to it. But the battery swapping system possesses the attribute of detachable and replaceable battery pack. The battery pack is detached from the vehicle when the battery energy is completely depleted and replaced with another charged battery pack. The replaced battery pack is charged separately at external charging stations. This feature of external charging is the primary reason for this very significant reduction in the downtime. The other dimension of

research reported in literature to reduce the downtime is by supplying energy to the vehicle from road through the contactless power transfer system [6, 7].

In addition to the range anxiety, the other main inhibition in EV adoption is the reliability and availability of the EV. The battery pack is the single point of energy source in a BEV. Hence the vehicle operation solely depends on the battery pack reliability. Lithium based cells for EV applications are manufactured by many companies and each company has its own manufacturing process which results in cells possessing different specifications in terms of volume, weight etc. However, one broad common specification of the Lithium ion cells is the three categories of form factors namely cylindrical, prismatic and pouch. The following, Table 1.2 lists the major specifications of some of lithium ion manufacturer's products for usage in the battery of the BEVs. As the lithium electrochemical cells commonly have a voltage of about 3.3 to 4.2 volts, a number of cells are usually interconnected to form a high power battery pack. The typical details of a battery pack construction and testing of battery pack are given in [8]. The design objectives of the battery pack for any EV primarily depends on these factors: the power required for maximum speed of the vehicle and the energy required covering the rated distance or range after one complete charging of the EV's battery [9]. A sample calculation of number of cells required to form a battery pack of 40 kW was performed and the values are given in Table 1.3.

Simply saying, the battery pack is built with numerous units of electrochemical cells dictated by the power and energy requirement. The effects of inter-cell contact resistances on the performance of battery pack are studied in [10, 11]. With single battery pack, the operation of the vehicle relies completely on one energy source. Although the electrochemical cells are of high quality, the reliability of the battery pack might be compromised even because of the connectors. Any fault in the battery pack affects the vehicle availability. A solution to increase the reliability of the system is to have modularity approach in the control of power extraction from the battery pack. This modular approach in designing the battery pack increases the availability of the system even in case of failure of one or more modules, and hence increases the reliability of operating an EV. Instead of having a one-size fits all battery pack, battery cells can be connected as micro-packs or modules. Therefore, modularization of the single pack to multiple micro-packs and controlling the power independently would have a significant influence on the operational reliability of an EV. By utilizing multiple micro-packs, it is

expected that any failures that would have become debilitating in single battery pack EVs becomes manageable with a multiple micro-pack EV. This research study considers a three various means of controlling the power/energy transfer from the available battery pack to the electric motor inside EV, primarily focusing on the operation and control of power converters employed.

Table 1.2 Lithium ion cell specifications [Source: Online]

Cell Manufacturer/ Model	Type	V	Ah	Weight (kg)	Radius (mm)	Length (mm)	Volume (cu.m)
A123/ANR26650	Cylindrical	3.3	.3	0.07	25.85	65.15	0.00013 66
A123/AHR32113	Cylindrical	3.3	4.5	0.21	32	113	0.00036 33
		V	Ah	Weight (kg)	Length (mm)	Width (mm)	Height (mm)
Thunder sky/ TS-LFP40AHA	Prismatic	3.2	40	1.6	116	46	190
Thunder sky /TS-LFP60AHA	Prismatic	3.2	60	2.5	115	61	203
A123/AMP20	Pouch	.3	20	0.496	160	7.25	227

Building a battery pack requires a robust mechanical design which takes into consideration the safety and cooling requirements. The mechanical design becomes more complex in the design process of micro-packs as the previously employed safety constraints are imposed. Such practical constraints fall outside the scope of this research. Rather, this research report examines the power converters and their control perspective when the energy storage system is modularized. With modularized battery micro-packs, the need for an individual power converter arises to control the power independently. One possible configuration using DC/DC converter for every battery micro-pack is shown in Fig. 1.3. Given that a DC/DC converter of high power becomes bulky, this potential set up of the battery micro-pack is not considered for analysis. Moreover, operation and

control of parallel connected DC/DC converters have been thoroughly addressed in literature for interconnecting multiple as well as heterogeneous energy sources [12-16].

Table 1.3 Sample calculations of number cells required to build a battery pack

Cell Manufacturer / Model	Number of cells in Series (Ns=400/cell voltage)	Ns rounded off (Ns')	Number of cells in parallel (Np=100/Ah)	Np rounded off (Np')	N=Ns'+Np'	Net Weight (kg)	Net Volume (cu.m)
For Battery pack of 400 volts , 100 amperes							
A123/ANR26650	121.21	125	43.47	5	5625	393.75	0.7689
Thunder sky/ TS-LFP40AHA	125	125	2.5	3	375	600	0.3801
A123/AMP20	121.21	125	5	5	625	310	0.1645
For Battery pack of 330 volts , 122 amperes							
A123/AHR32113	100	100	27.11	30	3000	615	1.0900
Thunder sky/ TS-LFP40AHA	103.13	105	3.05	3	315	504	0.3193
A123/AMP20	100	100	6.1	6	600	297.6	0.1579

In this work, Cascaded H-bridge (CHB) multi-level converter (Fig. 1.4) and Double Star Chopper Cell (DSCC) based multi-level converter with modularized energy storage system (Fig. 1.5) for BEV motor drive is investigated. Operation and control of CHB and DSCC for medium voltage applications based converter has been well studied in the literature. The problems associated with the CHB and DSCC converters for BEV has to be investigated.

Fig. 1.3 One way of battery micro-packs and inverter arrangement in an EV

With independent power sources, cascaded H-bridge (CHB) multi-level converter is a potential viable candidate for the power converter. This CHB based motor drive arrangement is shown in Fig. 1.4. The CHB is a special kind of multi-level converter that has been researched upon and utilized for high power applications like Stat-Coms. A part of this research report examines the usage of the CHB based motor drive for EV applications. State of Charge (SOC) is the measure of the available battery capacity in an electrochemical cell. This measure is extended to modules and to the battery pack in which SOC is the average of SOC of each cell in battery pack. In any battery energy storage system, the most common problem addressed by many researchers is the imbalance of SOC of cells in a particular module. This imbalance in the SOC of cells in any module drastically reduces the available capacity of other cells in that battery pack and thence effects a commensurate reduction in the capacity of the whole module. This imbalance in the SOC of cells within a battery module or a pack can be rectified by judiciously employing some dedicated circuits. Another important feature of the analysis in the imbalance of the SOC of cells is the variation of the same between different micro-packs. This imbalance between SOC of the micro-packs also requires to be addressed as it reduces the performance of the whole system.

To combine the advantage of replaceable with the modularization, the novel wireless power transfer concept is considered. Modular battery packs or micro-packs with Inductive Power Transfer (IPT) interface is investigated in this work considering that it

would make the battery swapping process more reliable, faster and convenient as the electrical locking mechanism in conventional battery swapping method could be eliminated. The conceptual representations of the proposed systems are shown in Figs. 1.6 and 1.7. The first topology is a contactless and replaceable BESS with multiple module-casings and is conceptually depicted in Fig. 1.6. The second topology is a contactless and replaceable BESS with one casing and multiple modules and is conceptually depicted in Fig. 1.7. Also the weight of the micro-pack would be much lesser compared to the whole battery pack which would bring in a significant reduction in the complexity associated with the design of battery exchange stations.

Fig. 1.4 Cascaded H-bridge based power train for BEV

Fig. 1.5 Double Star Chopper Cell DSCC-MBESS based power train for BEV

Fig. 1.6 Type 1: Inductive Power Transfer IPT-MBESS based power train for BEV

Fig. 1.7 Type 2: Inductive Power Transfer IPT-MBESS based power train for BEV

1.2 Major Contributions of this Thesis

With the motivations discussed in the previous section, the objectives of this research study can be stated as follows:

- To control the power flow effectively and independently when the battery system is modularized into micro-packs to increase the operational reliability of the vehicle
- To facilitate an efficient battery swapping process with the modularized battery micro-packs

Based on these objectives, a research work has been conducted and reported in this thesis. The major contributions of this thesis are as follows:

Proposed an inter-module State-of-Charge (SOC) balancing for Cascaded H-bridge (CHB) based Battery Energy Storage System (BESS). CHB is considered as an apt choice for modular power converter for independent power control of modular battery system. A fault tolerant control realization for CHB using Multi-Dimensional Pulse Width Modulation (MD-PWM) is presented. The proposed SOC balancing and fault tolerant operation are conveniently integrated with the PWM generation algorithm. The main benefit of the proposed control method is that it aids in improving the driving range in case of fault and unbalanced SOC conditions. Though the proposed methods are discussed for EV application, it can be applied to other applications like grid connected ESS.

Proposed a Double Star Chopper Cell (DSCC) based converter topology for electric vehicle BESS. The Multi-Dimensional Pulse Width Modulation (MD-PWM), originally proposed for CHB, is extended for the operation of DSCC. The integration of fault tolerant control with the PWM generation algorithm is proposed. The SOC balancing control using MD-PWM is developed for DSCC. The fault tolerant control and inter-module SOC balancing realization for DSCC MD-PWM is presented. The proposed converter topology and the modulation strategy integrated with fault tolerant control and the SOC balancing can also be extended for medium voltage applications.

Proposed two novel systems of Inductive Power Transfer based contactless and swappable BESS. The proposed systems would aid in a safe and reliable battery swapping for EVs, and hence helps in reducing the down-time or waiting time. The proposed systems are investigated from the power converters operation and the Inductive Power Transfer (IPT) perspective. The operation of the first topology is analysed with an equivalent circuit and an averaged state-space model. A feedback linearization technique to establish the DC link voltage regulation based on the phase shift ratio between the primary side and the secondary side voltages is implemented. The proposed dynamic modelling approach, and control method using feedback linearization can be extended to any application employing the IPT technique. The second topology is analysed with an equivalent electric circuit and the expressions for power flow are derived using steady state analysis of the equivalent circuit.

1.3 Organization of this Thesis

This thesis comprises of seven chapters, and the organization of which are summarized as follows:

A brief introduction about the electric vehicle and the battery pack has been given in the beginning of Chapter 1. The motivation of this research study has also been discussed. The objectives and the contributions of this research study have been introduced in the later part of this chapter. Chapter 2 briefly reviews the works related to this research study that were reported in the literature.

A Cascaded H-bridge (CHB) based BESS for an EV is discussed in Chapter 3. A module level SOC balancing control in CHB based BESS using multi-dimensional pulse width modulation technique (MD-PWM) is proposed. A design concept for the battery micro-pack construction for the symmetrical CHB converter in EV application is described. The MD-PWM strategy for CHB converter is explained. It is a generalized modulation strategy for cascaded H-bridge inverters. The proposed inter-module SOC balancing and fault tolerant control using MD-PWM are described. The operation of the proposed system and control methods are verified through experimental results from the laboratory hardware prototype and the results are discussed. The operation of CHB as rectifier for battery micro-pack charging is also discussed using simulation results.

A Double Star Chopper Cell (DSCC) converter based BESS is proposed and investigated for electric vehicle application in Chapter 4. A design concept for modularizing the battery pack into micro-packs for DSCC is described. MD-PWM for DSCC is introduced. The methods of operation for inverter using MD-PWM with integrated inter-module SOC balancing and fault tolerant control are described. The method of operation as synchronous rectifier using MD-PWM is discussed. The proposed system operation as inverter and synchronous rectifier are verified through simulations and the results are discussed.

Two novel systems of Inductive Power Transfer based contactless and swappable BESS for EV application are proposed and investigated in Chapter 5. The Type 1 and Type 2 of the proposed modularized battery systems with inductive coupling are detailed. Design guideline and simulation results of the proposed micro-pack system with inductive coupling are presented. A laboratory prototype for each type of the proposed system is implemented to demonstrate the feasibility of the proposed concept. Details of the

experimental setup and the simulation results as well as experimental verification of the proposed systems under different operating conditions are presented.

A comparative analysis of the proposed modular BESS based on the number and capacity of the lithium-ion cells required to construct the micro-pack, and the input current signature depending on the switching strategy are investigated in Chapter 6.

Chapter 7 concludes this thesis, and the recommendations for future research are discussed. Followed by this chapter are the List of author's publications, Bibliography and Appendices related to this research work.

Chapter 2 Literature Review

2.1 EV Charging

EV charging can be broadly classified into conductive charging and wireless charging [17-20]. Conductive charging refers to charging the EV battery pack by connecting either to an external charger or power supply equipment (for the vehicles with on-board charger) through a charging cable. Wireless charging refers to charging the EV battery pack by means of any of the wireless power transfer methods like inductive power transfer with or without magnetic resonance. Wireless charging can be either stationary or dynamic. Stationary wireless charging requires the vehicle to be parked at the designated wireless charging bed whereas the dynamic charging refers to the charging of EV battery back when the vehicle is in motion over a specially constructed road line. Online electric vehicles (OLEVs) [21, 22] can be included in the category of dynamic wireless charging. OLEVs can be considered having literally zero downtime if only the vehicle operation is considered and the power is transferred from the electrical roadbed to the drive directly. The disadvantages with OLEVs are the power transfer efficiency, and it is not feasible to provide electrical roadbed at all places.

Based on the time taken to recharge an EV battery and the power levels of the chargers, the charging methods are categorized as Level 1 (or slow), Level 2 (or medium) and Level 3 (or fast). Society of Automotive Engineers (SAE) and International Electro technical Commission (IEC) are coming up with standards for the EV charging [23-25]. Level 1 includes all the lower power chargers which require a downtime of more than 8 hours. As an alternative to level 1 slow charging, level 2 charging was developed. Level 3 includes very high power chargers capable of charging the battery pack in less than 30 minutes. A chart representing the classification of EV charging is shown in Fig. 2.1. The battery swapping methods have been under investigation in order to achieve further reduction in downtime and are considered as ultrafast when compared with the other charging methods. Typically, EVs are permanently equipped with battery packs. The battery swapping system allows for the battery packs to be replaceable. The battery pack is detached from the vehicle when the battery energy is completely depleted and replaced with another fully charged battery pack. This battery swapping or replacement can be done in designated stations. Battery swapping or exchange stations are designed

especially for certain type of vehicles. A Battery exchange station holds certain number of charged battery packs. When an EV reaches the exchange station, the battery pack is replaced.

Fig. 2.1 Categories in EV Charging

An algorithm for such a kind of automated battery replacing was given in [26]. A number of electrical and mechanical locking systems keep the battery pack attached with the vehicle chassis. These locking mechanisms are operated sequentially during the battery swapping process. Tesla motors demonstrated a successful battery swapping for its vehicle [27] in which the whole battery pack is replaced in 90 seconds. The replaced battery pack is recharged separately at external charging station. This external charging feature of battery swapping is the primary reason for the significant reduction in downtime. EV use is expected to increase with the availability of such convenient exchange stations. The reduction in downtime not only makes it an attractive option for personal vehicles but also for vehicles used in the public transportation system, including e-buses and e-taxis. The electrical interface between battery and vehicle chassis is a critical component in the design of the replaceable battery system. A good design of electrical interfaces prevents the creation and furtherance of hazards during swapping process. The mechanical vibration that is inherent in any vehicle motion may cause sparks between the electrical contacts. Therefore it is desirable to develop an appropriate contactless electrical interface system. Hence, the use of IPT for modularized energy storage in EV is investigated in this work.

Power converters play an important role in all the types of EV charging mentioned earlier. Various kinds of power converter topologies and control concepts have been proposed in literature. A review of various topologies of EV chargers was discussed in [3]. The architectures of EV chargers vary with the category it belongs. Level 1 chargers get its power input from a single-phase house hold power supply, whereas high power Level 2 and Level 3 chargers require three-phase power supply. Various designs of wireless charging topologies for EV application have been reported in literature [28]. Some researchers have also proposed integrated charger topologies utilizing three-phase two-level inverter and the electric motor in the power train to charge the EV battery pack [29-31]. In most of the commercially available electric vehicles, the power converters for the battery pack charging and the inverter for motor drive are two different components. In this work, modular power converters for the motor drive were investigated. The charging operation of the battery pack is not much investigated in this work, as the synchronous rectifier operation for battery charging is similar to the inverter operation. It is assumed that external three-phase power supply equipment is available for battery charging in case of the CHB and DSCC based systems. For the IPT based proposed systems, it is assumed that the charging of battery pack would be done externally to vehicle in a battery exchange station.

2.2 EV Discharging (Power Train)

A typical representation of a power train in a BEV is portrayed in Fig. 1.1. Three-phase AC motor is commonly used in any BEV due to its high efficiency and power density compared to DC motor of same power rating. It can be observed from the specifications of commercially available electric vehicles tabulated in Table 1.1 that three-phase AC Induction Motor (IM) and Permanent Magnet Synchronous Motor are widely used in BEV applications for commercial usages. Apart from these two commonly used motors, other types of motors like Brushless DC Motor (BLDC) [32], Switched Reluctance Motor (SRM) [33] have also been proposed for EV applications. The propulsion motor can also be a single motor coupled to the wheels through a mechanical transmission system like the one portrayed in Fig. 1.1, or it could be in-wheel motors like the ones proposed in [34]. A three-phase inverter unit in the power train is to generate necessary voltages to drive the motor. A two-level inverter shown in Fig. 2.2 is the basic

and conventionally used inverter topology for BEV application. The operation and control of the three-phase two-level inverter for any motor drive has been researched and developed thoroughly for many years. Various kinds of modulation strategies namely six pulse modulation, sinusoidal pulse width modulation, space vector modulation etc., are being used for the two-level inverter topology [35].

Fig. 2.2 Three-phase Two-level inverter

2.3 Cascaded H-bridge (CHB) Multi-level Converter

Cascaded H-bridge (CHB) is one of the important members of the family of multi-level converters [36] including diode clamped converter and capacitor clamped converter, which has its application in medium voltage electric drives [37-39], STATCOMs [40-43] etc. CHB has many advantages like modularity, low dv/dt and low voltage Total Harmonic Distortion (THD). However, the main disadvantage of CHB is that it requires isolated DC power supplies and transformer. With the approach of employing battery micro-packs for CHB, this particular disadvantage is circumvented. Many topological variations of CHB have been reported in literature. A three-phase CHB could be either star connected or delta connected. CHB can be classified as symmetrical having equal input DC voltages, and asymmetrical having unequal input DC voltages. The symmetrical star connected CHB based converter with modularized battery system is shown in Fig. 2.3, in which a generalized N H-bridge units per phase leg is illustrated. For the symmetrical CHB having N number of H-bridges in a phase, the voltage levels L in each phase is given by $L=2N+1$. N is usually called as cascade number.

Fig. 2.3 Cascaded H-bridge converter configuration

Various modulation techniques like phase shifted multi-carrier PWM (PS-PWM), level shifted multi-carrier PWM (LS-PWM), Space Vector modulation (SVPWM), staircase modulation, selective harmonic elimination (SHE-PWM) etc., are available for CHB multi-level converter [44-46]. A generalized modulation strategy called multi-dimensional PWM (MD-PWM) for CHB converter proposed in [47], is employed in this work to investigate the possibility of module level SOC balancing and fault tolerant operation.

Various fault tolerant operation techniques for CHB converter have been discussed in [48-52]. Fault in multi-level converters generally refers to the fault in the power electronic switches. In case of CHB, any fault in the power electronic switch requires that particular H-bridge to be bypassed. It requires additional bypass switch for every H-bridge unit. However, in this work, the bypassing of H-bridge is realized using the zero switching state of the H-bridge. Irrespective of the bypassing technique, all the fault tolerant techniques aim at operating the CHB to deliver a balanced three-phase current even when some of the H-bridge units are bypassed and not in operation. This has been achieved in two ways. First method is described as ‘without neutral shift’, achieved by disabling one H-bridge unit in the healthy phase legs for every faulty H-bridge, and distributing the load to remaining healthy H-bridge units. Second method is described as ‘with neutral shift’, achieved by adding zero sequence voltage or common mode voltage to the phase voltage reference to generate balanced line-line voltages. The second method

is more effective as it does not reduce the capacity of the system, and methods to obtain the zero sequence voltage for PS-PWM have been discussed in [49, 53]. A space vector modulation based fault-tolerant operation of CHB converter utilizing redundant switching state has been proposed in [52, 54]. In this report, a method for realizing the fault tolerant operation ‘with neutral shift’ using MD-PWM is presented.

The fault tolerant control operates the CHB in such a way that uneven amount of power are being extracted from the micro-packs, which subsequently results in an imbalance condition among their SOC. The imbalance in SOC of micro-packs is also possible due to change in the battery characteristics over aging, or even battery micro-packs could have been charged to different SOC levels. The entire system has to be stopped even when one of the micro-pack’s SOC reaches the minimum cut-off value. This imbalance in the SOC is undesirable in EV applications, as it affects the driving range of the vehicle. A module level SOC balancing for a grid connected CHB battery energy storage system based on modifying the phase voltage reference with the zero sequence voltage injection and PS-PWM has been demonstrated in [55]. A novel control method for the SOC balancing and micro-pack fault tolerant operation that can be conveniently integrated with the PWM generation algorithm is proposed in this report. Thus the proposed control methods do not require any extra hardware components. The proposed control method based on MD-PWM and reference phase voltage amplitude is convenient for digital implementation.

2.4 Double Star Chopper Cell (DSCC) Multi-level Converter

DSCC is a kind of modular multi-level converter (commonly abbreviated as MMC or M2C). MMC has a single DC supply and each sub-module being connected with a capacitor eliminates the need of a single larger DC link capacitor. The applications of MMC are usually in HVDC and STATCOMs. Various control methods for MMC based on DSCC have been presented in the literature with focus on capacitor voltage balancing, fault tolerant operation etc., [56-60]. The DSCC based converter with modularized battery system is shown in Fig. 2.4 where a generalized ‘n’ number of micro-packs per arm are illustrated.

Fig. 2.4 Double Star Chopper Cell (DSCC) converter configuration

The difference between the conventional DSCC based MMC and the proposed system is that each sub-module in DSCC is fed with battery micro-pack instead of capacitors and single larger DC supply. A configuration of DSCC with chopper cell for each electrochemical cell in a battery pack was discussed in [61, 62], whereas in this work, a chopper cell for a battery module or micro-pack with certain number of cells is investigated. Any modulation method applicable for a multi-level converter can be extended to DSCC. For instance, an improved PWM based technique was given in [63]. The Multi-dimensional Pulse Width Modulation (MD-PWM) was proposed for cascaded H-bridge multilevel converter [47] and has been modified for DSCC in this work. The MD-PWM with integrated inter-module SOC balancing and fault tolerant control for DSCC with battery micro-packs is proposed.

2.5 Inductive Power Transfer

Wireless power transfer concept has been utilized in myriad varied applications ranging from biomedical equipment to electric vehicle charging. It includes hand phone battery charging, EV battery charging, on-line EVs, industrial automation applications, robotic applications etc., [22, 64-70]. There are two common types of wireless power transfer viz., magnetic resonance power transfer and inductive power transfer. Magnetic resonance type power transfer refers to the transference of power over long distances. The operating frequency of these types of systems ranges in MHz [71-73]. Contrastingly, the inductive type power transfer (IPT) system refers to the transference of power over smaller distances (ranging in millimeters/ centimeters). Consequently, such systems exhibit an operating frequency which ranges in kHz [74-77].

Among various wireless power transfer techniques, IPT is found suitable for the proposed application. Numerous topologies of IPT systems, control strategies and design techniques have been reported in literature. Most of them were addressing the power transfer efficiency improvement from the perspectives of coil design [78], reactive power compensation [79], control strategy [80] etc. The topologies of IPT systems proposed in [67-69, 81] clearly demonstrate the feasibility of IPT systems for EV charging. The IPT system can also be used for bi-directional power transfer and was demonstrated in [81]. This implies that the IPT systems are also suitable for auxiliary functions of EV such as V2G. Modular and redundant low power IPT systems for high power applications were studied in [82] and [83]. In [82], the secondary coils of modular IPT systems were connected in parallel whereas in [83] various modular low power IPT systems were paralleled using intermediate LC network.

The proposed application of the IPT system requires power to be transferred in both directions across the inductive interface. The LCL compensated IPT system topology, shown in Fig. 2.5 and similar to the one proposed in [81] having the capability of controlling the power flow bi-directionally, is used as basic building blocks in the proposed systems. All the LCL components of the system are designed at the angular resonant frequency given by $\omega = 1/\sqrt{LC}$. The IPT system has an H-bridge power converter on either side to produce necessary voltage. The primary side converter is operated as inverter generating a square wave or quasi square wave V_m . The average value of V_m is controlled using the phase shift between the outputs of two legs of the H-bridge

converter. The secondary side converter operation is similar to that of primary side converter, operated as reversible rectifier for the bi-directional power control. The amount and the direction of power transferred between the primary and secondary sides can be controlled by manipulating either the voltage magnitude of the primary side converter V_m or by manipulating the relative phase angle between the voltages of primary (V_m) and secondary side (V_c).

Fig. 2.5 A Typical LCL IPT model

2.6 Summary

A brief overview on conventional methods of charging the EV batteries and EV power train has been presented. Cascaded H-bridge converter topology for modular battery pack system and a brief review of which has been presented. Fault tolerant operation of CHB converter is commonly based on Sinusoidal PWM and Space Vector modulation. A simpler and easier method of realizing fault tolerant operation using MD-PWM is proposed in this work. Furthermore, integrating the SOC balancing algorithm with fault tolerant control and modulation algorithm is proposed in this work. Double Star Chopper Cell converter topology based modular battery system is proposed for EV application in this work, and a brief review of which has been discussed. An overview about the Inductive Power Transfer concept has been presented. Generally, the IPT systems have been investigated and employed for battery charging or on-line EV applications. However, in this study, the IPT systems are investigated for the implementation of contactless battery system in EVs.

Chapter 3 Cascaded H-bridge Converter based Modularized Battery Energy Storage System (CHB-MBESS)

3.1 Introduction

Conventional battery electric vehicles (BEV) have a single battery pack with a two-level inverter for motor drive. Any fault in the battery pack or the two-level inverter affects the availability of the vehicle with the conventional design having a single battery pack. A modular battery pack or micro-pack would circumvent this problem and thus could be considered to increase the availability of the vehicle in case of fault. The reliability and availability of the vehicle would be increased when the power in each battery micro-packs is controlled independently and also when the system possesses ride through capability under fault conditions. This is achievable using modular power converters. One such system employing cascaded H-bridge (CHB) multilevel converter shown in Fig. 3.1 is discussed in this chapter.

This chapter presents a method for module level State-Of-Charge (SOC) balancing and fault tolerant operation of the CHB converter based battery energy storage system for EV application. The proposed methods are based on the multi-dimensional PWM (MD-PWM) strategy. It is shown that the proposed methods are conveniently integrated with the PWM generation algorithm. A conceptual design of modular battery pack (or micro-pack) system using lithium-ion cells for the CHB converter is described to show its feasibility for EV application. A brief review on the MD-PWM strategy and a detailed explanation on the method to integrate the inter-module SOC balancing and fault tolerant control with the PWM generation are presented.

A design concept for the battery micro-pack construction for the symmetrical CHB converter in EV application is described in Section 3.2. The MD-PWM strategy for CHB converter is briefed in Section 3.3. The proposed inter-module SOC balancing and fault tolerant control using MD-PWM are described in Section 3.4. The operation of the proposed system and control methods are verified through experimental results from the laboratory hardware prototype and the results are presented in Section 3.5. Battery micro-

packs charging is achieved through synchronous rectifier operation of the CHB, which is explained using simulation results in Section 3.6.

Fig. 3.1 CHB for MBESS

3.2 Battery Micro-pack Design Concept

This section describes a method for designing a modularized battery micro-packs using lithium ion cells for the CHB converter to show its feasibility for the EV application. The basic design parameters of the EV battery pack are the maximum power required for the maximum acceleration with the given vehicle weight, and the maximum energy capacity of the battery pack required for certain driving range. The voltage of the battery pack should be in accordance with the rated voltage of the drive motor. In order to generate a line voltage with peak value of V_p with CHB converter, the required micro-pack voltage is given by,

$$V_{dc} \geq \frac{V_p}{\sqrt{3}N} \quad (3.1)$$

where N is the number of H-bridge units with battery micro-pack in a given phase leg of the CHB converter. The rated power of the battery pack should be not less than P_{acc} , the power required for the maximum acceleration of the vehicle. To achieve this maximum

power with CHB converter and micro-packs, the rated power of each micro-pack should be

$$P_{bat}^{max} \geq \frac{P_{acc}}{3N} \quad (3.2)$$

With V_{dc} and P_{bat}^{max} using (3.1) and (3.2), the number of cells and the capacity (Ah) of the lithium ion cell for a battery micro-pack can be manipulated. In this work, the chosen values for investigation are $V_p = 400$ V and $P_{acc} = 100$ kW. These values are considered for the proposed design of micro-pack, so that it can be compared with the conventional battery pack design of a commercially available electric car: Nissan Leaf battery pack (24 kWh and 360 V) comprising of 192 Lithium ion cells [84]. The battery pack has the rated power of about 100 kW. The required V_{dc} should be ≥ 115.5 V, 77 V, 46 V, and the required P_{bat} should be ≥ 16.67 kW, 11.11 kW, 6.6 kW for the CHB with $N = 2, 3, 5$ respectively. For these V_{dc} and P_{bat} , the rated current of the micro-pack I_{bat} should be ≥ 144 A. Any automotive lithium ion cell with nominal voltage of 3.7 V and the nominal capacity of 35 Ah with maximum discharging capacity of at-least 5C can be chosen. With such lithium cell, the micro-pack can be constructed with 32, 21, and 13 cells in series for $N = 2, 3$, and 5 respectively. The total capacity of such CHB system would be ~24 kWh with each micro-pack being 4.14, 2.72, and 1.68 kWh for $N = 2, 3$, and 5 respectively. The parameters of the proposed micro-pack design and that of Nissan leaf battery pack are compiled in Table 3.1

Table 3.1 Parameters of the battery pack

Parameters	Nissan Leaf	CHB ($N=2$)	CHB ($N=3$)	CHB ($N=5$)
Total number of cells	192	192	189	195
Nominal Capacity of each cell (Ah)	60	35	35	35
Total Energy storage capacity (kWh)	24	24.8	24.5	25.3
Battery Pack Nominal Voltage (V)	360	118.4	77.7	48.1

It can be seen that the micro-packs for CHB could be constructed with almost same number of lithium ion cells, even with reduced capacity compared to the conventional single battery pack system with two-level inverter. It has to be noted that the design of Battery Management System (BMS) and thermal management would become complex with micro-pack system.

3.3 CHB and Multi-Dimensional PWM

Multi-dimensional pulse width modulation (MD-PWM) proposed in [47] is a generalized modulation approach for a CHB converter, which can be particularized to operate like PS-PWM or LS-PWM. The MD-PWM technique is based on the geometric representation of the output voltage of each H-bridge in co-ordinate axes as shown in Fig. 3.2. For the cascade number of two ($N = 2$), the control contour is a straight line bounded by a square, whereas for the cascade number of three ($N = 3$), the control region is a plane bounded by a cube. For cascade number more than three, the control region cannot be represented visually, but the concept can be extended.

The phase voltage of an i^{th} phase leg averaged over a period is the sum of average voltage outputs of j H-bridge units and is given by,

$$V_i = \sum_{j=1}^N V_{cell,j} \quad (3.3)$$

Therefore solution to (3.3) for any desired value of V_i is a point on the line for $N = 2$, and a point on the plane orthogonal to $[+V_{dc,1} ; +V_{dc,2} ; +V_{dc,3}]$ vector for $N = 3$ as shown in Fig. 3.2. As such, there can be infinite number of solutions to any desired value of V_i . This is a specific advantage of MD-PWM, since it provides an extra degree of freedom to choose a $V_{cell,j}^*$ depending on a specific design criterion. For instance, MD-PWM was used to control dc-link voltage in a CHB based synchronous rectifier in [85].

Once $V_{cell,j}^*$ is chosen for a desired phase voltage V_i^* in a switching period (T_{sw}), the next step is to determine the switching time and the switching sequence. The switching period is identical to the carrier wave frequency in SPWM. Each H-bridge can take three switching states $S \in (-1, 0, +1)$ which corresponds to the output voltage of $(-V_{dc}, 0, +V_{dc})$. The switching time is calculated using the equations (3.4) and (3.5).

Fig. 3.2 Control contour representation for MD-PWM

$$t_{on,j} = \frac{V_{cell,j}^*}{V_{dc,j}} \times T_{sw} \quad (3.4)$$

$$t_{off,j} = T_{sw} - t_{on,j} \quad (3.5)$$

During the time $t_{on,j}$, the j^{th} H-bridge cell is clamped to either state $S = +I$ for positive $V_{cell,j}^*$ or to state $S = -I$ for negative $V_{cell,j}^*$. And during the time $t_{off,j}$, the j^{th} H-bridge cell output is maintained zero which is state $S = 0$. The MD-PWM is easy to implement and can be extended to CHB of any levels.

3.4 Micro-pack SOC balancing and Fault Tolerant Control using MD-PWM

A phase leg in CHB has N H-bridge units each with a battery micro-pack. Let SoC_i be the average SOC of N battery micro-packs' SOC's (SoC_{i1} to SoC_{iN}) in phase i , with SoC_{ij} being the SOC of j^{th} micro-pack in phase i for all $i \in (A, B, C)$. And SoC is the average of SoC_A , SoC_B and SoC_C , which represents the SOC of the whole system.

$$SoC_i = \frac{1}{N} \sum_{j=1}^N SoC_{ij} \quad (3.6)$$

$$SoC = \frac{1}{3} (SoC_A + SoC_B + SoC_C) \quad (3.7)$$

The deviation of SoC_A , SoC_B and SoC_C from SoC is given by,

$$\Delta_i = SoC - SoC_i \quad (3.8)$$

The deviation of j^{th} micro-pack SOC in i^{th} phase (SoC_{ij}) from its average SOC (SoC_i) is given by,

$$\delta_{ij} = SoC_i - SoC_{ij} \quad (3.9)$$

3.4.1 Inter-phase SOC balancing

If all the phase legs have equal average SOC (i.e., $\Delta_A = \Delta_B = \Delta_C = 0$), then power extracted during motoring operation and the power delivered during charging operation should be distributed equally among three phases. But when the phase legs have difference in their average SOC values, the power extracted from or delivered to each phase leg should be varied according to their average SOC to achieve a balanced operation. This is achieved by adding or subtracting zero sequence voltage of fundamental frequency to the phase voltage references. The zero sequence voltage modifies the phase voltage generated, and hence affects the power extracted from each phase without affecting the output line-line voltage.

The MD-PWD requires the $V_{cell,ij}^*$ to be manipulated in every switching period. A particular design criterion can be used to manipulate $V_{cell,ij}^*$. In order to extract equal amount of power from each battery micro-packs, $V_{cell,ij}^*$ is calculated using,

$$V_{cell,ij}^* = \left(\frac{V_i^*}{N} \right) \quad (3.10)$$

where V_i^* is the desired average phase voltage over a switching period. To include the inter-phase SOC balancing control in MD-PWM, the reference for each H-bridge unit in (3.10) is modified into,

$$V_{cell,ij}^* = \left[\frac{(V_i^* - V_{bal}^*)}{N} \right] \quad (3.11)$$

where V_{bal}^* is the voltage modifier that represents the required zero sequence voltage at every switching period. It is obtained from the reference phase voltage amplitude V_i^* and the SOC deviation. The required voltage modifier to achieve the inter-phase SOC balancing is given by,

$$V_{bal}^* = -G_I [f_A \Delta_A V_A^* + f_B \Delta_B V_B^* (1-f_A) + f_C \Delta_C V_C^* (1-f_A)(1-f_B)] \quad (3.12)$$

where

$$f_i = \begin{cases} \text{floor} \left[\frac{(\Delta_i + |\Delta_i|)}{2^* \Delta_{max}} \right] & ; \forall i \in (A, B, C) \text{ and } |\Delta_{max}| > 0 \\ 0 & ; \forall i \in (A, B, C) \text{ and } \Delta_{max} = 0 \end{cases}$$

$$\Delta_{max} = \max(\Delta_A, \Delta_B, \Delta_C)$$

The V_{bal}^* follows the voltage reference of the phase leg having higher SOC deviation from the average SOC of three phase legs. Magnitude of the V_{bal}^* is determined based on the SOC deviation and the proportionality constant G_I . The proportionality constant G_I is the percentage of voltage reference that is required to be subtracted from each of the reference phase amplitudes. The value for G_I is obtained from the input power and SOC relation of each H-bridge unit given in Fig. 3.3(a). G_I is the representation of difference between the amounts of power delivered by each phase legs in CHB, which needs to be modified for certain time period to nullify their SOC differences. In Fig. 3.3(a), $P_{L,i}$ is the loss and P_i is the active power of i^{th} phase.

The closed loop transfer function is given by,

$$\frac{\Delta_i(s)}{P_L(s)} = \frac{\frac{1}{G_I P_i}}{(1+s \frac{NQ}{100 G_I P_i})} \quad (3.13)$$

where Q is the maximum capacity (in Joules) of each battery micro-pack. The value for G_I is manipulated from the time constant of the transfer function given by,

$$T = \frac{NQ}{100 G_I P_i} \quad (3.14)$$

For $N = 2$; $Q = 19008 \text{ J}$; $P_i = 250 \text{ W}$; if $G_1 = 0.3$, then time taken for SOC balancing among three phases would be 8.5 minutes theoretically.

Fig. 3.3 Control block diagram for SOC balancing - (a) Inter-phase SOC balancing (b) Intra-phase SOC balancing

3.4.2 Intra-phase SOC balancing

If all the micro-packs in a phase leg have equal SOC (i.e., $\delta_{i1} = \delta_{i2} = \dots = \delta_{iN} = 0$), then power extracted during motoring operation and the power delivered during charging operation should be distributed equally among all the micro-packs in a phase leg. This section explains a novel method that can be integrated with the PWM generation algorithm to achieve SOC balanced operation among the micro-packs within a phase leg. In order to vary the amount of power distribution with respect to SOC of individual battery micro-pack, (3.11) is modified to,

$$V_{cell,ij}^* = \left[\frac{(V_i^* + V_{bal}^*)}{N} \right] (1 - G_2 \delta_{ij}) \quad (3.15)$$

The parameter δ_{ij} modifies the $V_{cell,ij}^*$ in every switching period to vary the amount of power distributed among the H-bridge units cascaded in series. The parameter δ_{ij} is determined from the deviation of the j^{th} battery micro-pack's SOC (SoC_{ij}) from the

average SOC of all battery micro-packs in i^{th} phase leg. The value of G_2 is obtained in similar way as G_1 in the inter-phase balancing, based on the percentage of power that needs to be varied to nullify the SOC differences among the micro-packs in a phase leg. The control block diagram for intra-phase SOC balancing is shown in Fig. 3.3(b). From the closed loop transfer function, the proportionality gain G_2 is given by,

$$G_2 = \frac{Q}{100TP_{ij}} \quad (3.16)$$

For $Q = 19008 \text{ J}$; $P_{ij} = 125 \text{ W}$; if $G_2 = 0.4$, then SOC balancing among the micro-packs in a phase leg would take 6.3 minutes theoretically.

As the parameter δ_{ij} depends on the deviation of SoC_{ij} from SoC_i , δ_{ij} can be positive or negative i.e., δ_{ij} is positive when SoC_{ij} is smaller than SoC_i and δ_{ij} is negative when SoC_{ij} is greater than SoC_i . The minimum value that δ_{ij} can have is -1. But when $\delta_{ij} = -1$, then it is similar to faulty micro-pack and that corresponding H-bridge cell is clamped to zero voltage. This might cause voltage unbalance issues in 3-phase CHB converter, which requires additional control technique, and is explained in the following section. The maximum value of δ_{ij} depends on the modulation index and is determined based on the other outer control loop. The upper range of δ_{ij} is determined so that the corresponding H-bridge cell output voltage $V_{cell,ij} \leq V_{dc,ij}$. Hence the range of δ_{ij} can be defined as in (3.17).

$$-1 \leq G_2\delta_{ij} \leq \frac{N \times V_{dc,ij} - |V_i^*|}{|V_i^*|} \quad (3.17)$$

To prevent the inverter from entering into square wave modulating region, the upper cut-off of the parameter δ should be set to 0.278 so that the maximum modulation index becomes $4/\pi$.

3.4.3 Ride through operation under faulty condition

The motivation for investigating modular converters with micro-packs is to achieve the ride through capability under fault conditions, which is not feasible with the conventional single battery pack and the two-level inverter system. Various fault tolerant operation methods was proposed in literature for multilevel and modular converters [49,

53, 54]. The effective method of operation would be to distribute the load equally among the healthy micro-packs. The motivation in this study is neither on type or place of the fault occurrence nor on the fault detection methods. Instead, a fault tolerant operation realization using MD-PWM is studied in detail. Furthermore, it is assumed that bypassing an H-bridge could be possible using the zero switching state ($S = 0$). The following method is derived assuming that the fault signals are available from any of the real time state of health monitoring systems like the ones proposed in [11, 86]. In-order to achieve ride through capability under faulty condition, the reference voltage for each H-bridge cell given in (3.15) is modified into

$$V_{cell,ij}^* = \left[\frac{\varepsilon_{ij}(V_i^* + V_{bal}^*)}{\varepsilon_{i1} + \dots + \varepsilon_{ij} + \dots + \varepsilon_{iN}} \right] (1 - G_2 \delta_{ij}) \quad (3.18)$$

where ε_{ij} is the fault signal for the j^{th} micro-pack in i^{th} phase and is given by 1 for being healthy and 0 for being faulty. This distributes the load among the other healthy units in the same phase in order to generate balanced three-phase output without neutral shift. This method is not efficient, and does not perform well in higher modulation indexes. However with neutral shift, the load can be shared with the healthy units in other phase legs of the CHB. It is well known that the addition of a zero sequence voltage to all three-phase voltage references modifies each phase voltage output without affecting the line-line voltage. For integration with the MD-PWM, the required voltage modifier that represents the required zero sequence voltage for the neutral shift is obtained from the fault signals and the V_i^* . Thus with the proposed method, the fault tolerant control along with the SOC balancing control is conveniently integrated with the PWM generation algorithm.

To achieve ride through capability with efficient load distribution, (3.18) is modified into,

$$V_{cell,ij}^* = \left[\frac{\varepsilon_{ij}(V_i^* + V_{bal}^* - V_f^*)}{\varepsilon_{i1} + \dots + \varepsilon_{ij} + \dots + \varepsilon_{iN}} \right] (1 - G_2 \delta_{ij}) \quad (3.19)$$

With the voltage modifier for ride through capability under faulty condition given by,

$$V_f^* = \frac{N - \max(\beta_A, \beta_B, \beta_C)}{3\beta - \max(\beta_A, \beta_B, \beta_C)} [f_A V_A^* + f_B V_B^* (1 - f_A) + f_C V_C^* (1 - f_A)(1 - f_B)] \quad (3.20)$$

where

$$f_i = \begin{cases} \left\lfloor \frac{(\Delta f_i + |\Delta f_i|)}{2 * \Delta f_{max}} \right\rfloor & ; \forall i \in (A, B, C) \text{ and } \Delta f_{max} > 0 \\ 0 & ; \forall i \in (A, B, C) \text{ and } \Delta f_{max} = 0 \end{cases}$$

$$\Delta f_{max} = \max(\Delta f_A, \Delta f_B, \Delta f_C)$$

$$\beta_i = \frac{I}{N} \sum_{j=1}^N \varepsilon_{ij}$$

$$\beta = \frac{I}{3} (\beta_A + \beta_B + \beta_C)$$

$$\Delta f_i = \beta - \beta_i$$

The V_f^* follows the voltage reference of the phase leg where the fault occurs and with most number of faulty H-bridge units. Magnitude of the V_f^* is determined based on the number of faulty and healthy H-bridge units. The V_f^* is added to the voltage references of all the three phase legs, which distributes the load demand from the phase leg having more number of faults to the other phase legs. For example, with $N = 2$ and fault occurs in B1 ($\varepsilon_{B1} = 0$), the magnitude of V_f^* is given by $0.67V_B^*$.

3.5 Experimental Investigations

Two laboratory setups, for RL load and induction motor load, have been built to demonstrate the operation of CHB converter with battery micro-packs for the efficacy of the proposed SOC balancing control and fault tolerant control using MD-PWM. Both the laboratory setup consists of a three-phase five level CHB with two H-bridge units per phase made of IGBTs, each supplied with a battery module. Rechargeable lithium ion battery module UBI 2590 [87] rated at 30-V 6.2-Ah / 15-V 12.4-Ah was employed for the first setup with RL load. Each such UBI module has two terminals of 15-V 6.2-Ah which are considered as a battery micro-pack in this demonstration. Every battery micro-pack in the second setup has 16 lithium-ion-phosphate cells of 3.3-V 60-Ah connected in series. A three-phase 1-HP induction motor has been employed as load.

The proposed control strategy and the gate signal generation algorithm using the MD-PWM has been implemented in dSPACE DS1104 control board. The dSPACE board has eight ADC channels, of which six were used for six battery micro-pack voltage measurements and the remaining two channels were used to measure two line currents of the three-phase load. The coulomb counting method has been implemented in dSPACE controller for the real-time SOC estimation and is given by,

$$SoC_{ij}(t) = SoC_{ini,ij} - \frac{1}{Q_{bat}} \int V_{bat,ij} i_{bat,ij}(t) dt \quad (3.21)$$

Q_{bat} is the capacity of the battery in Wh, V_{bat} and i_{bat} are the battery micro-pack voltage and current respectively. The battery current is derived from the line current of the motor, and the switching state ($S_{ij} \in [1, 0, -1]$) of the corresponding H-bridge using,

$$i_{bat,ij}(t) = S_{ij} i_i(t) \quad (3.22)$$

where $i_i(t)$ is i_A , i_B and i_C . The i_A and i_B are measured and i_C is derived using $i_C = -i_A - i_B$ as the load is balanced. In order to execute the coulomb counting method, the initial SOC ($SoC_{ini,ij}$) of each battery micro-pack is required. And it was entered into the dSPACE controller offline. The parameters of the laboratory setup are listed in Table 3.2. A snapshot of the experimental setup is shown in Fig. 3.4.

3.5.1 With RL Load

In this demonstration a constant reference has been given to the converter control system. This is because that the proposed control is based on the difference in energy extracted from the battery micro-packs over a certain amount of time. Hence if it works for a constant modulation index, it would work for the varying modulation index as in real time conditions. The required phase reference voltage was set to 17 V_{rms} 50 Hz and maintained same for all the operating conditions in order to make a fair comparison.

Under normal operating condition, all the battery micro-packs would be healthy and fully charged or charged to same SOC level. In such cases, equal power has to be maintained across all the H-bridge units. The traces of SOC and the average power of six battery micro-packs under normal operating condition are shown in Fig. 3.5. All battery micro-packs were charged to 80% SOC. Since all battery modules have equal initial SOC,

the power delivered by each module is same at ~ 17.7 W with the total input power of 106 W. As it can be observed the input power traces from Fig. 3.5(a), all the battery micro-packs have equal discharge rate of about 0.19C.

Fig. 3.4 Experimental setup

To emulate the condition of unequal SOC, each battery micro-pack was charged to different SOC levels ($SoC_{A1}=44\%$, $SoC_{A2}=45\%$, $SoC_{B1}=45\%$, $SoC_{B2}=46.5\%$, $SoC_{C1}=44\%$, and $SoC_{C2}=45.5\%$). Without SOC balancing control, the MD-PWM particularized as PS-PWM operates in a way to extract equal power from the battery micro-packs. This would make the micro-pack having lower SOC to reach the minimum cut-off value (35%) and the system should be stopped at $t = 1600$ s from further operation, even though some of the battery micro-packs have the SOC higher than the cut-off value, and is shown in Fig. 3.5(b). The phase voltages generated by the CHB, the load line voltage and currents of 50 Hz under normal operating condition are given in Fig. 3.5(c).

Table 3.2 Experimental Setup Parameters – RL Load

Parameter	Value
Cascade Number N	2
Battery micro-pack rating	15 V and 6.2 Ah
Load Resistor R	3.3Ω
Load Inductor L	6 mH
Switching period Tsw	1 ms
Inter-phase SOC balancing constant $G1$	0.32
Intra-phase SOC balancing constant $G2$	0.44

With the unequal initial SOC conditions, the CHB is operated with the proposed SOC balancing control. Since the SOC values are not equal, the proposed control modifies the $V_{cell,ij}$. Fig. 3.6 portrays the operation of the system with unequal initial SOC values and operated with the proposed SOC balancing control. There exists an imbalance between the average SOC values of three phases. The inter-phase SOC balancing control makes phase-B to deliver more power than phase-A and phase-C. The intra-phase SOC balancing control considers the individual micro-pack SOC in that particular phase. Therefore at $t=0$ s, the power delivered by the micro-packs are $P_{A1}=14.5$ W, $P_{A2}=20.5$ W, $P_{B1}=17.5$ W, $P_{B2}=24.5$ W, $P_{C1}=11$ W, and $P_{C2}=18$ W altogether delivering 106 W. Consequently, the SOC values of micro-packs are balanced in about 17 minutes. This time taken to balance the SOC values is based on the values of G_1 , G_2 , and the rate of discharge. The procedure to determine these constants has been explained in the previous section. It is clear from the SOC traces shown in Fig. 3.6(a) that the SOC values of the micro-packs are still at about 36% when $t = 1600$ s, which is above the cut-off limit. It shows that the proposed SOC balancing control prolongs the system operation even though when the battery micro-packs' SOC values are not equal. The phase voltages, line voltages and line currents of 50 Hz are given in Fig. 3.6(b). It is clear that the proposed control makes the CHB to generate different phase voltages, still maintaining the balanced load current.

To emulate the condition of a faulty micro-pack, the micro-pack of the unit B1 is assumed to be faulty and the corresponding fault signal was set to 0. In order to generate a balanced three-phase output, the proposed algorithm modifies the $V_{cell,ij}$ of the rest of the healthy units to meet the load demand as explained in the previous section. Without SOC balancing control, the fault tolerant control makes the system to operate in such a way that the power extracted from the phases having healthy micro-packs delivers more power ($P_{A1}=18$ W, $P_{A2}=18$ W, $P_{B1}=0$ W, $P_{B2}=24$ W, $P_{C1}=23$ W, and $P_{C2}=23$ W altogether delivering 106 W). It results in the SOC imbalance among the healthy battery micro-packs. It is evident from the power and SOC traces given in Fig. 3.7(a). The corresponding phase voltages, line voltages and line currents of 50 Hz are presented in Fig. 3.7(b). However with the SOC balancing control integrated to the fault tolerant control, this imbalance is resolved as shown in Fig. 3.8. With SOC balancing control in loop, unequal amount of power is extracted from each of the healthy micro-packs to maintain the SOC of each micro-pack close to their average SOC. The corresponding

phase voltages, line voltages and line currents of 50 Hz with integrated fault tolerant and SOC balancing control are given in Fig. 3.8(b).

Fig. 3.5 System operation under normal operating conditions without SOC balancing control

Fig. 3.6 System operation under unbalanced operating conditions with SOC balancing control

Fig. 3.7 System operation under fault condition without SOC balancing control

Fig. 3.8 System operation under fault condition with SOC balancing control

3.5.2 With Induction Motor Load

The lithium cells used for this experiment are of 60 Ah capacities. But to demonstrate the SOC balancing operation in a reasonable time with 1-HP motor load, the capacity of lithium cell is considered as 6 Ah in the SOC calculation. Open loop constant V/Hz control has been implemented for the speed control of the induction motor. The photo of the setup is shown in Fig. 3.9. The parameters of the laboratory setup are listed in Table 3.3.

Fig. 3.9 Experimental setup

Table 3.3 Experimental Setup Parameters – Motor Load

Parameter	Value
Cascade Number N	2
Battery micro-pack rating	52.8 V, and 60 Ah
Load – Induction Motor	1-HP
Switching period Tsw	1 ms
Inter-phase SOC balancing constant G1	0.4
Intra-phase SOC balancing constant G2	0.5

Under normal operating condition, all the H-bridge units with battery micro-packs would be healthy and fully charged or charged to an identical SOC level. In such cases, equal power has to be maintained across all the H-bridge units. The traces of reference variables and the system parameters under normal operating condition with a step change

in the speed reference are shown in Fig. 3.10. Fig. 3.10(a) shows the speed reference and the corresponding voltage reference generated by the constant V/Hz control. Fig. 3.10(b) shows the reference signals generated by the proposed control methods. Since all battery modules have equal initial SOC and no fault condition, the V_f^* and V_{bal}^* are zero. Therefore, all the H-bridge units have equal voltage references ($V_{cell,ij}^*$) as shown in Fig. 3.10(b). The voltages generated by the three-phase five-level inverter, the motor line current, and the input power from the six battery micro-packs are presented in Fig. 3.10(c). It can be seen that the power delivered by each micro-pack is identical.

To emulate the fault condition, the unit B1 in phase-B leg is assumed to be faulty and the corresponding fault signal was set to 0. The traces of reference variables and the system parameters under fault condition (without SOC balancing control) with a step change in the speed reference are shown in Fig. 3.11. Fig. 3.11(a) shows the speed reference and the corresponding voltage reference generated by the constant V/Hz control. Fig. 3.11(b) shows the reference signals generated by the proposed control methods. V_f^* is the voltage modifier that needs to be added to all the phase voltage references for this fault condition. Since fault has occurred in phase-B, the V_f^* follows V_b^* as explained in Section 3.4.3, and the reference voltage the faulty unit, $V_{cell,B1}^*$ is zero. The other H-bridge units' voltage references ($V_{cell,ij}^*$) are also shown in Fig. 3.11(b). The voltages generated by the three-phase five-level inverter, the motor line current, and the input power from the six battery micro-packs are presented in Fig. 3.11(c). It can be seen from the voltage wave forms that the B-phase voltage has only three levels as only one H-bridge is in operation. The balanced motor line current verifies the proposed control scheme for the fault condition. The power extracted from the micro-pack B2 is higher than the other battery micro-packs, which eventually results in an unbalanced SOC condition.

Fig. 3.10 System operation under normal operating condition - (a) Speed reference and voltage reference from motor control unit, (b) Voltage references generated by the proposed control schemes, (c) Voltages generated by the converter, line currents of the motor, input power from the battery micro-packs

Fig. 3.11 System operation under fault condition without SOC balancing control - (a) Speed reference and voltage reference from motor control unit, (b) Voltage references generated by the proposed control schemes, (c) Voltages generated by the converter, line currents of the motor, input power from the battery micro-packs

To emulate the condition of unequal SOC, each battery micro-pack was charged to different SOC levels ($SoC_{A1}=21.5\%$, $SoC_{A2}=23\%$, $SoC_{B1}=22\%$, $SoC_{B2}=23.5\%$, $SoC_{C1}=20\%$, and $SoC_{C2}=21\%$). The traces of reference variables and the system parameters under unbalanced SOC condition (with SOC balancing control) with a step change in the speed reference are shown in Fig. 3.12. Fig. 3.12(a) shows the speed reference and the corresponding voltage reference generated by the constant V/Hz control. Fig. 3.12(b) shows the reference signals generated by the proposed SOC balancing control method. Since all battery modules are healthy, the V_f^* is zero. There exists an imbalance between the average SOC of three phases. The inter-phase SOC balancing control makes phase-B to deliver more power than phase-A and phase-C. The V_{bal}^* generated for this unbalanced condition is shown. The intra-phase SOC balancing control considers the individual micro-pack SOC in that particular phase. Corresponding H-bridge voltage references ($V_{cell,ij}^*$) are presented. The voltages generated by the three-phase five-level inverter, the motor line current, and the input power from the six battery micro-packs are presented in Fig. 3.12(c). It can be seen that the power extracted from the micro-packs are different and is according to their SOC difference from the average SOC of the system. For instance, SoC_{B2} is higher compared to other SOC and hence the power extracted from B2 is higher compared to other micro-packs.

Fig. 3.12 System operation under unbalanced SOC condition with the SOC balancing control - (a) Speed reference and voltage reference from motor control unit, (b) Voltage references generated by the proposed control schemes, (c) Voltages generated by the converter, line currents of the motor, input power from the battery micro-packs

SOC traces of the battery micro-packs under different operating conditions without any SOC balancing control and with the proposed SOC balancing control are portrayed in Fig. 3.13(a) and Fig. 3.13(b) respectively. The left plot in Fig. 3.13(a) is for balanced SOC condition. It can be observed that all the battery micro-packs exhibits equal discharge rate under balanced SOC condition. The middle plot in Fig. 3.13(a) shows the SOC traces under unbalanced SOC condition with initial SOC's given by $SoC_{A1}=21.5\%$, $SoC_{A2}=23\%$, $SoC_{B1}=22\%$, $SoC_{B2}=23.5\%$, $SoC_{C1}=20\%$, and $SoC_{C2}=21\%$. Without SOC balancing control, the MD-PWM implemented as PS-PWM operates in a way to extract equal amount of power from the battery micro-packs. This would make the micro-pack having lower SOC to reach the minimum cut-off value (15%) quicker, and the system should be stopped at $t = 850$ s from further operation, even though some of the battery micro-packs have the SOC higher than the cut-off value. With the unequal initial SOC conditions, the CHB is operated with the proposed SOC balancing control. The corresponding SOC traces are given in the left plot of Fig. 3.13(b). There exists an imbalance between the average SOC's of three phases. The inter-phase SOC balancing control makes phase-B to deliver more power than phase-A and phase-C. The intra-phase SOC balancing control considers the individual micro-pack SOC in that particular phase. Consequently, the SOC's of micro-packs are balanced in about 22 minutes. With the SOC balancing control, the micro-pack having lower SOC reaches the minimum cut-off value (15%) at $t = 1400$ s. Hence the proposed SOC balancing control prolongs system operation by about 9.1 minutes. Such an operational improvement is very crucial for an electric vehicle, as it has been found that usually personal transportation electric cars have a driving range of about 4.5 hours with average speed in city traffic condition [84].

Without SOC balancing control, the fault tolerant control makes the system to operate in such a way that the power extracted from the phases having healthy micro-packs delivers comparatively more power. It results in the SOC imbalance among the healthy battery micro-packs and is evident from the SOC traces given in the right plot of Fig. 3.13(a). However, with the SOC balancing control integrated to the fault tolerant control, this imbalance is resolved as shown in the right plot of Fig. 3.13(b). With the SOC balancing control in loop, unequal amount of power is extracted from each of the healthy micro-packs to maintain the SOC of each micro-pack close to their average SOC. Without SOC balancing control, the SOC of B2 micro-pack reaches 57.5 % at about $t = 510$ s. With the SOC balancing control, the SOC of B2 micro-pack reaches 57.5 % at

about $t = 620$ s. This verifies that the proposed SOC balancing schemes aids in prolonging the system operation time.

Fig. 3.13 SOC traces of battery micro-packs under different operating conditions - (a) without SOC balancing control (left: balanced SOC condition, middle: unbalanced SOC condition, right: fault condition), (b) with the proposed SOC balancing control (left: unbalanced SOC condition, right: fault condition)

3.6 Simulation of the CHB-MBESS for rectifier operation

In this section, synchronous rectifier operation of the CHB is presented to explain the charging of battery micro-packs from a three-phase power supply. The CHB converter could be used for both inverter and rectifier operations. The operation explained in this section is also applicable for the regenerative braking in which the energy obtained from braking would be used to charge the battery micro-packs. The modulation method for the synchronous rectifier operation of the CHB is same as that of the inverter operation which has been explained earlier in this chapter. The amount of power transferred is controlled based on the decoupled current control method. The average SOC of all the micro-packs is used as reference to generate the required i_d . The control concept is illustrated in Fig. 3.14, where SoC is the average SOC of all the battery micro-packs in the CHB. Based on

the required SoC^* , the power will be extracted from the three-phase supply equipment. When the three-phase reference voltages are obtained from the decoupled current control module, the CHB is operated in similar way to the inverter operation using MD-PWM with integrated SOC balancing and fault tolerant control.

Fig. 3.14 Synchronous rectifier control method

The inverter operation of the CHB with SOC balancing control has been explained in Section 3.4. The synchronous rectifier operation of the CHB for battery micro-packs charging was simulated by connecting the CHB to a three-phase source of 220 V, 50 Hz. Simulation was done for CHB with three H-bridge cells in a phase leg ($N = 3$). The line voltage output of the CHB and the line current when the micro-packs are charged are shown in Fig. 3.15. The fault tolerant and SOC balancing methods are also applicable to the rectifier operation.

The MD-PWM makes the CHB to operate as multilevel inverter which is evident in the voltage trace in Fig. 3.15. The second trace in Fig. 3.15 shows the three-phase load current. To verify the ride through capability of the CHB under fault condition, one of the micro-pack is considered faulty in phase 'b'. The results obtained are given in Fig. 3.16. The fault condition is induced at $t = 0.5$ s. As depicted, the line current shows no significant change even though one H-bridge is bypassed.

Fig. 3.15 CHB Phase voltages and currents during rectifier operation under normal condition

Fig. 3.16 CHB Phase voltages and currents during rectifier operation under fault condition

3.7 Simulation of the CHB-MBESS for regeneration

The concept of bi-directional power flow is essential in the battery operation and control for EV applications. This section discusses the regenerative operation of the CHB with MD-PWM. During the operation of EV, the mode of operation of the electric machine is as motor during driving or powering and as generator during when the kinetic energy of the EV is reduced and converted to the electrical energy. This mode of operation is called regenerative mode. During regenerative mode, the energy is extracted from the machine to charge the batteries through the power converter. These modes of operation will frequently and quickly alter each other during the EV operation.

Fig. 3.17 Simulation of PMSM with CHB

Permanent Magnet Synchronous Machine driven by the CHB with the modular battery micro-packs, as shown in Fig. 3.17, is modelled in Simulink to demonstrate the powering and regenerative mode of operation. The commonly used Field Oriented Control (FOC) is used in this simulation for the control of PMSM. Electric braking in EV is realized in different ways including DC braking, plug braking and regenerative braking [88-90]. Plug braking and regenerative braking can be realized using the FOC method. When the braking torque is less, negative quadrature axis currents results in regeneration and when the braking torque is higher, the braking eventually results in plug braking.

A 40 kW 4 pole pair PMSM is modelled in Simulink with the CHB-MBESS. The results from simulation of CHB-MBESS operation to demonstrate the regeneration capability is shown in Fig. 3.18. The speed of the machine is controlled at 400 rpm with load torque of 700 Nm, as it can be observed from the top two plots. The CHB converter is operated in the powering mode driving the converter. Corresponding line voltages generated by the CHB converter as inverter and the line currents drawn by the machine are presented in bottom two plots. At $t=2$ s, the speed command is reduced, which results in braking torque being generated and the velocity of the machine starts reducing. The CHB converter enters into regenerative mode and operated as rectifier, as it can be observed from the line voltage and line current plots.

Fig. 3.18 Simulation of CHB-MBESS operation to demonstrate regeneration capability

3.8 Summary

- ❖ MD-PWM, a generalized modulation technique for the CHB converter has been reviewed.
- ❖ A control strategy for the SOC balancing between the battery micro-packs and realization of fault ride through operation using MD-PWM has been explained.
- ❖ The advantage of the proposed methods is that it can be conveniently integrated with the MD-PWM algorithm.
- ❖ The results obtained from the laboratory setup of a five-level three-phase CHB with six battery modules of 15-V 6.2-Ah for RL load have been presented.

- ❖ The results obtained from the laboratory setup of a five-level three-phase CHB driving an induction motor with six battery modules of 52.8-V 60-Ah have been presented.
- ❖ The results have verified that the fault tolerant and SOC balancing operation can be achieved with CHB using MD-PWM. This shows that the proposed micro-pack system with CHB would improve the EV performance in the event of fault, which is not possible with the conventional single battery pack and two-level converter designs.
- ❖ The operation of CHB as rectifier for battery micro-packs charging and for regeneration capability has also been demonstrated using simulation results.

Chapter 4 Double Star Chopper Cell Converter based Modularized Battery Energy Storage System (DSCC-MBESS)

4.1 Introduction

Fig. 4.1 DSCC for MBESS

Conventional battery electric vehicles (EV) have a single battery pack with a two level inverter for motor drive. The battery pack is constructed by interconnecting certain number of electrochemical cells in series (and/or parallel) to achieve required energy and power capacity. The required power and energy capacity depends on the vehicle size, maximum speed and driving range per charging etc. Any fault in the battery pack affects

the availability of the vehicle. With the motive of increasing the availability of an EV, a modular battery pack or micro-pack system is proposed and investigated in this work. The reliability and availability of the vehicle is increased when the power in each battery micro-packs are controlled independently and when the system possesses ride through capability under fault conditions. This is achieved using modular power converters. One such system employing Cascaded H-bridge converter has been discussed in Chapter 3. Another type of modular power converter named Double star chopper cell (DSCC) multilevel converter shown in Fig. 4.1, is investigated in this chapter. DSCC has been investigated by various researchers for medium voltage applications. However in this work, DSCC for electrical vehicle application is investigated. The difference between the conventional DSCC based MMC and the proposed system is that each sub-module in DSCC is fed with battery micro-pack instead of capacitors and single larger DC supply.

A design concept for modularizing the battery pack into micro-packs for DSCC is described in Section 4.2. The DSCC converter can be operated either as an inverter to drive the EV motor or as a synchronous rectifier connected to external three-phase power supply equipment to charge the battery micro-packs. The multi-dimensional pulse width modulation (MD-PWM) for DSCC is introduced in Section 4.3. The methods of operation for inverter using MD-PWM with integrated inter-module SOC balancing and fault tolerant control are described in Section 0. The method of operation as synchronous rectifier using MD-PWM is discussed in Section 4.5. The proposed system operation as inverter and synchronous rectifier are verified through simulations and the results are presented in Section 4.6.

4.2 Battery Micro-pack Design Concept

The criteria for designing a modularized battery micro-packs using lithium ion cell for the DSCC converter is described in this section. The basic design parameters of the EV battery pack are the maximum power required (P_{max}^*) for maximum acceleration with the given vehicle weight and the maximum energy capacity (E_{max}) of the battery pack to attain certain driving range. The voltage of the battery pack should be in accordance with the rated line-line voltage of the drive motor. In order to generate a line voltage with peak value of V_p with DSCC, the required micro-pack voltage is given by,

$$V_{bat} \geq \frac{2V_p}{\sqrt{3}n} \quad (4.1)$$

where n is the number of modules (or chopper cells with battery micro-packs) in a given arm of the DSCC. The rated power of the battery pack should be greater than the power required for maximum acceleration P_{acc} of the vehicle. To achieve this maximum power with DSCC and micro-packs, the rated power of each micro-pack should be

$$P_{bat}^{max} \geq \frac{P_{acc}}{6n} \quad (4.2)$$

With the V_{bat} and P_{bat}^{max} calculated using (4.1) and (4.2), the Ah of the lithium ion cell and the number of cells for a battery micro-pack can be manipulated. In this work, the chosen values for investigation are $V_p = 400$ V and $P_{acc} = 100$ kW. Therefore the required V_{bat} should be ≥ 154 V and the required P_{bat} should be ≥ 5.6 kW for $n = 3$. From V_{bat} and P_{bat} , the rated current of the micro-pack, I_{bat} , should be ≥ 36.4 A. Any lithium ion cell with nominal voltage of 3.7 V and the nominal capacity of 10 Ah with maximum discharging capacity of at least 5 *C-rate* can be chosen for this application. With such lithium cell chosen, the micro-pack can be constructed with 42 cells in series. The total capacity of such system with DSCC and 18 micro-packs would be 28 kWh with each micro-pack being 1.55 kWh (with nominal voltage of 155.4 V). The total number of lithium cells required to build such a system is 756.

Table 4.1 Parameters of the battery pack

<i>Parameters</i>	<i>DSCC</i>	<i>Conventional (Nissan Leaf)</i>
Number of Li-ion cells	756 (18 Micro-packs with 42 cells each)	192
Number of power IGBTs	36	6
Nominal Capacity of each cell	10 Ah	33 Ah
Total Energy storage capacity	28 kWh	24 kWh
Nominal Battery pack Voltage	155.4 V (for each micro-pack)	360 V

In-order to compare with the conventional battery pack design, a commercially available electric car is taken into consideration. Nissan Leaf battery pack (24 kWh and 360 V) has 192 Lithium ion cells [84]. Both the systems would have similar output performance. The proposed DSCC based system requires approximately four times more number of cells but with one third of the nominal capacity of the conventional system.

4.3 DSCC and Multi-dimensional PWM

The operation of DSCC as inverter to obtain multilevel three-phase output depends on the manipulation of the reference phase voltage. The references for each arm is obtained in such a way that positive arm current and negative arm current are out of phase so that they cancel out the flux in the coupling inductor in each leg. The positive and negative arm references in a phase leg are given by,

$$v_{arm (+)}^* = \frac{\sum_{i=1}^n V_{bat,i}}{2} - v_p^* \quad (4.3)$$

$$v_{arm (-)}^* = \frac{\sum_{i=1}^n V_{bat,i}}{2} + v_p^* \quad (4.4)$$

where v_p^* is the required phase voltage and $V_{bat,i}$ is the i^{th} battery micro-pack voltage in the corresponding arm.

Any pulse width modulation schemes like phase shifted (PS-PWM), level shifted (LS-PWM), or space vector modulation can be applied to DSCC like any other multilevel converter. Multidimensional modulation (MD-PWM) proposed in [47] is a generalized modulation approach for a Cascaded H-Bridge (CHB) multilevel converter topology. It can be specialized to any of the commonly used modulation techniques like PS-PWM or LS-PWM. For CHB, the output of each H-bridge in a phase leg can be at any of the three voltage levels $(+V_{bat}, 0, -V_{bat})$ depending on the switching state, where V_{bat} is the input DC voltage. The MD-PWM technique is based on the geometric depiction of the output voltage generated by each H-bridge inverter in co-ordinate axes. The control contour is a straight line for two H-bridges in series and the control region is a cube for having three H-bridges cascaded. For cascade number more than three, the control region cannot be denoted graphically but the concept can be extended [47] .

Fig. 4.2 Control region for three cascaded chopper cells ($n = 3$)

MD-PWM strategy for a DSCC is briefed in this section. An arm in the DSCC has a stack of chopper cells connected in series. Multidimensional modulation can be applied for an arm in the DSCC as the stack voltage of an arm averaged over a period is the sum of average voltage outputs of the chopper cells in it and is given by,

$$V_{arm} = \sum_{i=1}^n V_{cell,i} \quad (4.5)$$

where the average voltage output of each chopper cell, $V_{cell,i}$ ($\forall i=1,2,3\dots n$) can have any value from 0 to $V_{bat,i}$ ($\forall i=1,2,3\dots n$) in a period with $V_{bat,i}$ being the battery micro-pack voltage. Therefore solution to (4.5) for any desired value of V_{arm} is a point on a plane orthogonal to $[V_{bat,1} ; V_{bat,2} ; V_{bat,3}]$ vector bounded by the cubical control region. An example of a solution set is shown as a dashed triangular plane inside the cube in Fig. 4.2. Intrinsically, there exists infinite number of solutions to any desired value of V_{arm} . This existence of wide range of solution set is a specific advantage of MD-PWM since it provides an extra degree of freedom to choose a $V_{cell,i}^*$ ($\forall i=1,2,3\dots n$) depending on a design criterion. The only difference between MD-PWM for DSCC and CHB is that the $V_{cell,i}$ in DSCC can be only positive as it has chopper cell.

Once the $V_{cell,i}^*$ ($\forall i=1,2,3\dots n$) for the desired V_{arm} in a switching period (T_{sw}) is obtained, the next step is to determine the switching time and the switching sequence. The $V_{cell,i}^*$ ($\forall i=1,2,3\dots n$), switching sequence and switching time determination is based on a particular design criteria such as to particularize PS-PWM or LS-PWM. Each chopper cell in an arm can take two switching states $S \in [0, 1]$ which correspondingly generates the output voltage of $[0, V_{bat}]$. The switching time is calculated using the equations (4.6) and (4.7) where t_{on} is for state $S = 1$ and t_{off} is for state $S = 0$.

$$t_{on,i} = \frac{|V_{cell,i}^*|}{V_{bat,i}} \times T_{sw} ; \forall i=1,2,3\dots n \quad (4.6)$$

$$t_{off,i} = T_{sw} - t_{on,i} ; \forall i=1,2,3\dots n \quad (4.7)$$

4.4 Micro-pack SOC balancing and Fault Tolerant Control using MD-PWM

SOC variation among the cells in a battery pack affects the performance of the battery pack. Thus various cell balancing strategies have been proposed and investigated by various researchers. With modules or micro-packs with independent power flow control, the balancing of SOC among the micro-packs becomes mandatory. Thus the methods for balancing the SOC among the micro-packs in a phase leg and also across three phase legs are introduced. The proposed balancing method requires no extra component. Instead, the strategy is integrated into the modulation technique.

Each phase leg in DSCC has $2n$ (n in each arm) chopper cells with battery micro-packs. Let $SoC_{i,P}$ be the average SOC of n battery micro-packs' SOC's ($SoC_{i1,P}$ to $SoC_{in,P}$) in the positive arm of phase i , with $SoC_{ij,P}$ being the SOC of j^{th} micro-pack in the positive arm of phase i for all $i \in (A,B,C)$. Similarly, let $SoC_{i,N}$ be the average SOC of n battery micro-packs' SOC's ($SoC_{i1,N}$ to $SoC_{in,N}$) in the negative arm of phase i , with $SoC_{ij,N}$ being the SOC of j^{th} micro-pack in the negative arm of phase i for all $i \in (A,B,C)$. Let SoC_i be the average SOC of $2n$ battery micro-packs' SOC's in phase i and is given by,

$$SoC_i = \frac{1}{2} (SoC_{i,P} + SoC_{i,N}) \quad (4.8)$$

where

$$SoC_{i,P} = \frac{1}{n} \sum_{j=1}^n SoC_{ij,P} \quad (4.9)$$

$$SoC_{i,N} = \frac{1}{n} \sum_{j=1}^n SoC_{ij,N} \quad (4.10)$$

And the overall SOC of the energy storage system, SoC is defined by the average of SoC_A , SoC_B and SoC_C , and is given by,

$$SoC = \frac{1}{3} (SoC_A + SoC_B + SoC_C) \quad (4.11)$$

The deviation of SoC_A , SoC_B and SoC_C from SoC is given by,

$$\Delta_i = SoC - SoC_i \quad (4.12)$$

The deviation of j^{th} micro-pack SOC in i^{th} phase (SoC_{ij}) from its average SOC (SoC_i) is given by,

$$\delta_{ij} = SoC_i - SoC_{ij} \quad (4.13)$$

4.4.1 Inter-phase SOC balancing

If all the phase legs have equal average SOC (i.e., $\Delta_A = \Delta_B = \Delta_C = 0$), then power extracted during motoring operation and the power delivered during charging operation should be distributed equally among three phases. But when the phase legs have difference in their average SOC values, the power extracted from or delivered to each phase leg should be varied according to their average SOC to achieve a balanced operation. This is achieved by adding or subtracting zero sequence voltage of fundamental frequency to the phase voltage references. The zero sequence voltage modifies the phase voltage generated, and hence affects the power extracted from each phase without affecting the output line-line voltage.

The MD-PWD requires the reference voltage $V_{cell,ij,P}^*$ and $V_{cell,ij,N}^*$ for every j^{th} chopper cell unit in the positive arm and the negative arm respectively in a i^{th} phase to be

manipulated in every switching period. As explained in Section 4.3, a particular design criterion can be used to manipulate $V_{cell,ij,P}^*$ and $V_{cell,ij,N}^*$. In order to extract equal amount of power from each battery micro-packs, $V_{cell,ij}^*$ is calculated using,

$$V_{cell,ij,P}^* = \left(\frac{V_{arm(+)}^*}{n} \right) \quad (4.14)$$

$$V_{cell,ij,N}^* = \left(\frac{V_{arm(-)}^*}{n} \right) \quad (4.15)$$

where $V_{arm(+)}^*$ and $V_{arm(-)}^*$ are the desired arm voltages over a switching period defined in (4.3) and (4.4) respectively. To include the inter-phase SOC balancing control in MD-PWM, the reference for each chopper cell unit in (4.14) and (4.15) are modified into,

$$V_{cell,ij,P}^* = \left(\frac{V_{arm(+)}^* - V_{bal}^*}{n} \right) \quad (4.16)$$

$$V_{cell,ij,N}^* = \left(\frac{V_{arm(-)}^* - V_{bal}^*}{n} \right) \quad (4.17)$$

where V_{bal}^* is the voltage modifier that represents the required zero sequence voltage at every switching period. It is obtained from the reference phase voltage amplitude V_i^* and the SOC deviation. The required voltage modifier to achieve the inter-phase SOC balancing is given by,

$$V_{bal}^* = -G_I [f_A \Delta_A V_A^* + f_B \Delta_B V_B^* (1-f_A) + f_C \Delta_C V_C^* (1-f_A)(1-f_B)] \quad (4.18)$$

where

$$f_i = \begin{cases} \left\lfloor \frac{(\Delta_i + |\Delta_i|)}{2 * \Delta_{max}} \right\rfloor & ; \forall i \in (A, B, C) \text{ and } |\Delta_{max}| > 0 \\ 0 & ; \forall i \in (A, B, C) \text{ and } \Delta_{max} = 0 \end{cases}$$

$$\Delta_{max} = \max(\Delta_A, \Delta_B, \Delta_C)$$

The V_{bal}^* follows the voltage reference of the phase leg having higher SOC deviation from the average SOC of three phase legs. Magnitude of the V_{bal}^* is determined based on the SOC deviation and the proportionality constant G_I . The proportionality constant G_I is the percentage of voltage reference that is required to be subtracted from

each of the reference phase amplitudes. The value for G_I is obtained from the input power and SOC relation of each chopper cell unit given in Fig. 4.3(a). G_I is the representation of difference between the amounts of power delivered by each phase legs in the DSCC converter, which needs to be modified for certain time period to nullify their SOC differences. In in Fig. 4.3(a), $P_{L,i}$ is the loss and P_i is the active power of i^{th} phase.

Fig. 4.3 Control block diagram for SOC balancing for DSCC BESS - (a) Inter-phase SOC balancing (b) Intra-phase SOC balancing

The closed loop transfer function is given by,

$$\frac{\Delta_i(s)}{P_L(s)} = \frac{\frac{1}{G_I P_i}}{(1 + s \frac{2nQ}{100 G_I P_i})} \quad (4.19)$$

where Q is the maximum capacity (in Joules) of each battery micro-pack. The value for G_I is manipulated from the time constant of the transfer function given by,

$$T = \frac{2nQ}{100 G_I P_i} \quad (4.20)$$

4.4.2 Intra-phase SOC balancing

If all the micro-packs in an arm have equal SOC (i.e., $\delta_{i1} = \delta_{i2} = \dots = \delta_{iN} = 0$), then power extracted during motoring operation and the power delivered during charging operation should be distributed equally among all the micro-packs in an arm. This section explains a novel method that can be integrated with the PWM generation algorithm to achieve SOC balanced operation among the micro-packs within the arm. In order to vary the amount of power distribution with respect to SOC of individual battery micro-pack, (4.16) and (4.17) are modified to,

$$V_{cell,ij}^* = \left[\frac{(V_{arm}^* + V_{bal}^*)}{N} \right] (1 - G_2 \delta_{ij}) \quad (4.21)$$

where V_{arm}^* is the desired average arm voltage over a switching period of T_{sw} . V_{arm}^* could be $V_{arm(+)}^*$ for the positive arm or $V_{arm(-)}^*$ for the negative arm

The parameter δ_{ij} modifies the $V_{cell,ij}^*$ in every switching period to vary the amount of power distributed among the chopper cell units cascaded in series. The parameter δ_{ij} is determined from the deviation of the j^{th} battery micro-pack's SOC (SoC_{ij}) from the average SOC of all battery micro-packs in that particular arm. The value of G_2 is obtained in similar way as G_1 in the inter-phase balancing, based on the percentage of power that needs to be varied to nullify the SOC differences among the micro-packs in a phase leg. The control block diagram for intra-phase SOC balancing is shown in Fig. 4.3(b). From the closed loop transfer function, the proportionality gain G_2 is given by,

$$G_2 = \frac{Q}{100TP_{ij}} \quad (4.22)$$

As the parameter δ_{ij} depends on the deviation of SoC_{ij} from SoC_i , δ_{ij} can be positive or negative i.e., δ_{ij} is positive when SoC_{ij} is smaller than SoC_i and δ_{ij} is negative when SoC_{ij} is greater than SoC_i . The minimum value that δ_{ij} can have is -1. But when $\delta_{ij} = -1$, then it is similar to faulty condition and that corresponding chopper cell is clamped to zero voltage. This might cause unbalance between the positive and negative voltages of that phase and ultimately leads to unbalance in the three-phase output of the DSCC. Hence it requires additional control technique, and is explained in the following section.

The maximum value of δ_{ij} depends on the modulation index and is determined based on the other outer control loop. The upper range of δ_{ij} is determined so that the corresponding chopper cell output voltage $V_{cell,ij} \leq V_{bat,ij}$. Hence the range of δ_{ij} can be defined as in (4.23).

$$-1 \leq G_2 \delta_{ij} \leq \frac{n \times V_{bat,ij} - |V_{arm}^*|}{|V_{arm}^*|} \quad (4.23)$$

4.4.3 Ride through operation under faulty condition

The motivation for investigating modular converters with micro-packs is to achieve the ride through capability under fault conditions, which is not feasible with the conventional single battery pack and the two-level inverter system. Various fault tolerant operation methods was proposed in literature for multilevel and modular converters [49, 53, 54]. The effective method of operation would be to distribute the load equally among the healthy micro-packs. The motivation in this study is neither on type or place of the fault occurrence nor on the fault detection methods. Instead, a fault tolerant operation realization using MD-PWM is studied in detail. The following method is derived assuming that the fault signals are available from any of the real time state of health monitoring systems like the ones proposed in [11, 86]. The DSCC converter requires balanced DC voltage between the positive and negative bus. Hence with one micro-pack being faulty in any of the arm, one module should be bypassed in rest of the arms to achieve balanced output. A method to achieve this using MD-PWM is discussed in this section. The disadvantage in DSCC is that the ride through operation cannot be achieved when fault occurs at the top switch of any of the chopper cell. Otherwise this method is applicable to the occurrence of either fault in battery micro-pack or in the bottom switch of any of the chopper cell. Furthermore, it is assumed that bypassing a chopper cell could be possible using the zero switching state ($S = 0$). In-order to achieve ride through capability under faulty condition, the reference voltage for each H-bridge cell given in (4.21) is modified into

$$V_{cell,ij}^* = \left[\frac{\varepsilon_{ij}(V_{arm}^* + V_{bal}^*)}{\varepsilon_{i1} + \dots + \varepsilon_{ij} + \dots + \varepsilon_{iN}} \right] (1 - G_2 \delta_{ij}) \quad (4.24)$$

And the V_{arm}^* is modified into,

$$V_{arm (+)}^* = \frac{\sum_{i=1}^n \varepsilon_{ij,P} V_{bat,ij,P}}{2} - v_p^* \quad (4.25)$$

$$V_{arm (-)}^* = \frac{\sum_{i=1}^n \varepsilon_{ij,N} V_{bat,ij,N}}{2} + v_p^* \quad (4.26)$$

where ε_{ij} is the fault signal for the j^{th} micro-pack in i^{th} phase ($\varepsilon_{ij,P}$ is for the micro-pack in the positive arm and $\varepsilon_{ij,N}$ is for the micro-pack in the negative arm) and is given by 1 for being healthy and 0 for being faulty. This distributes the load among the other healthy units in the same phase in order to generate balanced three-phase output without neutral shift. This method is not efficient, and does not perform well in higher modulation indexes.

4.5 Rectifier operation of the DSCC for battery charging

The battery micro-packs in the vehicle require recharging when the stored energy is utilized. In this section, the synchronous rectifier operation of the DSCC is explained to charge the battery micro-packs from a three-phase power supply. Adopting this method eliminates the need for either the off-board charger or the on-board charger, as the same DSCC converter is used for both inverter and rectifier as well. The three-phase power supply equipment is though needed for charging the vehicle. The operation explained in this section is also applicable for the regenerative braking in which the energy obtained from braking would be used to charge the battery micro-packs. The modulation method for the synchronous rectifier operation of the DSCC is same as that of the inverter operation which has been explained in the previous section. The amount power transferred is controlled based on the decoupled current control method. The average SOC of all the micro-packs is used as reference to generate the required i_d . The control concept is illustrated in Fig. 4.4, where SoC is the average SOC of all the battery micro-packs in the DSCC. Based on the required SoC^* , the power will be extracted from the three-phase supply equipment. When the three-phase reference voltages are obtained from

the decoupled current control module, the DSCC is operated in similar way to the inverter operation using MD-PWM with integrated SOC balancing and fault tolerant control.

Fig. 4.4 Synchronous rectifier control method

4.6 Simulation Results

The DSCC based converter with $n=3$ was simulated in the Simulink and PLECS with three-phase RL load to demonstrate the inverter operation. The parameters of the system used in simulation are given in Table 4.2. A typical equivalent circuit model of the battery cell is given in Fig. 4.5.

Fig. 4.5 Electric circuit model of a Lithium ion cell

Lithium ion battery cell can be modelled as equivalent electric circuit model or mathematical model. Though the mathematical model is more accurate in finding the internal characteristic of the battery, it requires heavy computation. To use in a system level simulation like an electric drive system simulation, the equivalent electric circuit model shown in Fig. 4.5 is sufficient as it is simple and approximates the battery behaviour very well. The parameters of the equivalent circuit model can be found by experimentation [91] [92], and in this simulation, the battery cell is modelled as open-circuit voltage with internal resistance of 0.0745 ohms. The open circuit voltage V_{oc} is calculated using (4.27) and (4.28). SOC is calculated using the very well-known Coulomb counting method as given in (4.29).

$$V_{oc,d} = E_0 - k \frac{Q_o}{Q_o - Q} i^* - k \frac{Q_o}{Q_o - Q} Q \quad (4.27)$$

$$V_{oc,c} = E_0 - k \frac{Q_o}{0.1Q_o - Q} (i^* + Q) \quad (4.28)$$

$$SoC = SoC_{initial} - \frac{1}{Q_o} \int_0^t V_{bat} i(t) dt \quad (4.29)$$

where $V_{oc,d}$ is the open circuit voltage (V) when battery is discharging, $V_{oc,c}$ is the open circuit voltage (V) when battery is charging, E_0 is the nominal voltage (V), k is the polarization constant (Ah^{-1}) or polarization resistance (Ohms), i^* is the low frequency current dynamics (A), Q is the extracted capacity (Ah), Q_o is the maximum battery capacity (Ah)..

Table 4.2 Parameters of the proposed system used in simulation

<i>Parameters</i>	<i>Value</i>
Load R_L	10 Ω
Load L_L	6 μH
Coupling Inductor L_c	1 μH
Switching Period T_{sw}	0.5 ms
Sample Time T_s	1 μs

The inverter operation of the DSCC with SOC balancing control has been explained in Section 0. The reference for output phase voltage was set at 220 V, 50 Hz in the simulation. The MD-PWM makes the DSCC to operate as multilevel inverter which is evident in the voltage trace v_{ab} in Fig. 4.6. The second trace in Fig. 4.6 shows the three-phase load current. There is a modulation index change induced at $t = 0.5$ s.

Fig. 4.6 Load line voltage and current traces under normal operation

To verify the ride through capability of the DSCC under fault condition, one of the micro-pack is considered faulty in the positive arm in phase ‘a’. Therefore corresponding ε is zero, and to obtain balanced output, one module in each arm is bypassed by making the corresponding ε to be zero. The results obtained are given in Fig. 4.7. The fault condition is induced at $t = 0.5$ s. As depicted, the line voltage and current shows no significant change even though one module in each arm is bypassed.

Fig. 4.7 Load line voltage and current traces with one faulty module per arm from $t=0.5$ s

Fig. 4.8 depicts the SOC of 18 battery micro-packs of the DSCC during inverter operation. To verify the SOC balancing algorithm, the initial SOC values of the micro-packs was set between 79.5% and 80% randomly. The micro-pack with higher average SOC is discharged more than the lower average SOC value. Hence the SOC values get balanced eventually which is evident in Fig. 4.8.

Fig. 4.8 SOC of 18 Micro-packs

Fig. 4.9 Line voltage and current traces during rectifier operation

The battery micro-packs are required to be recharged when the average SOC value reaches the specified minimum value. The synchronous rectifier operation of the DSCC for battery micro-packs charging has been described in Section 4.5. This operation was simulated by connecting the DSCC to a three-phase source of 220 V, 50 Hz. The line voltage output of the DSCC and the line current when the micro-packs are charged at about 5 *C-rate* is shown in Fig. 4.9. The fault tolerant and SOC balancing methods are also applicable to the rectifier operation.

4.7 Simulation of the DSCC-MBESS for regeneration

As discussed in Section 3.7 for the CHB, the concept of bi-directional power flow is essential for DSCC in the battery operation and control for EV applications. This section discusses the regenerative operation of the DSCC with MD-PWM. Permanent Magnet Synchronous Machine (PMSM) PMSM driven by the DSCC with the modular battery micro-packs as shown in Fig. 4.10, is modelled in Simulink to demonstrate the powering and regenerative mode of operation. The commonly used Field Oriented Control (FOC) is used in this simulation for the control of PMSM.

Fig. 4.10 Simulation of PMSM with DSCC-MBESS

A 40 kW 4 pole pair PMSM is modelled. The results from simulation of DSCC-MBESS operation to demonstrate the regeneration capability is shown in Fig. 4.11. The speed of the machine is controlled at 400 rpm with load torque of 700 Nm, as it can be observed from the top two plots. The DSCC converter is operated in the powering mode driving the converter. Corresponding line voltages generated by the DSCC converter as inverter and the line currents drawn by the machine are presented in bottom two plots. At $t=2$ s, the speed command is reduced, which results in braking torque being generated and the velocity of the machine starts reducing. The DSCC converter enters into regenerative mode and operated as rectifier, as it can be observed from the line voltage and line current plots.

Fig. 4.11 Simulation of DSCC-MBESS operation to demonstrate regeneration capability

4.8 Comparison with Cascaded H-bridge converter

4.8.1 Advantages

- More number of battery micro-packs with independent power control per phase, and hence the fault tolerant level for same number of power electronic switches is higher for DSCC when compared to CHB.

-
- The battery micro-packs for DSCC requires battery cells of lower current capacity (Ah) to achieve the same converter power capacity and output voltage levels when compared to CHB.
 - SOC balancing could be achieved between the upper arm group of micro-packs and the lower arm group provides an extra degree of freedom in the SOC balancing strategy, which is not possible in the CHB.

4.8.2 Disadvantages

- The power loss in the coupled inductor per phase is considered an inherent disadvantage of the DSCC based converter systems.
- Increased number of battery cells though with lesser current capacity makes battery management and thermal management more complex compared to CHB based system.
- Another inherent disadvantage in DSCC is that the ride through operation cannot be achieved when fault occurs at the top switch of any of the chopper cell.
- Any fault in one of the micro-packs in the upper arm requires one micro-pack to be bypassed in the bottom arm, and thus reducing the overall capacity of the system. It is considered as a disadvantage when compared to CHB.

Considering the disadvantages of DSCC, it can be concluded that CHB is the preferred choice for EV application with modular battery micro-packs.

4.9 Summary

- ❖ A modularized battery system with DSCC based modular multilevel converter has been proposed for a battery operated electric vehicle (EV).
- ❖ A conceptual design of lithium ion based battery micro-packs for DSCC has been described.

- ❖ Multidimensional pulse width modulation (MD-PWM) with integrated inter-module SOC balancing and fault tolerant control has been proposed and explained.
- ❖ The methods of operation of the DSCC as an inverter and as a synchronous rectifier using MD-PWM with integrated inter-module SOC balancing and fault tolerant control have been discussed.
- ❖ The performances of the DSCC for SOC balancing and fault tolerant operation have been verified through simulations and the results have been presented. The results shows that the fault ride through capability of the DSCC based converter with battery micro-packs would increase the availability of the EV when compared with the conventional battery pack system.
- ❖ Operation of the DSCC as rectifier for the regeneration capability has also been demonstrated using simulation results.

Chapter 5 Inductive Power Transfer based Modularized Battery Energy Storage System (IPT-MBESS)

5.1 Introduction

Wide usage of EVs is largely hampered by limited driving range, lengthy battery recharging time and unavailability of convenient charging infrastructure. Two novel modularized energy storage systems with inductive interface are discussed in this chapter for convenient battery swapping that aids in overcoming these limitations. Each modularized energy storage system consists of a plurality of battery modules which are wirelessly coupled to the EV through inductive power transfer (IPT) technology. The proposed systems are described in detail, and models are presented to analyse their steady-state behaviours. A design guideline for a 24 kWh 80 kW battery micro-pack system is discussed. Performances of the proposed topologies are investigated using simulations. To demonstrate the applicability, prototype systems of 1.5 kW are implemented and tested under various operating conditions. Results convincingly indicate that the proposed systems improve the vehicle's availability under fault condition. And with inductive interface the proposed systems are attractive for EVs to be convenient alternatives to plug-in EV charging.

The modularized battery pack with independent power control is termed as modules or micro-packs in this work. The number of modules or the micro-packs (N) depends on the vehicle design requirements. The conceptual representation of the proposed modular EV energy storage systems with inductive coupling is illustrated in Fig. 5.1 and Fig. 5.2. A typical Inductive Power Transfer (IPT) system has two coils: the primary or the sending coil and the secondary or the receiving coil. The proposed system would have hollow casings containing the IPT secondary side that are fixed to the EV. The battery micro-pack with the IPT primary side is designed to be in a container that fits into the hollow casing attached to the EV. Type 1 of the proposed system depicted in Fig. 5.1 has one casing for each module. The output of each casing is cascaded to form the DC link of the three-phase motor drive inverter. Type 2 system has one casing that accommodates all the modules as depicted in Fig. 5.2. Hence Type 2 system employ less number of power switches when compared with Type 1. The output capacitor of the secondary forms the DC link of the three-phase inverter.

Fig. 5.1 Conceptual representation of Type 1 proposed system

Fig. 5.2 Conceptual representation of Type 2 proposed system

The proposed systems have no physical electrical connection and hence are highly suitable for quick swapping without any possible electrical hazards. The proposed systems are also highly suitable for smart charging where the EV batteries are charged with wide range of input power to meet the power system constraints based on certain optimization algorithms [93]. The external charging of the proposed modular battery packs is not addressed in this work. The operation of the proposed systems for the motor drive operation when the battery micro-packs are discharging is investigated in this chapter.

Type 1 and Type 2 of proposed modularized battery systems with inductive coupling are detailed in Sections 5.2 and 5.3 respectively. A design guideline and simulation results for a 24 kWh 80 kW battery micro-pack system with inductive coupling are presented. A laboratory prototype of 1.5 kW for each type of the proposed system is implemented to demonstrate the suitability of the proposed concept. Details of the experimental setup and the simulation results as well as experimental verification of the proposed systems under different operating conditions are presented.

5.2 Type 1: Multiple module-casings

5.2.1 Steady-State analysis and System Operation

Type 1 system has one module per casing as depicted in Fig. 5.1. Series cascaded connection of multiple module-casings forms the entire energy storage system. Each module-casing arrangement in this system is an identically constructed IPT system. The secondary side reversible rectifier is connected to a DC link capacitor. Each output capacitor of the module-casing setup is cascaded to form the DC link for the three-phase VSI drive as depicted in Fig. 5.3. The steady-state analysis of an IPT system was detailed in [81] and a brief of which is given in this section.

Following analysis is for k^{th} module-casing setup neglecting the resistive losses. $V_{m,k}$ is generated by the primary side converter which causes the current $I_{p,k}$ in the primary or module side coil. This primary current induces the voltage $V_{ci,k}$ in the secondary or casing side coil which is given by,

$$V_{ci,k} = j\omega M_c I_{p,k} \quad (5.1)$$

Similarly the current $I_{s,k}$ in the casing side coil induces voltage in the module side coil and is given by,

$$V_{mr,k} = -j\omega M_c I_{s,k} \quad (5.2)$$

where M_c is the mutual inductance between the module side and casing side coil in k^{th} IPT system. M_c is considered same for all the k IPT systems and is assumed to be constant as there is no movement required for the IPT coils in the proposed application.

By T- π equivalence analysis of the LCL circuit, the primary side inductor currents of the k^{th} IPT system are given by,

$$I_{m,k} = j \frac{V_{mr,k}}{\omega L_{p,k}} \quad (5.3)$$

$$I_{p,k} = -j \frac{V_{m,k}}{\omega L_{m,k}} \quad (5.4)$$

Similarly the secondary side inductor currents are given by,

$$I_{s,k} = j \frac{V_{c,k}}{\omega L_{s,k}} \quad (5.5)$$

$$I_{c,k} = -j \frac{V_{ci,k}}{\omega L_{c,k}} \quad (5.6)$$

The input power from the battery module of k^{th} IPT system is given by

$$P_{b,k} = \text{Re}\{V_{m,k} I_{m,k}^*\} = \frac{|V_{m,k}| V_{c,k} M_c \sin \theta_k}{\omega L_{m,k} L_{s,k}} \quad (5.7)$$

It can be concluded from (5.7) that the amount of power extracted from the primary side is controlled by either $|V_{m,k}|$ or θ_k which is the relative phase angle between $V_{c,k}$ and $V_{m,k}$. The magnitude of voltage $V_{c,k}$ depends on the capacitor voltage $V_{d,k}$ which in turn depends on the power extracted from the primary side and the load demand. In the proposed system, the power transferred from the battery should compensate for the power required by the three-phase VSI drive, power required by the capacitor to maintain its voltage and the losses in the IPT system.

Fig. 5.3 Electric circuit representation of the proposed system

For the three-phase inverter operation, a constant DC link voltage is required. Hence the operation of IPT system aims at maintaining a constant total DC link voltage which is the sum of voltage outputs of each module-casing IPT system. In the proposed system, the converters at either side of the wireless interface are operated in the square wave mode. The power transferred between the primary and secondary coil is controlled by manipulating the relative phase angle θ_k between the voltages $V_{m,k}$ and $V_{c,k}$. The power transferred will be maximum for $\theta_k = \pi/2$ and zero for $\theta_k = 0$, for given magnitude of $V_{m,k}$ and $V_{c,k}$. The maximum value of $V_{c,k}$ varies with changes in the capacitor voltage $V_{d,k}$. The $V_{d,k}$ in turn depends on the relative phase angle which controls the amount of power extracted from the primary coil through the wireless interface. The power flow direction depends on the sign of the relative phase angle between $V_{m,k}$ and $V_{c,k}$. For motoring operation, power should be transferred from battery to motor for which the $V_{c,k}$ should lag $V_{m,k}$ as depicted in Fig. 5.4. Whereas for regenerative braking operation, power should be transferred from motor to the battery for which the $V_{c,k}$ should lead $V_{m,k}$.

The output voltage of k^{th} IPT system is controlled by varying the corresponding power transferred by manipulating the relative phase angle θ_k between $V_{m,k}$ and $V_{c,k}$ as explained earlier. Each module-casing is controlled independently to regulate its output voltage and hence the amount of power transferred. The total DC link voltage V_{dc} is the sum of k capacitor voltages $V_{d,k}$. Hence the reference value of required output voltage $V_{d,k}^*$ is obtained from the required DC link voltage as follows:

$$V_{d,k}^* = \frac{V_{dc}^*}{\beta_k} \quad (5.8)$$

where β_k is varied to achieve any desired performance. If β_k is varied, then consequently $V_{d,k}^*$ varies affecting the power extraction. The voltage reference setting for each of the IPT system can be used to implement the control of power extraction from each battery micro-pack. This could be based on the state of charge reference or any fault detection based algorithms.

5.2.2 Averaged State-Space Modelling

Fig. 5.4 Equivalent circuit of one module-casing set up over half switching period T'_{sw}

A dynamic multivariable state-space model for the IPT system has been discussed in [94]. In this section, an averaged state space modelling of the proposed system is discussed. The IPT system in the module-casing can be represented as an equivalent circuit which is illustrated with dynamic variables in Fig. 5.4, in which the inverter and the motor are represented as current source in the equivalent circuit. The relative phase angle θ is expressed as the duty ratio D in terms of the switching period T'_{sw} ($=T_{sw}/2$). Since all the module-casings contain identical components, analysis of a single battery module-casing combination suffices to investigate the proposed system. The IPT system will be operated at a frequency higher than the inverter switching frequency. The model equations are developed for a period which is half of the regular switching period of this IPT system. The induced or reflected voltage on the mutually coupled coils are modelled as an AC source as shown in Fig. 5.4 and are given by,

$$v_{mr} = -M \frac{di_s}{dt} \quad (5.9)$$

$$v_{cr} = M \frac{di_p}{dt} \quad (5.10)$$

The duty ratio D between the voltages v_m and v_c is shown in Fig. 5.4. The state equations are based on the circuit configuration over the period DT'_{sw} and $(1-D)T'_{sw}$. The state equations are given by,

$$\frac{di_m(t)}{dt} = \left(\frac{1}{L_m}\right)v_m - \frac{1}{L_m}v_p - \frac{R_m}{L_m}i_m \quad (5.11)$$

$$\frac{dv_p(t)}{dt} = \frac{1}{C_p}i_m - \frac{1}{C_p}i_p \quad (5.12)$$

$$\frac{di_p(t)}{dt} = \frac{1}{L_p}\left(1 + \frac{M^2}{b}\right)v_p + \frac{R_p}{L_p}\left(1 + \frac{M^2}{b}\right)i_p - \frac{MR_s}{b}i_s - \frac{M}{b}v_s \quad (5.13)$$

$$\frac{di_s(t)}{dt} = \frac{M}{b}v_p - \frac{R_p}{b}i_p - \frac{R_sL_p}{b}i_s - \frac{L_p}{b}v_s \quad (5.14)$$

$$\frac{dv_s(t)}{dt} = \frac{1}{C_s}i_s - \frac{1}{C_s}i_c \quad (5.15)$$

$$\frac{di_c(t)}{dt} = \frac{1}{L_c}v_s - \frac{1}{L_c}v_c - \frac{R_c}{L_c}i_c \quad (5.16)$$

$$\frac{dv_d(t)}{dt} = \frac{1}{C_d}i_c - \frac{1}{C_d}I_L \quad (5.17)$$

$$b = L_pL_s - M^2 \quad (5.18)$$

$$\langle v_c \rangle = D(v_d) - (1-D)v_d \quad (5.19)$$

The state-space model averaging with the switching function have been studied and applied to various power electronic converters [95, 96]. The averaged state-space model of the proposed system considering the duty ratio D as input, v_m as disturbance and V_d as output is developed with the state variables and inputs averaged over the period T'_{sw} as given by,

$$\dot{X} = f(X) + g(X)u + p(X)P \quad (5.20)$$

$$y = h(X) \quad (5.21)$$

where,

$$X = [\langle i_m \rangle \langle v_p \rangle \langle i_p \rangle \langle i_s \rangle \langle v_s \rangle \langle i_c \rangle \langle v_d \rangle]^T$$

$$u = [\langle D \rangle]$$

$$P = [\langle v_m \rangle]$$

$$f(X) = \begin{bmatrix} \frac{-R_m}{L_m} x_1 + \frac{-1}{L_m} x_2 \\ \frac{1}{C_p} x_1 - \frac{1}{C_p} x_3 \\ \frac{1}{L_p} \left(1 + \frac{M^2}{b}\right) x_2 + \frac{R_p}{L_p} \left(1 + \frac{M}{b}\right) x_3 + \frac{-MR_s L_p}{b} x_4 + \frac{-ML_p}{b} x_5 \\ \frac{M}{b} x_2 + \frac{-R_p}{b} x_3 + \frac{-R_s L_p}{b} x_4 + \frac{-L_p}{b} x_5 \\ \frac{1}{C_s} x_4 - \frac{1}{C_s} x_6 \\ \frac{1}{L_c} x_5 - \frac{R_c}{L_c} x_6 \\ \frac{1}{C_d} x_6 - \frac{I_L}{C_d} \end{bmatrix}$$

$$g(X) = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ \frac{2}{L_c} x_7 \\ 0 \end{bmatrix}; \quad p(X) = \begin{bmatrix} \frac{-1}{L_m} \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}; \quad h(X) = x_7; \quad b = L_p L_s - M^2$$

5.2.3 DC link voltage control based on feedback linearization

To control the output voltage of casing side converter, the phase information of the v_m needs to be passed to the casing side controller. It can be done either through the wireless communication or through estimation from the measured values of the system states on the casing side [97]. The following controller is developed on the assumption that the phase information of v_m is available at the casing side. Each casing side converter has its own controller and is identical. This section contains the derivation of secondary

side controller based on non-linear control theory with feedback linearization technique [98-102]. The control objective is to control the output voltage of the casing side converter at a constant desired value even with perturbations on the load or source (battery module) side. By input-output linearization theory, a state feedback control law $u(\mathbf{X})$ has to be derived which makes the input-output map of the non-linear system to a linear map: $\dot{y} = v$. The state feedback control law can be obtained using:

$$u = \frac{-L_f^\rho h(\mathbf{X})}{L_g L_f^{\rho-1} h(\mathbf{X})} + \frac{v}{L_g L_f^{\rho-1} h(\mathbf{X})} \quad (5.22)$$

This state feedback decouples the output from the disturbances when i^{th} Lie derivative of $h(\mathbf{X})$ along $f(\mathbf{X})$ and then along $g(\mathbf{X})$ is equal to zero i.e., $L_p L_f^i h(\mathbf{X}) = 0$ for all $i = 0$ to $\rho - 1$

The derivation of Lie derivatives for the calculation of relative degree is provided in Appendix I. The relative degree of the output of the system under consideration is obtained as $\rho=2$ from the Lie derivatives $L_g h(\mathbf{X})=0$ and $L_g L_f h(\mathbf{X}) = 2x_7/C_d L_c$. Therefore from (5.22) the control law for the casing side controller is given by,

$$u = \frac{R_c x_6 - x_5 + L_c C_d v}{2x_7} \quad (5.23)$$

The obtained control law u represents the duty ratio D which is allowed to vary between 0 to 50 %, meaning that the maximum value for u can be 0.5 even when the v_d is zero. The relative phase angle θ is obtained from D using a direct look-up table, which correspondingly varies from 0 to $\pi/2$.

Verification of reduction of non-linear input-output relation to linear mapping:

$$\begin{aligned} L_f^2 h(\mathbf{X}) + L_g L_f h(\mathbf{X}) u &= \frac{1}{C_d L_c} x_5 - \frac{R_c}{C_d L_c} x_6 + \frac{2}{C_d L_c} x_7 u \\ &= \frac{1}{C_d} \dot{i}_c \\ &= \ddot{v}_d \end{aligned}$$

Thus it can be inferred that $\dot{y} = v$.

Verification of disturbance decoupling:

$$L_p h(\mathbf{X}) = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1] \begin{bmatrix} 1 \\ \frac{1}{L_m} \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = 0$$

$$L_p L_f h(\mathbf{X}) = \left[0 \ 0 \ 0 \ 0 \ 0 \ \frac{1}{C_d} \ 0 \right] \begin{bmatrix} 1 \\ \frac{1}{L_m} \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = 0$$

Thus $L_p L_f^i h(\mathbf{X}) = 0$ for all $i = 0$ to $\rho - 1$ is verified for the system.

Fig. 5.5 Block diagram of the control scheme

Fig. 5.6 Block diagram of the DC link voltage loop

The obtained non-linear feedback control law (5.23) is associated with the linear control of the output dynamics $\dot{y} = v$. The block diagram of the non-linear control strategy of the output voltage of the casing side converter is shown in Fig. 5.5. The PI controller

can be designed based on the block diagram of the DC link voltage loop of an IPT system is given in Fig. 5.6.

5.2.4 Design Concept and Simulation results of 80 kW system – Type 1 Topology

A design guide line to construct battery micro-packs for an EV is discussed in this section. The basic design parameters of the EV battery pack are maximum acceleration and the driving range. The maximum acceleration determines the maximum power required for the given vehicle weight and the maximum energy capacity of the battery pack is derived from the required driving range. The voltage of the battery pack should be in accordance with the rated line-line voltage of the drive motor. In-order to compare with the conventional battery pack design, a commercially available EV is taken into consideration. Nissan Leaf battery pack (with capacity of 24 kWh and nominal voltage of 360 V) has 192 Lithium ion cells of 30 Ah [84]. The battery micro-packs should provide similar performance like the battery pack of the considered EV. Hence, the parameters of this battery pack are used as required values for designing of the proposed systems.

The choice of number of micro-packs N affects the design of the micro-pack and the IPT system. In the following discussion, $N = 3$ is considered as an example to explain the design concept. It has to be noted that the number of modules could be optimized considering the factors like weight of the micro-packs, efficiency of the IPT system at specific power levels etc. With $N = 3$ and total number of lithium ion cells to be 192, each battery micro-pack could be constructed with 64 cells connected in series. Hence each battery micro-pack would have the nominal voltage of 240 V and nominal capacity of 7.92 kWh.

The design of inductive interface for the proposed application is based on the primary side and secondary side voltages, and the maximum power required to be transferred. Assuming $L_m = L_p = L_s = L_c = L$, $C_p = C_s = C$ and $\theta = 90^\circ$ for the maximum power transfer, the following expressions can be used in designing the inductive interface.

$$L = \frac{V_b V_d k_c}{\omega P_{max}} \quad (5.24)$$

$$C = \frac{1}{\omega^2 L} \quad (5.25)$$

Where k_c is the coupling co-efficient between the module side and casing side coils, ω is the operating frequency.

The main objective of this proposed system is to increase the ride through capability during fault conditions. The total power from the battery micro-packs with inductive interface should be 80 kW to match the performance of the considered example car (Nissan Leaf in this case). In-order to achieve this with at least two micro-packs in operation, P_{max} of each inductive interface should be at least 40 kW. Therefore for this design, P_{max} is taken as 50 kW considering the losses in the inductive interface. Operating frequency of 10 kHz, coupling coefficient of 0.6, P_{max} to be 50 kW (for $N = 3$), and V_b to be 240 V are taken as the desired parameters. For Type 1 system with multiple module-casings, the total DC link voltage at the three-phase inverter input is the sum of output voltages of each module-casing. Hence V_d should be 120 V for $N = 3$ and V_{dc} to be 360 V. Therefore for the Type 1 system, L and C are calculated to be 8.25 μ H and 30.7 μ F respectively. It has to be noted that even when the battery micro-pack is supplying 40 kW, the discharge is about 5 C-rate which is still under the operating limit. The design guidelines explained earlier shows that the proposed systems are technically feasible and hence are applicable to EVs. However the mechanical design and thermal management of the battery micro-packs in the proposed systems would be challenging.

Simulation model of the proposed Type 1 system of 80 kW were developed in Simulink PLECS environment in order to study the performance under various operating conditions. The proposed system acts as a power source to the three-phase inverter. For the operation of three-phase inverter a constant DC link voltage is required. Hence the primary objective in these systems is to maintain a constant DC link voltage. This is achieved by supplying enough amounts of power to meet the load demand and the power required by the DC link capacitor. This simulation was conducted to verify that the power extracted from each battery micro-pack is independently and simultaneously controllable along with the primary objective of maintaining a constant DC link voltage. The simulations were carried out for three scenarios: equal amount of power extracted from each of the battery micro-pack, unequal amount of power extracted and one faulty micro-pack unable to contribute to the load demand.

The capacitor voltages at the DC link of three-phase inverter is shown in Fig. 5.7 and the corresponding input powers from each battery micro-pack is given in Fig. 5.8.

During the period $0 \leq t \leq 1$ s, three battery packs contribute equally to the load demand of 78 kW while maintaining the DC link voltage of 360 V by maintaining 120 V at the output of each IPT. During the period $1 \leq t \leq 2$ s, 78 kW is supplied with 28.5 kW from first module, 26.4 kW from second module and 24.3 kW from third module. During the period $2 \leq t \leq 3$ s, the second module is considered faulty and removed from operation and the remaining two battery packs contribute 39.5 kW each to the load demand of 78 kW. The difference between the sum of input power from the battery micro-packs and the input power to the three-phase inverter is due to the loss in the IPT system.

Fig. 5.7 DC Link Voltages for Type 1 system

Fig. 5.8 Battery micro-pack power and input power to three-phase inverter for Type 1 system

Each module-casing setup in Type 1 system is operated independently and it requires no synchronization between them. This is illustrated in Fig. 5.9 which shows the currents in the primary side coils (lower plot) and the total DC link voltage (upper plot). To verify the robustness of the system, a disturbance was created at 0.25 s by changing the phase of the second module. For the period $t \leq 0.25$ s, all three modules were operated synchronously in same phase and for $t > 0.25$ s, second module is operated with a phase difference of 60° from first and second module, which is reflected in the primary side coil current traces in Fig. 5.9. The operation the system was not disturbed by this asynchronous operation which is clear from the DC link voltage.

Fig. 5.9 Type 1 system under asynchronous operation

5.2.5 Experimental Setup of 1.5 kW System – Type 1 Topology

A laboratory prototype of the proposed system with $N = 2$ was constructed to demonstrate the concept. The proposed application requires transferring power over a shorter distance of 10 mm – 20 mm. Another advantage in this application is that the coils are stationary all the times during the power transfer operation. There are two types of coils that are suitable for the proposed application: planar coils and coils with cores. A detailed analysis of various types of pads with planar coils was given in [103]. For this demonstration, an E shaped ferrite core based power transfer coils was built and used. The E80/38/20 3c90 from ‘Ferrotec’ was used for this study. Both primary and secondary cores were built by stacking four E cores. ‘Type 2 50/32’ Litz wire was used for the coil windings. The parameters of the coils built are given in

Table 5.1. A snap shot of the IPT coils built for the demonstration of the proposed systems is shown in Fig. 5.10.

Fig. 5.10 Experimental Setup – Type 1 Topology

The converters for IPT systems were built with MOSFETs and gate signals were generated using a DSP. A three-phase RL was used as the load. IGBT based three-phase inverter was controlled using dSPACE and a RL load was employed as load. Laboratory DC power supplies were employed in place of battery modules. The parameters of the experimental setup are given in Table 5.2. Simulation models with the parameters

specified were developed in Simulink PLECS environment in order to study the performance under various operating conditions.

Table 5.1 IPT Coil Parameters at 20 kHz – Type 2

	Self-Inductance (μH)	ac resistance ($\text{m}\Omega$)
Coil 1	29.8	54
Coil 2	29.4	53.2
Coil 3	29.7	50.8
Coil 4	29.7	52

The simulation and experimental verification of the operation of the proposed systems were carried out for three scenarios: case 1 – equal amount of power extracted from both the inputs, case 2 – unequal amount of power extracted from the input sources and case 3 – one faulty module unable to contribute to the load demand. The desired DC link voltage for the three-phase inverter was chosen to be 50 V. The three-phase inverter was operated at the modulation index of 0.9 with a constant three-phase RL load.

Table 5.2 Setup parameters – Type 1

Parameters	1.5 kW System (Prototype and simulation model)	80 kW System (simulation Model)
$V_{b,k}$	50 V	240 V
$L_{m,k}; L_{c,k}$	30 μH	8.25 μH
$C_{p,k}; C_{s,k}$	2.2 μF	30.7 μF
$C_{in}; C_d$	200 μF	6600 μF
Operating Frequency f (IPT)	19.6 kHz	10 kHz
Coupling Coefficient k_c	0.6	0.6

The power converters on either side of module-casing were operated in the square wave mode and the amount of power transferred over the air gap was controlled by manipulating the relative phase angle between the voltages. The relative phase angle between $V_{m,k}$ and $V_{c,k}$ could be varied from 0 to $\pi/2$ radians for zero to maximum power flow. The relative phase angle θ_k was manipulated according to the DC link voltage reference $V_{d,k}^*$ for each of the IPT system. The total DC link voltage V_{dc} is the sum of the

two capacitor voltages $V_{d,k}$ for $i = 1, 2$. Hence the reference value of the required output voltage $V_{d,i}^*$ was obtained from the required DC link voltage as follows:

$$V_{d,i}^* = V_{dc}^* / \beta_i \quad (5.26)$$

In (5.26), β_i was varied to execute the three scenarios – equal power extraction, unequal power extraction and one module being faulty. For instance, $\beta_i = 2$ extracts equal amount of power from each battery module. This can be included in the control strategy to handle any State of Charge imbalances amongst the battery micro-packs.

Case 1:

To extract equal amount of power from each of the input modules, the voltage references for each of the IPT system was set at 25 V so that the total DC link voltage could be 50 V. The voltage and currents of the IPT components during this scenario are presented in Fig. 5.11. Both the primary side converters were operated in square wave mode at the IPT resonant frequency of 19.6 kHz. The relative phase angle of $V_{c,k}$ with respect to $V_{m,k}$ was controlled so that output voltages $V_{d,k}$ were regulated at 25 V.

Case 2:

This scenario was carried out to emulate the condition of unequal power extraction from the battery micro-packs. The difference in the amount of power extracted from the micro-packs was obtained by setting different voltage references to the IPT system, with total desired DC link voltage set as 50 V. For this case, $V_{d,1}^*$ was set as 21 V and $V_{d,2}^*$ was set as 29 V. The corresponding voltage and current waveforms are shown in Fig. 5.12. Because of unequal output voltage, different amount of was extracted from the inputs to meet the same load demand.

Case 3:

This scenario was carried out to emulate the condition of one faulty module and to verify the efficacy of the proposed system for its redundancy. For this case, $V_{d,1}^*$ was set as 50 V and $V_{d,2}^*$ was set as 0 V. The second IPT system was completely turned off by keeping all the switches of the power converters open. The corresponding voltage and current waveforms of the active IPT system are shown in Fig. 5.13. The voltages and currents of the inactive IPT system are all zero and hence not shown here. The load demand was entirely supported by one IPT module by regulating the DC link voltage at

50 V. During this operation, the load current passes through the anti-parallel diodes of the secondary H-bridge converter of the inactive IPT system.

Load side three-phase inverter output:

The output voltage and current of the three-phase load side inverter when the DC link voltage regulated at 50 V are shown in Fig. 5.14. Both Type 1 and Type 2 systems were operated such that they act as a controllable power source to the three-phase inverter. The operation of three-phase inverter was controlled using the dSPACE controller. A fast space vector modulation was implemented to generate the necessary three-phase output.

(a)

Fig. 5.11 Voltage and current traces of the IPT components of Type 1 system for equal input power extraction - (a) Simulated and (b) Experimental

Fig. 5.12 Voltage and current traces of the IPT components of Type 1 system for unequal input power extraction - (a) Simulated and (b) Experimental

Fig. 5.13 Voltage and current traces of the IPT components of Type 1 system with one faulty input module - (a) Simulated and (b) Experimental

Fig. 5.14 Three-phase inverter output

5.3 Type 2: Single casing and multiple modules

5.3.1 Steady-State analysis and System Operation

Type 2 system would have a hollow casing attached to the vehicle and is designed with multiple compartments each capable of accommodating a battery module with IPT primary side as shown in Fig. 5.2. The IPT secondary side is built on the casing. This arrangement has multiple primary coils and one secondary coil of the IPT system. Each primary coil is coupled to the secondary coil with a mutual inductance M_c . The mutual inductance between the primary coils is M_{ij} . The advantage of this topology when compared with the Type 1 topology is that the single casing makes the mechanical and safety design of the vehicle to be relatively simpler. However, with this arrangement, the mutual coupling between the primary coils of the IPT system is inevitable and consequently the power is exchanged between the battery modules. The power circulation among the primary coils is undesirable as the battery modules are supposed to supply power to the motor driving the vehicle.

The equivalent circuit of the Type 2 system is given in Fig. 5.15. An expression for the power transferred between the module and casing is derived considering the mutual coupling between primary and secondary (M_c) as well as the mutual coupling between the primary coils (M_{ij}). A solution to improve the performance of the system by avoiding the problem of mutual coupling between primary coils is derived.

As depicted, the casing side of this type has one converter which should be rated accordingly with the three-phase drive inverter. Each of the primary or module side coil is coupled with one casing side coil. With N battery modules constituting the primary side of the IPT system, the LCL circuit of k^{th} IPT system is tuned to the frequency ω which is given by,

$$\omega^2 = \frac{1}{L_{m,k}C_{p,k}} = \frac{1}{L_{p,k}C_{p,k}} = \frac{1}{L_sC_s} = \frac{1}{L_cC_s} \quad (5.27)$$

Fig. 5.15 Electric circuit representation of Type 2 system: Multiple modules in single casing

The reflected voltage in the primary coils depends on the currents in the secondary coil as well as the current in the other primary coils. Thus the induced voltages in the respective coils of this type of IPT system because of the coupling through mutual inductance are given by,

$$V_{ci} = \sum_{k=1}^N j\omega M_{c,k} I_{p,k} \quad (5.28)$$

$$V_{mr,k} = -j\omega M_{c,k} I_s + \sum_{k=1, k \neq l}^{N-1} j\omega M_{kl} I_{p,k} \quad (5.29)$$

where $M_{c,k}$ is the mutual inductance between module side coil and casing side coil of k^{th} IPT system and M_{kl} is the mutual inductance between primary side coils of k^{th} and l^{th} IPT systems.

The primary side inductor currents of the k^{th} IPT system are given by,

$$I_{m,k} = j \frac{V_{mr,k}}{\omega L_{p,k}} \quad (5.30)$$

$$I_{p,k} = -j \frac{V_{m,k}}{\omega L_{m,k}} \quad (5.31)$$

The secondary side inductor currents are given by,

$$I_s = j \frac{V_c}{\omega L_s} \quad (5.32)$$

$$I_c = -j \frac{V_{ci}}{\omega L_c} \quad (5.33)$$

Neglecting the switching and conduction losses of the H-bridge converter interfacing the battery module and the primary side of the IPT system, the input power from k^{th} battery module is given by,

$$P_{m,k} = \text{Re}\{V_{m,k} I_{m,k}^*\} \quad (5.34)$$

Using (5.29) and (5.30),

$$I_{m,k} = j \frac{-j\omega M_{c,k} I_s + \sum_{k=1, k \neq l}^{N-1} j\omega M_{kl} I_{p,k}}{\omega L_{m,i}} \quad (5.35)$$

Therefore,

$$P_{b,k} = |V_{m,k}| \frac{V_c \frac{M_{c,k}}{L_s} \sin \theta_k - \sum_{k=1, k \neq l}^{N-1} M_{kl} \frac{V_{m,k}}{L_{m,k}} \sin \theta_{kl}}{\omega L_{m,k}} \quad (5.36)$$

$$= P_{(b-c),k} - \sum_{l=1, l \neq k}^{N-1} P_{(b-b),kl} \quad (5.37)$$

where θ_{kl} is relative phase angle between $V_{m,k}$ and $V_{m,l}$ of k^{th} and l^{th} modules; θ_k is the relative phase angle between $V_{m,k}$ of k^{th} module and V_c of casing side. Thus the power extracted from the battery module contains two terms: the first term represents the power transferred to the secondary coil or casing represented as $P_{(b-c),i}$ and the second term is the power transferred to the rest of the module side coils represented as $P_{(b-b),i}$. The power exchange between the battery modules in the primary side is undesirable during motoring operation. Therefore to nullify the circulating power amongst battery modules in the primary side, the relative phase angle θ_{kl} should be zero. Meaning that $V_{m,k}$ for all $k = 1$ to N should be in same phase. And to maximise the power transferred from primary side to secondary side of the wireless interface θ_k for all $k = 1$ to N should be 90° .

The output power of the IPT system at the secondary side can be derived in the similar manner and is given by,

$$P_o = \text{Re}\{V_c I_c^*\} \quad (5.38)$$

$$= \frac{V_c M_c}{\omega L_s} \sum_{i=1}^n \frac{V_{m,i}}{L_{p,i}} \sin \theta_i \quad (5.39)$$

$$= \sum_{i=1}^n P_{(m-c),i} \quad (5.40)$$

Thus output power is the sum of input power from all the battery micro-packs (if losses are not considered) at the primary side, provided $V_{m,k}$ are all in phase. The three-phase inverter operation requires a constant DC link voltage across the capacitor C_d at the

output of the casing side converter. The voltage at the input of the casing side converter i.e., V_c determines the maximum output voltage of the converter. From (5.39), it is clear that V_c depends on P_o which in turn depends on $P_{(m-c),k}$ and hence on $V_{m,k}$. Hence for the type 2 system, the power transferred is controlled by manipulating the required voltage $V_{m,k}$. The H-bridge inverter at primary side is controlled using phase shift modulation and hence the average output is controlled by varying the phase shift $\alpha_{p,k}$ between the two legs of the H-bridge converter. The DC link voltage is controlled by varying the power transferred from each input module which is achieved by manipulating the phase shift $\alpha_{p,k}$. The casing side converter is operated as full bridge rectifier for motoring operation. Each module side converter is controlled independently according to the power reference. The power reference setting for each IPT system can be used to implement the control of power extraction from each battery micro-pack. This could be based on the state of charge reference or any fault detection based algorithms. The real time implementation of this system requires wireless communication on either side of the inductive interface which is not addressed in this work.

5.3.2 Design Concept and Simulation results of 80 kW system – Type 2 Topology

The battery micro-pack design concept is same as that of the Type 1 topology as explained in Section 5.2.4. With $N = 3$ and total number of lithium ion cells to be 192, each battery micro-pack could be constructed with 64 cells connected in series. Hence each battery micro-pack would have the nominal voltage of 240 V and nominal capacity of 7.92 kWh. The design of inductive interface of the Type 2 topology for the proposed application is also done using the equations (5.24) and (5.25).

The main objective of this proposed system is to increase the ride through capability during fault conditions. The total power from the battery micro-packs with inductive interface should be 80 kW to match the performance of the considered example car (Nissan Leaf in this case). In-order to achieve this with at least two micro-packs in operation, P_{max} of each inductive interface should be at least 40 kW. Therefore for this design, P_{max} is taken as 50 kW considering the losses in the inductive interface. Operating frequency of 10 kHz, coupling coefficient of 0.6, P_{max} to be 50 kW (for $N = 3$), and V_b to be 240 V are taken as the desired parameters. For the Type 2 system with one casing, the V_d should be same as $V_{dc} = 360$ V. Therefore for type 2 system, L and C are calculated to be 16.5 μ H and 15.35 μ F respectively.

Simulation model of the proposed Type 2 system of 80 kW were developed in Simulink PLECS environment in order to study the performance under various operating conditions. The proposed system acts as a power source to the three-phase inverter. For the operation of three-phase inverter a constant DC link voltage is required. Hence the primary objective in these systems is to maintain a constant DC link voltage. This is achieved by supplying enough amounts of power to meet the load demand and the power required by the DC link capacitor. This simulation was conducted to verify that the power extracted from each battery micro-pack is independently and simultaneously controllable along with the primary objective of maintaining a constant DC link voltage. The simulations were carried out for three scenarios: equal amount of power extracted from each of the battery micro-pack, unequal amount of power extracted, and with one faulty micro-pack unable to contribute to the load demand.

The DC link voltage at the input of three-phase inverter is shown in Fig. 5.16 and the corresponding input powers from each battery micro-pack is given Fig. 5.17. Fault may occur in any of the battery modules in the primary side of the IPT system. To ensure uninterrupted operation of the vehicle, the output of remaining active battery modules need to be adjusted to meet the load demand. This scenario has been simulated by setting α_p of the corresponding module side converter to zero and proportionately adjusting the remaining module side converter phase shift angles. For $0 \leq t \leq 1$ s, the second module is considered faulty and removed from operation and the remaining two battery packs contribute 40.5 kW each to the load demand of 80 kW.

A different amount of power has to be extracted from the battery modules at the primary side of the IPT system, as the proposed system has multiple battery modules and there might exist an imbalance condition in their SOC's. This is achieved by varying α_p of the individual converter at the module side. For $1 \leq t \leq 2$ s, 80 kW load demand is supplied with 31.4 kW from first module and third module each, and 18.2 kW from second module.

When all battery modules with their IPT system are in operation, the output power demand should be distributed equally amongst all the battery modules so that the SOC of the battery modules remain equal. For $2 \leq t \leq 3$ s, three battery packs contribute equally to the load demand of 80 kW while maintaining the DC link voltage of 360 V. The

difference between the sum of input powers from the battery micro-packs and the input power to the three-phase inverter is due to the loss in the IPT system.

Fig. 5.16 DC Link Voltages for Type 2 system

Fig. 5.17 Battery micro-pack power and input power to three-phase inverter for Type 2 system

5.3.3 Experimental Setup of 1.5 kW System – Type 2 Topology

A laboratory prototype for the type 2 systems with $N = 2$ was constructed to demonstrate the concept. Like the Type 1 topology, the Type 2 requires transferring power over a shorter distance of 10 mm – 20 mm. For Type 2 system, the primary side cores were made by staking three E cores and the secondary core was made by staking seven E cores. ‘Type 2 50/32’ Litz wire was used for the coil windings. The parameters of the coils built are given in Table 5.3. A snap shot of the experimental setup is shown in Fig. 5.18.

Table 5.3 IPT Coil Parameters at 20 kHz – Type 1

	Self-Inductance (μH)	ac resistance ($\text{m}\Omega$)
Coil 1	30.4	55
Coil 2	29.9	56
Coil 3	30.9	59

Table 5.4 Setup parameters – Type 2

Parameters	1.5 kW System (Prototype and simulation)	80 kW System (Simulation)
$V_{b,k}$	50 V	240 V
$L_{m,k}; L_{c,k}$	30 μH	16.5 μH
$C_{p,k}; C_{s,k}$	2.2 μF	15.35 μF
$C_{in}; C_d$	200 μF	6600 μF
Operating Frequency f (IPT)	19.6 kHz	10 kHz
Coupling Coefficient k_c	0.6	0.6

Fig. 5.18 Experimental setup – Type 2 Topology

The converters for IPT systems were built with MOSFETs and gate signals were generated using a DSP. IGBT based three-phase inverter was controlled using dSPACE and a RL load was employed as load. Laboratory DC power supplies were employed in place of battery modules. The parameters of the experimental setup are given in Table 5.4. Simulation models with the parameters specified were developed in Simulink PLECS environment in order to study the performance under various operating conditions.

Type-2 system was built with two primary sides, representing two modules, and one secondary side for single casing. Like in Type-1 system, the objective here was also to maintain a constant DC link voltage to the three-phase inverter. The approach to control the amount of power transferred requires manipulation the primary or module side input voltage $V_{m,k}$. The output voltage of the module side converter $V_{m,k}$ was controlled by manipulating the phase shift $\alpha_{p,k}$ between the two legs of the H-bridge converter. The RMS voltage of $V_{m,k}$ is maximum when $\alpha_{p,k}$ is π radians and zero when $\alpha_{p,k}$ is 0 radians. The secondary or the casing side converter was operated as a full bridge rectifier when the power flow is from the source to load. The module side converters were controlled such that $V_{m,1}$ and $V_{m,2}$ were in same phase, in order to alleviate the problem of circulating powers amongst the primary coils. The change in $V_{m,k}$ causes variations in the three-phase inverter DC link voltage V_{dc} . In this manner, difference in the input power extracted for a given output power was achieved similar to type 1 system. Thus by manipulating the power extracted from each module independently, the DC link voltage can be regulated for the changes in the load demand. State of Charge imbalances amongst the battery micro-packs can also be managed in this type.

Case 1: Both the primary side converters was operated to ensure that the power extracted was equal and the DC link voltage was regulated at 50 V. Corresponding voltage and current waveforms of the IPT system are presented in Fig. 5.19.

Case 2: One of the primary sides was controlled to supply more power and other one to supply lesser power. This demonstrates the proposed system's ability to operate in unequal and independent power control. Corresponding voltage and current waveforms of the IPT system are presented in Fig. 5.20.

Case 3: For this case, $V_{m,2}^*$ was set as 0 V. The corresponding voltage and current waveforms of the IPT systems are shown in Fig. 5.21. The entire load demand was supported using one input module with the DC link voltage regulated at 50 V. Even though the second input was commanded to be zero, there was an induced current in the primary coil of the inactive module because of its mutual inductance with the other active primary coil. This is potentially uncontrollable which could be considered as disadvantage when compared with Type-1 system.

Fig. 5.19 Voltage and current traces of the IPT components of Type 2 system for equal input power extraction - (a) Simulated and (b) Experimental

Fig. 5.20 Voltage and current traces of the IPT components of Type 2 system for unequal input power extraction - (a) Simulated and (b) Experimental

(a)

(b)

Fig. 5.21 Voltage and current traces of the IPT components of Type 2 system with one faulty input module - (a) Simulated and (b) Experimental

5.4 Summary

- ❖ Two topologies of modularized battery systems with inductive coupling between battery pack and EV have been proposed for battery swapping.
- ❖ The operation and control of the proposed systems has been described, presenting steady state analysis and simulation results of a 24 kWh 80 kW system.
- ❖ Laboratory prototypes of 1.5 kW systems with IPT interfaces have been implemented and the results from both simulations and the laboratory prototypes have been presented and discussed.

Table 5.5 Comparison of Type 1 and Type 2 IPT MBESS

Type 1: Multiple module-casings	Type 2: Single casing and multiple modules
More number of power electronic components, and hence the complex controller is required. Also the power loss in the secondary side is higher.	Lesser number of power electronic components, and hence the loss at the secondary side is comparatively lesser.
IPT coils design and construction is easy as there is one primary coil for one secondary coil.	IPT coils design and construction for multiple primary coils for one secondary coil is relatively complex.
There is no power circulation problem among primary side coils.	Power circulation problem exists among the primary side coils, which can be solved by control methods. Also the circulating current in the primary of the bypassed module is unavoidable.
All micro-packs does not have to be placed at same place in the vehicle, as only the DC link is common for all the IPT system. This gives an extra degree of freedom in designing the vehicle.	All micro-packs have to be placed at same place in the vehicle as there is one secondary for all the primary coils.
Since the module-casings are identical and independent, there is a scope of individual slave controller, and one master controller architecture.	There is no scope for multiple slave controllers kind or architecture in Type 2.

- ❖ Based on the advantages and disadvantages of both types of IPT MBESS, it can be concluded that Type 1 can be considered as the preferred choice for EV applications.
- ❖ The results convincingly conclude that the power flow from the battery modules can be controlled independently. Thus the operational reliability and the availability of an EV could be potentially enhanced by employing the proposed system which offers modularity and makes the battery swapping process more safe, convenient and reliable.

Chapter 6 Comparison of the Proposed Systems with Conventional System for EV BESS

6.1 Introduction

This Chapter compares certain characteristics of the CHB, the DSCC and the IPT systems introduced in the previous chapters with the conventional BESS based on three-phase two-level inverter. Firstly, the number and capacity of the lithium-ion cell required to construct the battery micro-pack based BESS is compared. Secondly, the input current signature is compared. The input current profile gives the information regarding the harmonic composition, and hence would be useful for the loss analysis and thermal management design as well as in the DC link capacitor selection.

6.2 Battery Micro-pack Requirement Comparison

This section discusses a methodology for the design of modularized battery micro-packs using lithium ion cells for the systems under study, to show their feasibility for the EV application. The basic design parameters of the EV battery pack are the maximum power required for the maximum acceleration with the given vehicle weight, and the maximum energy capacity of the battery pack required for certain driving range. The terminal voltage of the battery pack (or the micro-pack) depends on the power converter topology and the drive motor voltage requirement. In this study, the chosen values for investigation are: the rated peak value of the line-line voltage of the drive motor given by $V_p = 400$ V and the rated power given by $P_{acc} = 100$ kW. These values are considered so that the micro-pack design can be compared with the conventional battery pack design of a commercially available electric car: Nissan Leaf battery pack (24 kWh and 360 V) comprising of 192 Lithium ion cells [84]. The Nissan Leaf battery architecture is as portrayed in Fig. 6.1. As it can be seen, it has 96 series connected modules of two 33 Ah Lithium ion cells in parallel.

Fig. 6.1 Nissan-Leaf battery pack architecture

In order to make a fair comparison for the number and capacity of the lithium ion cells required to construct battery micro-pack for the proposed systems, the similar architecture of Nissan-Leaf battery pack i.e., two cells in parallel for a module is considered. The design of micro-packs for the proposed systems has been discussed in the respective chapters. In this section, the design results are compiled together to compare the requirements. The design results are obtained on the common ground that the fault in one of the micro-pack (unit considering the power electronic components) does not affect the performance of the entire system. Therefore, the number of micro-packs is chosen as three per phase for CHB based BESS, three per arm for DSCC based BESS, and three for both the types of IPT based BESS. The design results are compiled in Table 6.1.

It can be seen that the micro-packs for IPT BESS could be constructed with almost same number of lithium ion cells, even with reduced capacity compared to the conventional single battery pack system with two-level inverter. For CHB BESS, the number of cells is increased by two times whereas the capacity is reduced by half compared to conventional design. For DSCC, the number of cells required is about eight folds higher than that for CHB, but the capacity is reduced by eight fold. It has to be noted that the design of Battery Management System (BMS) and thermal management would become complex with micro-pack system.

Table 6.1 Battery Micro-pack design results comparison

	Convention -al BESS	CHB MBESS	DSCC MBESS	IPT MBESS
Number of battery pack (or micro-pack)	1	9	18	3
Required micro-pack voltage	360 V	77 V	154 V	120 V
Required micro-pack power	100 kW	11.11 kW	5.5 kW	33.3 kW
Required current capacity	277.77 A	144 A	36 A	277.5 A
Choice of lithium ion cell Ah	33 Ah	15 Ah	4 Ah	28 Ah
Total number of cells required	192	378	1512	195

6.3 Input Current Analysis

The current drawn from the battery pack determines the losses in it and hence the heat generated. This influences the thermal management design of the battery pack. The harmonic composition of the input current can be used to study the behaviour of the battery, as well as can be used for aging analysis of the battery pack. The input current composition is required to be analysed to optimize the DC-link filter Capacitor selection. The input current of the power converters employed for motor drives has different signatures depending on the type of the converter. The switching strategy and the switching frequency also affect the harmonic composition of the input current. In this section, the input current signature is analysed for its harmonic composition for the conventional two-level inverter, CHB and IPT based systems. As discussed in the previous section, number of lithium-ion cells required to construct the battery micro-packs for the DSCC based MBESS is quite high compared to the other systems. This is an inevitable disadvantage of the DSCC MBESS as it increases the complexity in the pack design. Therefore, the input current profile comparison in this section does not include the DSCC MBESS. A theoretical analysis based on the switching function for the computation of the inverter input current as well as the time based simulation results are presented in this section.

6.3.1 Three-phase Two-level Inverter based BESS

Conventionally, a typical EV has a single battery pack (with a DC-link Capacitor bank) connected to a three-phase two-level inverter driving the traction motor as portrayed in Fig. 6.2. The commonly used modulation strategy - Sinusoidal Pulse Width Modulation (SPWM) is considered here to study the signature of the inverter input current. A method to obtain the expression for inverter input current based on the switching function expression and the inverter output current was given in [104] and is briefed here.

Fig. 6.2 Three-phase Two-level Inverter Motor drive

The switching pulses generated by SPWM can be expressed in Fourier series as follows:

$$\left. \begin{aligned} S_A(\omega t) &= \sum_{m=-\infty}^{+\infty} H_m e^{jm\omega t} \\ S_B(\omega t) &= S_A\left(\omega t - \frac{2\pi}{3}\right) = \sum_{m=-\infty}^{+\infty} H_m e^{jm(\omega t - \frac{2\pi}{3})} \\ S_C(\omega t) &= S_A\left(\omega t + \frac{2\pi}{3}\right) = \sum_{m=-\infty}^{+\infty} H_m e^{jm(\omega t + \frac{2\pi}{3})} \end{aligned} \right\} \quad (6.1)$$

In pulse width modulation waveform, a typical k^{th} pulse is shown in Fig. 6.3. The switching function is defined as set of such pulses. The number of pulses in a fundamental frequency is based on the switching frequency. The modulating angle for k^{th} pulse is obtained using,

$$\alpha_{l,k} = \frac{\delta}{4} \left(1 + M \sin \left(\alpha_k - \frac{\delta}{4} \right) \right) \quad (6.2)$$

$$\alpha_{r,k} = \frac{\delta}{4} \left(1 + M \sin \left(\alpha_k + \frac{\delta}{4} \right) \right) \quad (6.3)$$

$$\delta = \frac{2\pi}{p} \quad (6.4)$$

Where M is the modulation index, α_k is the centre of k^{th} pulse from origin, and δ is the switching period defined by number of pulses ' p ' in the fundamental period.

Fig. 6.3 A typical pulse in a PWM waveform

The amplitude of the m^{th} harmonic in such k^{th} pulse is given by,

$$H_{m,k} = \frac{je^{-jm\alpha_k}}{2\pi m} [e^{jm\alpha_{l,k}} - e^{jm\alpha_{r,k}}] \quad (6.5)$$

Therefore the harmonic coefficient H_m for p pulses in the fundamental period is given by,

$$H_m = \sum_{k=1}^p H_{m,k} \quad (6.6)$$

The phase-A current can be expressed as,

$$i_A(\omega t) = \sqrt{2}I \cos(\omega t - \phi) \quad (6.7)$$

Similarly phase-B and phase-C currents are expressed as,

$$i_B(\omega t) = i_A \left(\omega t - \frac{2\pi}{3} \right) \quad (6.8)$$

$$i_C(\omega t) = i_A \left(\omega t + \frac{2\pi}{3} \right) \quad (6.9)$$

Therefore the total input current to the inverter is given by,

$$i_{inv} = i_A S_A + i_B S_B + i_C S_C \quad (6.10)$$

$$i_{inv} = \left\{ \begin{aligned} & \left[\sqrt{2}I \cos(\omega t - \phi) * \left(0.5 + \sum_{m=-\infty}^{+\infty} H_m e^{jm\omega t} \right) \right] + \\ & \left[\sqrt{2}I \cos\left(\omega t - \phi - \frac{2\pi}{3}\right) * \left(0.5 + \sum_{m=-\infty}^{+\infty} H_m e^{jm\left(\omega t - \frac{2\pi}{3}\right)} \right) \right] + \\ & \left[\sqrt{2}I \cos\left(\omega t - \phi + \frac{2\pi}{3}\right) * \left(0.5 + \sum_{m=-\infty}^{+\infty} H_m e^{jm\left(\omega t + \frac{2\pi}{3}\right)} \right) \right] \end{aligned} \right\} \quad (6.11)$$

This expression can be used to realize the input current signature and to compute the harmonic composition. The inverter input current waveform and its harmonic composition computed using Matlab is shown in Figs. Fig. 6.4(a) and Fig. 6.4(b) respectively.

The input current waveform obtained from the time based simulation of two-level inverter in Simulink is shown in Fig. 6.5(a). Corresponding harmonic composition is also shown in Fig. 6.5(b). It can be seen that the computed and simulated results are in good agreement.

Fig. 6.4 Two-level Inverter (a) DC link current trace and (b) its harmonic composition with $f_o=50$ Hz – Computed

Fig. 6.5 Two-level Inverter DC link current trace and its harmonic composition with $f_o=50$ Hz - Simulated

6.3.2 Cascaded H-bridge multi-level converter based MBESS

Cascaded H-bridge converter based battery energy storage system for electric vehicle has been described in Chapter 3. It consists of series connected of single phase H-bridge converter given in Fig. 6.6. The battery current depends on the switching of the H-bridge converter. The input current derivation of a single-phase H-bridge converter is as follows. The switching pulses of the two legs in a H-bridge converter can be expressed in Fourier series as follows:

$$\left. \begin{aligned} S_A(\omega t) &= \sum_{m=-\infty}^{+\infty} S_m e^{jm\omega t} \\ S_B(\omega t) &= S_A(\omega t - \pi) = \sum_{m=-\infty}^{+\infty} S_m e^{jm(\omega t - \pi)} \end{aligned} \right\} \quad (6.12)$$

Fig. 6.6 Single-phase H-bridge

The current drawn from leg-A can be expressed as,

$$i_A(\omega t) = \sqrt{2}I \cos(\omega t - \phi) \quad (6.13)$$

Similarly for leg-B is expressed as,

$$i_B(\omega t) = i_B(\omega t - \pi) \quad (6.14)$$

Therefore the total input current of the inverter is given by,

$$i_{inv} = i_A S_A + i_B S_B \quad (6.15)$$

$$i_{inv} = \left\{ \left[\sqrt{2}I \cos(\omega t - \phi) * \sum_{m=-\infty}^{+\infty} S_m e^{jm\omega t} \right] + \left[\sqrt{2}I \cos(\omega t - \phi - \pi) * \sum_{m=-\infty}^{+\infty} S_m e^{jm(\omega t - \pi)} \right] \right\} \quad (6.16)$$

This expression can be used to realize the input current signature and to compute the harmonic composition. The inverter input current waveform and its harmonic composition computed using Matlab is shown in Fig. 6.7(a) and Fig. 6.7(b) respectively. The input current waveform obtained from the time based simulation of CHB inverter in Simulink is shown in Fig. 6.8(a). Corresponding harmonic composition is also shown in Fig. 6.8(b). It can be seen that the computed and simulated results are in good agreement.

Fig. 6.7 CHB MBESS DC link current trace and its harmonic composition with $f_o=50$ Hz
– Computed

Fig. 6.8 CHB MBESS (a) DC link current trace and (b) its harmonic composition with $f_o=50$ Hz – Simulated

6.3.3 Inductive Power Transfer based MBESS

Both Type-1: Multiple module-casings and Type-2: Multiple modules in single casing IPT BESS has single-phase H-bridge converter like the one shown in Fig. 6.6 at the primary side connected to the battery micro-pack. Hence the same expression obtained in eqn (6.16) can be used to analyze the battery current in IPT MBESS also. The difference here is that the H-bridges in IPT-BESS are operated in either square wave mode or the quasi-square wave mode instead of PWM like in the CHB MBESS. The primary side converters' input current waveform and its harmonic composition computed using Matlab is shown in Fig. 6.9(a) and Fig. 6.9(b) respectively. The input current waveform obtained from the time based simulation of IPT MBESS in Simulink is shown in Fig. 6.10(a). Corresponding harmonic composition is also shown in Fig. 6.10(b).

Fig. 6.9 IPT MBESS (a) DC link current trace and (b) its harmonic composition with $f_o=10$ kHz – Computed

Fig. 6.10 IPT MBESS (a) DC link current trace and (b) its harmonic composition with $f_o=10$ kHz - Simulated

Fig. 6.11 Two-level Inverter DC link current's harmonic composition

Fig. 6.12 CHB Inverter DC link current's harmonic composition

Fig. 6.13 IPT Type-1 Input current's harmonic composition

Fig. 6.14 IPT Type-2 Input current's harmonic composition

Table 6.2 Summary of harmonic content in the input current for different systems with switching frequency as fundamental frequency

Frequency (Hz)	2 Level Inverter	CHB	IPT Type 1	IPT Type 2
5	2,4	1,2	2,4	2,4
20	2,4,8	1,2	2,4	2,6
45.5	4,8	1,2	2,4	2,6

Table 6.3 Summary of harmonic content in the input current for different systems with load current frequency as fundamental frequency

Frequency (Hz)	2 Level Inverter	CHB	IPT Type 1	IPT Type 2
5	66	2,22	-	-
20	-	20	-	-
45.5	72,78	5,50	-	-

The harmonic composition in the battery current depends on the load current frequency as well as the switching frequency of the converter. Fig. 6.11, Fig. 6.12, Fig. 6.13 and Fig. 6.14 shows the harmonic composition of the input current for different output frequencies of the two-level inverter, CHB, IPT Type-1 and IPT Type-2 respectively. Table 6.2 and Table 6.3 are the summary of input currents' harmonic composition of different systems showing the prominent harmonics in every category. The harmonic composition at different motor speeds gives an impression about the possible losses in the battery pack or micro-pack with respect to different power converters. In IPT systems, the motor speed has very negligible impact on the input current. But on the other hand, battery micro-packs in IPT would have high frequency components than the CHB and two-level inverter.

6.4 Summary

- ❖ The number and capacity of the lithium-ion cells required construct battery micro-pack has been compared.

	Convention -al BESS	CHB BESS	DSCC BESS	IPT BESS
Choice of lithium ion cell Ah	33 Ah	15 Ah	4 Ah	28 Ah
Total number of cells required	192	378	1512	195

- ❖ The input current profile and its harmonic composition have been compared. It can be concluded that the H-bridge based converter induces more harmonics in battery current than the conventional three leg inverters. This information is crucial for selecting breakers for the battery pack.

Chapter 7 Conclusions and Recommendations

7.1 Conclusions

The operational reliability and the down-time of EV are considered as the debilitating disadvantages for a widespread implementation of the EV based personal transport. It has been hypothesized that modularizing the battery pack into micro-packs and controlling the power independently should significantly increase the operational reliability of the EVs. To realize this, Cascaded H-bridge (CHB) and Double Star Chopper Cell (DSCC) based modular battery energy storage systems have been proposed and discussed. The down-time of any usual EV is higher than gasoline vehicles because the EV requires long periods of time for charging. To aid in overcoming this disadvantage, a unique solution using battery swapping concept has been proposed and discussed.

Modular battery pack system or micro-pack concept for CHB has been introduced to increase the availability of the EV in the event of fault. Review of the multi-dimensional pulse width modulation (MD-PWM) technique for a cascaded H-bridge inverter has been conducted. It may be concluded that this modulation technique facilitated simpler digital implementation of this technique for control of the CHB inverter. A control strategy has been proposed for the SOC balancing between the battery micro-packs in the CHB. This proposed control strategy aims to deliver a semblance of uniform operation of the whole battery system irrespective of the current SOC of each micro-pack. This is achieved by tasking the inter-module SOC balancing to deliver a steady output. The realization of fault tolerant operation using MD-PWM has also been proposed. The advantage of the proposed methods is that it can be conveniently integrated with the MD-PWM algorithm. The results obtained from the laboratory setup of a five-level three-phase CHB with six battery modules of 15-V 6.2-Ah have been presented for RL load. The results obtained from the laboratory setup of a five-level three-phase CHB driving an induction motor with six battery modules of 52.8-V 60-Ah have also been discussed. The results have verified that the fault tolerant and SOC balancing operation can be achieved with CHB using MD-PWM. This shows that the proposed micro-pack system with CHB would improve the EV performance in the event of fault, which is not possible with the conventional single battery pack and two-level converter designs. It has

to be mentioned that the proposed control strategies are independent of load, and as such it can be used for grid connected systems as well.

A modularized battery system with DSCC based modular multilevel converter has been proposed for a battery operated electric vehicle (EV). A conceptual design of a 28-kWh lithium ion based battery micro-pack system for DSCC has been described. Multidimensional pulse width modulation (MD-PWM) with integrated inter-module SOC balancing and fault tolerant control for DSCC has been proposed and explained. The methods of operation of the DSCC as an inverter and as a synchronous rectifier using MD-PWM with integrated inter-module SOC balancing and fault tolerant control have been discussed. The performances of the DSCC for SOC balancing and fault tolerant operation have been verified through simulations and the results have been presented. The results shows that the fault ride through capability with reduced capacity is possible with the DSCC based converter with battery micro-packs. This would certainly increase the availability of the EV when compared with the conventional battery pack system.

Another objective of reducing the down-time of an EV is achieved by implementing battery swapping regime. This is primarily realized with the implementation of a contactless interface between the battery micro-pack and the vehicle, and it has been discussed in this report. To realize a replaceable and modularized energy storage system with wireless interface, two system topologies have been proposed for a battery operated EV. Such an EV system derives substantial benefits from the synergies created by modularization and the decreased down-time. The two perspectives of analysis of the proposed systems focus upon the power converter operation and IPT system. The operation of the first topology has been analysed with an equivalent circuit and an averaged state-space model. A feedback linearization technique to establish the DC link voltage regulation based on the phase shift ratio has been implemented. The power flow is controlled using the module side converter to maintain the DC link voltage at the three-phase inverter driving the three-phase load. The second topology has been analysed with an equivalent electric circuit and the expressions for power flow have been derived using steady state analysis of the equivalent circuit. The working and power flow control of the proposed systems have been studied through simulations of a 24-kWh 80-kW system. The improvement in the operational reliability of EV has been verified by operating the system with one inactive battery micro-pack emulating the faulty situation. The

robustness of this technique is indicated given that the infallibility of the battery system during the scenarios of inactive or faulty battery micro-pack.

Laboratory prototypes of 1.5-kW systems for each topology with IPT interfaces have been implemented and the results from both simulations and the laboratory prototypes have been presented and discussed. The results convincingly conclude that the power flow from the battery modules can be controlled independently. Thus the operational reliability and the availability of an EV could be potentially enhanced by employing the proposed system which offers modularity and also would make the battery swapping process more safe, convenient and reliable. Nevertheless the increase in volume, weight and losses are inevitable in the proposed systems when compared to the conventional battery pack. Also further investigations on the IPT coil design for high power, mechanical design of the battery micro-packs with IPT system, and the design of thermal management would be required for real time implementation. The difficulties associated with high power IPT systems could be avoided by increasing the number of modules ' N ' so that power of each module is lesser which would then require finding the optimum value for ' N '.

The common conception among the customers in adapting to electric vehicles in view of reliability and the downtime of EVs to charge the batteries have been considered as the motivation for this study. This study has considered the power converters and control methods to realize independent power control with modular battery systems. It can be concluded from this study that Cascaded H-bridge based converter could be considered as the preferred choice for modular battery systems for EV, and Type 1 IPT MBESS could be considered as the preferred choice for replaceable modular battery system.

There are various other fields of study that are associated with the EV technology that are considered as main considerations for the EV manufacturers. It has to be mentioned that following are some considerations in the EV technology that has not been addressed in this study:

From Vehicle design perspective – EV occupant crash safety, occupant electrical safety, flexibility in battery charging interface, aftermarket installations etc.

From battery technology perspective – EV battery safety, battery durability, battery recycling etc.

From power system distribution perspective – battery chargers installations as either home-based or on-street or commercial, upstream interface with distribution system etc.

From other perspectives – Maintenance and repair, rescue and recovery etc.

In conclusion, it is believed that the verified results of the proposed methods enhance the reliability and the robustness of the EV operation. This will certainly serve to improve the prospects of this nascent sector and facilitate in expansion of EV technology.

7.2 Recommendations for Future Research

- ✓ Modularity in the BESS design for EV increases the operational reliability of the entire system, as discussed in this thesis. This research work focused completely on the system operation and control. A number of challenges are to be investigated for a successful realization of the proposed kinds of modular BESS for EV application. It includes economic analysis, energy efficiency, thermal and mechanical design, safety standards consideration, vehicle body design etc.
- ✓ Battery pack contributes major portion to the cost of any electric vehicle. Hence a through cost estimation could be done for the proposed systems, especially for the battery swapping concept.
- ✓ All the proposed systems have the advantage of modularity and are independent on the type of energy source. Hence, ESS hybridization could be explored. The integration of ultra-capacitor with the battery in the proposed systems could be investigated.
- ✓ The proposed methods for fault tolerant operation in modular systems are based on the assumption of fault signals are available. The investigation into the identification and signalling of fault should be done. Also fault isolation strategies for each of the discussed systems needs to be investigated.
- ✓ Though the systems under study are investigated for an EV application, the proposed modular converters can be extended to any grid integrated ESS. Thus the challenges associated with grid integration could be investigated. The CHB and DSCC would be advantageous for the ESS in the range of MWh, because of the modularity.

- ✓ Input current signature and its harmonic composition for various converter systems were investigated in this work. This information can be used to study the losses in the battery pack, which would be helpful for the thermal management design and the energy efficiency study. This would also be helpful to investigate the DC link capacitor selection.
- ✓ Smart charging or co-ordinated charging of EV and its impacts on the utility power system are studied by various researchers. In certain situations, co-ordinated charging requires charging EVs in a charging platform in different power levels. This can be easily achieved using the proposed systems because of the reduced power levels of the battery micro-packs with independent power control compared to the conventional high power EV battery pack. The challenges and benefits from this perspective can be investigated to further the benefits of modular BESS.

Author's Publications

1. Kandasamy, K; Vilathgamuwa, D.M.; Foo, G., "Inter-module SoC balancing control for CHB based BESS using multi-dimensional modulation," Industrial Technology (ICIT), 2013 IEEE International Conference on, vol., no., pp.1630,1635, 25-28 Feb. 2013.
2. Kandasamy, K; Vilathgamuwa, D.M.; Madawala, U.K.; Kuhn, R., "Contactless and replaceable modularized battery energy storage system for electric vehicles," Industrial Electronics Society, IECON 2013 - 39th Annual Conference of the IEEE , vol., no., pp.4594,4599, 10-13 Nov. 2013.
3. Kandasamy, K; Vilathgamuwa, D.M.; Tseng, King-Jet, "Double star chopper cell converter for battery electric vehicles with inter-module SoC balancing and fault tolerant control," Industrial Electronics Society, IECON 2014 - 40th Annual Conference of the IEEE , vol., no., pp.2991,2996, Oct. 29 2014-Nov. 1 2014.
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5. Kandasamy, K; Vilathgamuwa, D.M.; Madawala, U.K.; Tseng, King-Jet, "Inductively coupled modular battery system for electric vehicles", IET Power Electronics, 2015, DOI: 10.1049/iet-pel.2014.0553, IET Digital Library, <http://digital-library.theiet.org/content/journals/10.1049/iet-pel.2014.0553>.

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Appendix I Derivation for calculation of relative degree and Z domain system dynamics of the proposed Type 1 IPT-BESS system

Calculation of relative degree:

$$\frac{\partial h(\mathbf{X})}{\partial \mathbf{x}} = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1]$$

$$L_g h(\mathbf{X}) = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1] \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ \frac{2}{L_c} x_7 \\ 0 \end{bmatrix} = 0$$

$$L_f h(\mathbf{X}) = [0 \ 0 \ 0 \ 0 \ 0 \ 0 \ 1] \begin{bmatrix} \frac{-R_m}{L_m} x_1 + \frac{-1}{L_m} x_2 \\ \frac{1}{C_p} x_1 - \frac{1}{C_p} x_3 \\ \left(\frac{1}{L_p} + \frac{M^2}{b} \right) x_2 + \frac{R_p}{L_p} \left(1 + \frac{M}{b} \right) x_3 + \frac{-MR_s L_p}{b} x_4 + \frac{-ML_p}{b} x_5 \\ \frac{M}{b} x_2 + \frac{-R_p}{b} x_3 + \frac{-R_s L_p}{b} x_4 + \frac{-L_p}{b} x_5 \\ \frac{1}{C_s} x_4 - \frac{1}{C_s} x_6 \\ \frac{1}{L_c} x_5 - \frac{R_c}{L_c} x_6 \\ \frac{1}{C_d} x_6 - \frac{I_L}{C_d} \end{bmatrix} = \frac{1}{C_d} x_6 - \frac{I_L}{C_d}$$

$$\frac{\partial L_f h(\mathbf{X})}{\partial \mathbf{x}} = [0 \ 0 \ 0 \ 0 \ 0 \ \frac{1}{C_d} \ 0]$$

$$L_g L_f h(\mathbf{X}) = \left[0 \ 0 \ 0 \ 0 \ 0 \ \frac{1}{C_d} \ 0 \right] \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ \frac{2}{L_c} x_7 \\ 0 \end{bmatrix} = \frac{2}{C_d L_c} x_7 \neq 0$$

Thus the relative degree of the output of the system under consideration is obtained as $\rho=2$ from the Lie derivatives $L_g h(\mathbf{X})=0$ and $L_g L_f h(\mathbf{X}) = 2x_7/C_d L_c$.

Choosing the transformation matrix for co-ordinate transformation from X to Z domain:

Since the relative degree ρ is less than n , in the transformation matrix, $\phi(\mathbf{X}) = [\phi_1(x)\phi_2(x)\phi_3(x)\phi_4(x)\phi_5(x)\phi_6(x)\phi_7(x)]^T$, $\phi_1(x)$ and $\phi_2(x)$ are chosen from the Lie derivatives as follows.

$$\phi_1(x) = h(\mathbf{X}) = x_7$$

$$\phi_2(x) = L_f h(\mathbf{X}) = \frac{1}{C_d} x_6 - \frac{I_L}{C_d}$$

The remaining row of the transformation matrix has to be chosen so that its Jacobian matrix is non-singular to achieve diffeomorphism between the X and Z co-ordinates.

$$\frac{\partial \phi(x)}{\partial X} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 1 \\ 0 & 0 & 0 & 0 & 0 & \frac{1}{C_d} & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}$$

Therefore,

$$\mathbf{Z} = \phi(x) = \begin{bmatrix} z_1 \\ z_2 \\ z_3 \\ z_4 \\ z_5 \\ z_6 \\ z_7 \end{bmatrix} = \begin{bmatrix} x_7 \\ \frac{1}{C_d} x_6 - \frac{I_L}{C_d} \\ x_5 \\ x_4 \\ x_3 \\ x_2 \\ x_1 \end{bmatrix}$$

Similarly the inverse mapping is given by,

$$\mathbf{X} = \Phi^{-1}(Z) = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \\ x_5 \\ x_6 \\ x_7 \end{bmatrix} = \begin{bmatrix} z_7 \\ z_6 \\ z_5 \\ z_4 \\ z_3 \\ C_d z_2 + I_L \\ z_1 \end{bmatrix}$$

The dynamics of the system in Z domain can be obtained as follows:

$$\begin{aligned} \dot{z}_1 &= z_2 \\ \dot{z}_2 &= L_f^2 h(\Phi^{-1}(Z)) + L_g L_f h(\Phi^{-1}(Z))u \\ &= \frac{1}{L_c C_d} x_5 - \frac{R_c}{L_c C_d} x_6 + \left(\frac{2}{L_c C_d} x_7 \right) u \Big|_{x=\Phi^{-1}(Z)} \\ &= \frac{1}{L_c C_d} z_3 - \frac{R_c}{L_c} z_2 - \frac{I_L}{L_c C_d} + \left(\frac{2}{L_c C_d} z_1 \right) u \\ \dot{z}_3 &= L_f \Phi_3(\Phi^{-1}(Z)) + L_g \Phi_3(\Phi^{-1}(Z))u \\ &= \frac{1}{C_s} x_4 - \frac{1}{C_s} x_6 \Big|_{x=\Phi^{-1}(Z)} \\ &= \frac{1}{C_s} z_4 - \frac{C_d}{C_s} z_2 - \frac{I_L}{C_s} \\ \dot{z}_4 &= L_f \Phi_4(\Phi^{-1}(Z)) + L_g \Phi_4(\Phi^{-1}(Z))u \\ &= \frac{M}{b} x_2 + \frac{-R_p}{b} x_3 + \frac{-R_s L_p}{b} x_4 + \frac{-L_p}{b} x_5 \Big|_{x=\Phi^{-1}(Z)} \\ &= \frac{M}{b} z_6 + \frac{-R_p}{b} z_5 + \frac{-R_s L_p}{b} z_4 + \frac{-L_p}{b} z_3 = L_f \Phi_5(\Phi^{-1}(Z)) + L_g \Phi_5(\Phi^{-1}(Z))u \\ &= \left(\frac{1}{L_p} + \frac{M^2}{b} \right) x_2 + \frac{R_p}{L_p} \left(1 + \frac{M}{b} \right) x_3 + \frac{-MR_s L_p}{b} x_4 + \frac{-ML_p}{b} x_5 \Big|_{x=\Phi^{-1}(Z)} \\ &= \left(\frac{1}{L_p} + \frac{M^2}{b} \right) z_6 + \frac{R_p}{L_p} \left(1 + \frac{M}{b} \right) z_5 + \frac{-MR_s L_p}{b} z_4 + \frac{-ML_p}{b} z_3 \\ \dot{z}_6 &= L_f \Phi_6(\Phi^{-1}(Z)) + L_g \Phi_6(\Phi^{-1}(Z))u = \frac{1}{C_p} x_1 - \frac{1}{C_p} x_3 \Big|_{x=\Phi^{-1}(Z)} \\ &= \frac{1}{C_p} z_7 - \frac{1}{C_p} z_5 \end{aligned}$$

$$\begin{aligned}\dot{z}_7 &= L_f \phi_7(\phi^{-1}(Z)) + L_g \phi_7(\phi^{-1}(Z))u = \frac{-R_m}{L_m} x_1 + \frac{-1}{L_m} x_2 \Big|_{x=\phi^{-1}(Z)} \\ &= \frac{-R_m}{L_m} z_7 + \frac{-1}{L_m} z_6\end{aligned}$$

The dynamics of the system in Z domain is given by,

$$\dot{z}_1 = z_2$$

$$\dot{z}_2 = \frac{1}{L_c C_d} z_3 - \frac{R_c}{L_c} z_2 - \frac{I_L}{L_c C_d} + \left(\frac{2}{L_c C_d} z_1\right)u$$

$$\dot{z}_3 = \frac{1}{C_s} z_4 - \frac{C_d}{C_s} z_2 - \frac{I_L}{C_s}$$

$$\dot{z}_4 = \frac{M}{b} z_6 + \frac{-R_p}{b} z_5 + \frac{-R_s L_p}{b} z_4 + \frac{-L_p}{b} z_3$$

$$\dot{z}_5 = \left(\frac{1}{L_p} + \frac{M^2}{b}\right) z_6 + \frac{R_p}{L_p} \left(1 + \frac{M}{b}\right) z_5 + \frac{-MR_s L_p}{b} z_4 + \frac{-ML_p}{b} z_3$$

$$\dot{z}_6 = \frac{1}{C_p} z_7 - \frac{1}{C_p} z_5$$

$$\dot{z}_7 = \frac{-R_m}{L_m} z_7 + \frac{-1}{L_m} z_6$$

The control law expressed in Z coordinate is

$$\begin{aligned}u &= \frac{-L_f^2 h(X)}{L_g L_f h(X)} + \frac{v}{L_g L_f h(X)} \Big|_{x=\phi^{-1}(Z)} \\ &= \frac{R_c x_6 - x_5 + L_c C_d v}{2x_7} \Big|_{x=\phi^{-1}(Z)} \\ &= \frac{R_c (C_d z_2 + I_L) - z_3 + L_c C_d v}{2z_1}\end{aligned}$$

Appendix II Matlab Script for Input Current Computation

```

% for three-phase three leg converter
sum_t=[];
An_n=[];
Hm_m=[];
ia=[]; ib=[]; ic=[];
phi=atan(w*L/R);
Sa=[]; Sb=[]; Sc=[];
idc=[];
delT=0.00001;
for t=0:delT:0.02
    Sa=[Sa 0];
    Sb=[Sb 0];
    Sc=[Sc 0];
    ia=[ia I*sin((w*t)-phi)];
    ib=[ib I*sin((w*t)-phi-(2*pi/3))];
    ic=[ic I*sin((w*t)-phi+(2*pi/3))];
end
for m=-500:1:500
    Hm_m=[Hm_m 0];
end

Hm=0; del0=del/4;alpha=0;
for ii=1:P
    del1ii=del0*(1+M*sin((alpha-del0)));
    del2ii=del0*(1+M*sin((alpha+del0)));

    for m=-500:1:500
        Hm_m1=[];
        if m~=0
            Hmm=((1i*-1)*(1/(2*pi*m)))*(exp(-1*i*m*del2ii)-exp(1*i*m*del1ii))*exp(-1*i*m*alpha);
            Hmm_a_t=[];Hmm_b_t=[];Hmm_c_t=[];
            for t=0:delT:0.02
                Hmm_a_t=[Hmm_a_t Hmm*exp((1*i*m*w*t))];
                Hmm_b_t=[Hmm_b_t Hmm*exp((1*i*m*((w*t)-2*pi/3)))]];
                Hmm_c_t=[Hmm_c_t Hmm*exp((1*i*m*((w*t)+2*pi/3)))]];
            end
            Sa=Sa+(Hmm_a_t);
            Sb=Sb+(Hmm_b_t);
            Sc=Sc+(Hmm_c_t);
            Hm_m1=[Hm_m1 Hmm];
        end
    end
    Hm=Hm+Hmm;
    alpha=alpha+(del);
    Hm_m=Hm_m+Hm_m1;
end
Sa=0.5+(Sa);
Sb=0.5+(Sb);
Sc=0.5+(Sc);

for t=1:numel(ia)
    idc =[idc abs(ia(t)*Sa(t) + ib(t)*Sb(t) + ic(t)*Sc(t))];
end
Fs=1/f;
L=numel(idc);

```

```

Y=fft(idc,110)/110;
figure; bar(abs(Y)) ;
figure;plot(abs(idc));
% for single-phase two leg H-bridge converter
sum_t=[];
An_n=[];
Hm_m=[];
ia=[]; ib=[]; ic=[];
phi=atan(w*L/R);
Sa=[]; Sb=[]; Sc=[];
idc=[];
delT=0.00001;
for t=0:delT:0.02
    Sa=[Sa 0];
    Sb=[Sb 0];
    Sc=[Sc 0];
    ia=[ia I*sin((w*t)-phi)];
    ib=[ib I*sin((w*t)-phi-(pi))];
end

Hm=0; del0=del/4;alpha=0;
for ii=1:P
    del1ii=del0*(1+M*sin((alpha-del0)));
    del2ii=del0*(1+M*sin((alpha+del0)));

    for m=-110:1:110
        if m~=0
            Hmm=((1i*-1)*(1/(2*pi*m)))*(exp(-1*i*m*del2ii)-exp(1*i*m*del1ii))*exp(-1*i*m*alpha));
            Hmm_a_t=[];Hm_b_t=[];Hm_c_t=[];
            for t=0:delT:0.02
                Hmm_a_t=[Hmm_a_t Hmm*exp((1*i*m*w*t))];
                Hm_b_t=[Hm_b_t Hmm*exp((1*i*m*((w*t)-pi)))]];
            end
            Sa=Sa+(Hmm_a_t);
            Sb=Sb+(Hm_b_t);
        end
    end
    Hm=Hm+Hmm;
    alpha=alpha+(del);
end
Sa=0.5+(Sa);
Sb=0.5+(Sb);
for t=1:numel(ia)
    idc =[idc (ia(t)*Sa(t) + ib(t)*Sb(t))];
end
Fs=1/delT;
L=numel(idc);
NFFT= 2^nextpow2(Fs/f);
Y=fft(idc,NFFT)/L;
figure; bar(abs(Y(1:110))) ;
figure;plot(abs(idc));

```