Integrated Battery Chargers for Electric Vehicle Traction Drives with Synchronous Machines

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Abstract

The transport sector is increasingly moving towards more electrification to benefit from enhanced reliability and energy efficiency. The automotive transport sector, which includes passenger and service vehicles, has also jumped on to the bandwagon of electrification. The wave of electrification brings with it challenges that must be resolved to realize its gains. One such challenge is that of ease and speed of charging for the large batteries that are required by electric vehicles to ensure longer range on a single charge. Fast chargers for electric vehicles are often heavy and large in volume because of their high-power rating and therefore cannot be placed on the vehicle. Chargers that can be placed on the vehicle are not fast enough since they need to be smaller in size and weight to be placed on board. The thesis focuses on chargers that provide a compromise between the bulkiness and the power rating of the charger.

The thesis proposes to integrate battery chargers into the existing traction drives of electric vehicles. The integration is achieved by using the traction motor windings as the filter inductances to interface the charger to the grid by using minor reconfiguration. Windings that are excited with a current produce a pulsating torque on the rotor shaft, the nature of which varies with the type of machine used. Since most traction drives use synchronous machines, the thesis proposes torque cancellation strategies that can be used for the three most common types of synchronous machines used such as the surface mounted permanent magnet synchronous machine, the interior permanent magnet machine, and the synchronous reluctance machine.

Since the winding inductances of synchronous machines depend on the rotor position, they form a set of unequal and unbalanced filter inductance for the charger. Power balance with linear quadratic regulator control strategy that allows the use of such filter inductances is proposed. A modulation strategy to reduce the circulating current in the parallel connected converters of the integrated battery charger is also proposed. Furthermore, the analysis of the machine during charging shows that the charger power rating cannot be designed to have the same rating as the traction motor windings and must be limited to less than that to avoid stator saturation. Validation of the proposed strategies is done by using co-simulation of finite element analysis for machines and time domain transient simulator for the power electronics and control first, then validated on hardware setups.
Résumé

Pour bénéficier d’une efficacité énergétique et d’une fiabilité accrues, le secteur des transports évolue progressivement vers une plus grande électrification. Notamment le secteur de l’automobile, incluant les véhicules de tourisme et de service, qui a pris le train en marche de l’électrification. Pour profiter des avantages de l’électrification, de nombreux défis doivent être résolus. Parmi ces défis on trouve celui de rendre facile et rapide la charge pour les grandes batteries, qui sont nécessaires pour les véhicules électriques afin d’assurer une plus grande autonomie avec une seule charge. Les chargeurs rapides pour véhicules électriques sont souvent lourds et prennent beaucoup de place en raison de leur puissance nominale élevée et ne peuvent donc pas être placés dans le véhicule. Les chargeurs qui respectent les exigences de poids et de volume ne sont pas assez rapides. La thèse se concentre sur les chargeurs qui offrent un compromis entre l’encombrement et la puissance nominale.

La thèse propose d’intégrer des chargeurs de batterie dans les systèmes de traction existants des véhicules électriques. L’intégration est réalisée en utilisant les enroulements du moteur de traction comme une inductance de filtre pour connecter le chargeur au réseau en faisant une reconfiguration mineure. L’excitation des enroulements par un courant produit un couple sur l’arbre du rotor, dont la nature varie en fonction du type de machine utilisé. Comme la plupart, des systèmes de traction utilisent des machines synchrones, la thèse propose des stratégies d’annulation de couple pour les trois types de machines les plus courantes, la machine à aimants permanents en surface, la machine à aimants permanents encastrées et la machine synchrone à réluctance. Puisque les inductances d’enroulement des machines synchrones dépendent de la position du rotor, elles forment un ensemble d’inductance de filtre inégal et déséquilibrée pour le chargeur. Un équilibrage de puissance avec une stratégie de contrôle de régulateur quadratique linéaire qui permet l’utilisation de telles inductances de filtre est également proposée. Une stratégie de modulation pour réduire le courant circulant dans les convertisseurs connectés en parallèle du chargeur de batterie intégré est également proposée. De plus, l’analyse de la machine pendant le chargement montre que la puissance nominale du chargeur ne peut pas être dimensionnée de la même manière que les enroulements du moteur de traction et doit être inférieure à celle-ci pour éviter la saturation du stator. La validation des stratégies proposées est réalisée en utilisant d’abord la co-simulation de l’analyse par éléments finis pour les machines et la simulation temporelle transitoire pour l’électronique de puissance et le contrôle. Enfin, une validation sur un banc d’essai a été effectuée.
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Primarily I would express my gratitude towards the Almighty God who made it all possible.

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“That man can have nothing but what he strives for; that (the fruit of) his striving will soon come in sight: Then will he be rewarded with a reward complete.”

An-Najm 53:39-41 – The Quran
Preface

In the interest of full disclosure, five publications resulted from this thesis including three conference papers and two journal articles (as listed below). The candidate was the primary contributor in defining the problem, proposing the solution, planning and executing the software and hardware experiments, analyzing the data, and in writing the resulting articles and this thesis. The contributions of each co-author are also listed.

A) Journals


Mr. Diego Mascarella and Dr. Loncheng Tan contributed in terms of providing editorial and technical comments on the candidates work and logistical support in conducting hardware and software experiments. Prof. Geza Joos contributed in supervising the primary author and helping him shape and refine the contributions of the paper.


Mr. Diego Mascarella and Dr. Loncheng Tan from McGill University contributed in terms of providing editorial and technical comments on the candidates work and logistical support in conducting hardware and software experiments. Prof. Geza Joos contributed in supervising the primary author and helping him shape and refine the contributions of the paper.

B) Conferences


Mr. Diego Mascarella from McGill University contributed in terms of providing editorial and technical comments on the candidates work and logistical support in conducting hardware and software experiments. Prof. Geza Joos from McGill University contributed in supervising the primary author and helping him shape and refine the contributions of the paper. Mr. Tony Coulombe and Mr. Jean-Marc Cyr from TM4 Inc. have provided logistical support for experimentation at the industrial partners’ (of the project) facilities.

Mr. Diego Mascarella from McGill University contributed in terms of providing editorial and technical comments on the candidates work and logistical support in conducting hardware and software experiments. Prof. Geza Joos from McGill University contributed in supervising the primary author and helping him shape and refine the contributions of the paper. Prof. Gerry Moschopoulos from Western University has provided editorial guidance to the candidate in shaping the paper on circulating current mitigation and formulating the proposal for the thesis.


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<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>BEV</td>
<td>Battery Electric Vehicle</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DER</td>
<td>Distributed Energy Resource</td>
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<tr>
<td>DPWM</td>
<td>Discontinuous Pulse Width Modulation</td>
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<tr>
<td>EV</td>
<td>Electric Vehicle</td>
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<tr>
<td>EMTP</td>
<td>Electro-Magnetic Transient Program</td>
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<tr>
<td>FEA</td>
<td>Finite Element Analysis</td>
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<tr>
<td>HEV</td>
<td>Hybrid Electric Vehicle</td>
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<tr>
<td>ICE</td>
<td>Internal Combustion Engine</td>
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<tr>
<td>IBC</td>
<td>Integrated Battery Charger</td>
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<tr>
<td>IPM/IPMSM</td>
<td>Interior Permanent Magnet Synchronous Machine</td>
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<tr>
<td>LQR</td>
<td>Linear Quadratic Regulator</td>
</tr>
<tr>
<td>PHEV</td>
<td>Plugin Hybrid Electric Vehicle</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
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<tr>
<td>RCP</td>
<td>Rapid Control Prototyping</td>
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<tr>
<td>SAE</td>
<td>Society of Automotive Engineers</td>
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<tr>
<td>SPM/SPMSM</td>
<td>Surface mounted Permanent Magnet Synchronous Machine</td>
</tr>
<tr>
<td>SRM</td>
<td>Synchronous Reluctance Machine</td>
</tr>
<tr>
<td>V2G</td>
<td>Vehicle to Grid</td>
</tr>
<tr>
<td>VRM</td>
<td>Variable Reluctance Machine</td>
</tr>
<tr>
<td>VSC</td>
<td>Voltage Source Converter</td>
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<td>ZVC</td>
<td>Zero-Vector Crossing</td>
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List of Symbols

\( \alpha \)  Displacement between the two phase-sets of a six-phase synchronous machine

\( B \)  Flux density

\( C \)  Capacitance

\( D_p \)  Demagnetization proximity

\( f \)  Force density

\( \phi \)  Angular position in the air gap

\( I \)  Current

\( J \)  Rotor inertia

\( L \)  Inductance

\( \lambda \)  Flux

\( \mu_0 \)  Permeability of free space

\( N \)  Neutral

\( P \)  Active power

\( p \)  Number of poles

\( Q \)  Reactive power

\( S_x \)  Switching state of the VSC arm connected to phase \( x \)

\( t \)  Time in seconds

\( \tau \)  Torque

\( \theta_g \)  Grid phase angle

\( \mathbf{u}_{mag} \)  Unit vector representing the direction of the magnet flux

\( V \)  Voltage

\( Z \)  Impedance

\( \omega_g \)  Grid frequency

\( \omega_r \)  Rotor speed
1 Introduction

1.1 Background

Recently electric vehicles (EV) have gained much attention in the wake of the volatility of oil prices that affect the use of internal combustion engine (ICE) based vehicles. EVs, when compared to their ICE counterparts, do not require fossil fuels and are therefore more climate-friendly. EVs are categorized into further two categories: hybrid electric vehicles (HEV), and battery electric vehicles (BEV). HEVs have a hybrid source of energy, i.e., the fuel and a battery both. These vehicles utilize the battery as a buffer that is discharged at low efficiency operating points of the ICE and is charged back by the ICE at its high efficiency operating points. Some variants of HEVs have a large rechargeable battery for an all-electric mode of operation that extends the range of the vehicle [1]. The batteries for electric vehicles mainly consist of either flow batteries (including fuel cells) or solid-state batteries. Flow batteries can be almost instantaneously recharged by replacing its electrolyte. However, flow batteries have a problem of safe handling of the electrolytes which has been a major hurdle in its market acceptance. The solid-state batteries on the other hand are an accepted technology in the industry (esp. the lithium ion batteries). Solid-state batteries store their energy in the electrodes that gets discharged as the battery is used. To replenish this energy, the batteries must be provided by an external electric source. Such batteries with a higher capacity require a dedicated charging interface that could be plugged into the electric power supply. Hence such vehicles are termed plugin hybrid electric vehicles (PHEV). BEVs on the other hand, only have a single source of energy, i.e., the battery. BEVs require a dedicated charging interface that could be plugged into the electric power supply for recharge as well. Batteries for PHEVs and BEVs are designed to have as much capacity as possible to provide a lengthier all-electric range for the vehicle. However, since the size and weight of the battery are proportional to its capacity, a tradeoff must be made between the size and weight, and the range that the vehicle can provide [2].

PHEVs and BEVs require a charger interface to be charged from the electric power supply. There are two types of battery chargers for EVs according to the modes in which the power is transmitted to the battery: wireless; and conductive (wired) charging as shown in Figure 1-1. Wireless chargers can be further divided into inductive chargers and capacitive chargers [3]. The
transmitter on the charger and the receiver on the vehicle are coupled electromagnetically for inductive chargers and electrostatically for capacitive chargers. Wireless chargers for electric vehicles provide a more convenient and automated way of charging. Although the theoretical efficiency of the wireless chargers can be very high, the efficiency of practical charger designs is low because it is highly sensitive to the alignment of, and the distance between the power transmitter and receiver [4].

Conductive chargers do not have the efficiency issues of wirelessly coupled chargers and are therefore more widely accepted in the automotive industry. Conductive chargers can be classified based on their power ratings and by their placement on or off the vehicle. The Society of Automotive Engineers (SAE) has categorized EV chargers based on their power rating in J1772 [5] as shown in Table 1-1. Alongside the classification shown, there is another class of chargers labeled as AC Level 3 chargers that are rated at a power greater than 19.2kVA and can have a single-phase or three-phase interface [6, 7]. Regarding size, AC Level 1 chargers are small enough to be put on the vehicle and are thus called on-board chargers [8]. The power ratings of AC Level 2 chargers allow certain components to be placed on the vehicle while requires the rest to be placed outside the vehicle as dedicated charging hardware. Due to their power ratings, DC Level 1 and Level 2 chargers are large, heavy and require a direct connection the battery DC terminals to charge it [9-11]. Due to their large size, weight and bulkiness, these chargers along with (partial) AC
Level 2 chargers, cannot be placed on the vehicle and are thus called off-board chargers. On-board chargers provide flexibility in terms of opportunity charging, i.e., the user can plug it in any outlet with the suitable socket, but takes a long time to charge. Off-board chargers, on the other hand, have dedicated infrastructure and can be found in a gas station type commercial installation and can charge the batteries sooner when compared to on-board chargers [12]. These chargers, however, do not provide flexibility and cost more than on-board chargers.

Integrated battery chargers (IBC) provide a good compromise between cost, flexibility and the time it takes to charge the battery. IBCs reduce the number of dedicated components for charging by making use of the already existing components of the traction drive in EVs to contribute in the charger circuit. Since charging and traction are mutually exclusive functions, i.e., the EV traction drive is not used when the EV battery is being charged, and when the EV is being driven its battery is not being charged. Therefore, the already existing components of the traction drive can be repurposed for charging. A typical EV drivetrain with a dedicated on-board charger, a dedicated

![figure](image.png)

Figure 1-2 Types of EV battery chargers (a) on-board charger; (b) off-board charger; and (c) integrated battery charger.
off-board charger, and an IBC is shown in Figure 1-2 (a), (b) and (c) respectively. Each figure shows the components of the EV drivetrain that are on the vehicle and those that are off the vehicle. The traction drive in the figures can be a one-stage DC-AC converter or a two-stage converter that cascades a DC-DC and a DC-AC converter. Due to the high variation of the battery voltage due to its state of charge variation, a two-stage solution is preferred. A bidirectional interface with the grid provides an opportunity of value addition to the charger such that the charger can exchange active and reactive power from the grid in either direction. This feature enables the charger to participate in the vehicle to grid (V2G) ancillary services [13] such as frequency regulation [14], load and generation smoothing [15], reactive power [16] and voltage support [17].

1.2 Literature Review

This section reviews the existing integrated battery chargers and their topologies.

1.2.1 Integrated battery chargers

There are various approaches towards integrating the charger with the existing traction drive ranging from partial to complete integration. Partial integration means the IBC does not utilize the traction inverter or the motor windings and additional components. IBCs that only utilized the traction inverter (or parts of it) were proposed in [18-26]. The IBC presented in [18] used a single leg of the traction inverter to form a fly-back converter fed by an uncontrolled rectifier. The topologies in [19-22] integrated a single-phase battery charger into the DC-DC stage of the traction drive. The chargers proposed in [23-25] were based on unconventional traction drives. All these
chargers required additional semiconductor switches in the conduction path of the traction mode of operation which adds losses to the traction mode of operation. Therefore, much effective integration of the battery charger is achieved by utilizing the existing topology such that both the traction inverter and the traction motor are a part of the IBC and no major additional components, like semiconductor switches or passive components, are required.

1.2.2 Single-phase integrated battery chargers

The IBCs presented in [27-30] utilized both, the machine windings and the traction inverter. However, these chargers required an additional rectifier at the input, and the IBC in [30] required access to the neutral point of the traction motor. The rectifier not only results in added components to the topology but also renders the charger capable of only operating with unity power factor in the charging direction. Single-phase IBC topologies were presented in [31, 32] that utilized the windings and drive of a switched reluctance traction motor without requiring additional components. Single-phase charging topologies based on two motor drives are proposed in [33-36], based on a traction drive with four in-wheel traction motors in [37], and on a drive with multiphase machines in [38]. Single-phase IBC in [39, 40] was proposed to have two of the phase windings connected in series to get the best inductance, while the rotor position was proposed to be aligned to the minimum torque position while charging. Although these topologies did not require additional components, their single-phase interfaces limit the maximum power ratings the chargers could be designed for and therefore result in slow charging of the battery.

1.2.3 Three-phase integrated battery chargers

Three-phase interfaced IBCs provide an opportunity to form high power, and therefore faster, chargers. However, when three-phase machine windings are excited with high enough three-phase currents, they can produce a significant torque on the shaft of the motor. The torque produced will depend on the motor technology. In general, the three-phase chargers are further classified into IBCs based on three-phase motor traction drives and IBCs with multiphase motor traction drives.

1.2.3.1 Three-phase high power integrated battery chargers with three-phase machines

A three-phase IBC based on a three-phase surface mounted PM traction drive was proposed in [41], based on interior PM traction drives in [42], based on interior PM traction drives with open-ended windings in [43] and based on interior PM traction drives with damper windings in [44, 45].
The torque generated on the drive shaft was not addressed for both, and therefore the only way to limit the torque on the shaft is to limit the current in the motor windings, resulting in low power charging. Most of these designs did not address the problem of the torque produced on the shaft of the motor while charging and the solution suggested was to lock the rotor mechanically. In [44, 45] the authors proposed designing the damper windings so that the torque produced is minimal, despite that, applying brakes was suggested in case the torque was produced. There are two consequences to the suggested solutions, 1) the charger is limited to low current charging to limit the torque developed on the shaft; 2) if the charger operates with a higher current, the torque produced will result in unwanted noise, vibration, and deterioration of the vehicle’s mechanical drive train. The torque on the shaft produces a small displacement since the rotor inertia for the motors used in the electric vehicular application is much higher during the charging mode (with its parking brake activated) than the driving mode. This displacement may not result in moving the vehicle; however, it will result in unwanted vibrations and noise on the powertrain shaft, which will result in the lifetime reduction of the electric vehicle and require enhanced suspension design to mitigate mechanically.

Another IBC based on an interior permanent magnet traction motor with open-ended windings was proposed in [46, 47]. The torque cancellation for the machine was achieved by connecting the three-phase grid interface to the midpoints of the machine windings. The requirement of the windings mid-points make requires a specially designed machine for the charger. IBCs based on a three-phase machine that required an additional three-phase current source rectifier for the grid interface were proposed in [43, 48]. Although this topology resulted in a bidirectional charger, it required a full rated current source rectifier adding to the weight and size of the charger. Torque generation was not an issue for this charger as the machine windings were used as part of a DC-DC converter carrying equal and DC currents, locking the rotor in its position. A three-phase IBC for three-phase machines was proposed in [49]. This charger required a three-phase AC-DC buck converter that only allows unity power factor rectification. Therefore, the proposed charger requires an additional rectifier, resulting in increased cost and weight, and is not capable of participating in V2G schemes. An isolated three-phase IBC based on a specially designed internal permanent magnet machine was presented in [50-52]. Isolation was achieved by winding reconfiguration, where the series windings of each phase are disconnected and arranged as two isolated sets of windings. The torque on the shaft was addressed by allowing the motor to rotate.
while charging such that one set of the isolated windings maintained the speed of the rotor synchronized to the grid acting as a motor while the other set operated as generator windings charging the battery. The rotation caused during the charging operation, however, required an additional mechanical clutch on the drivetrain and the rotation itself added losses to the charging process.

1.2.3.2 Three-phase high power integrated battery chargers with multiphase machines

Recently, multiphase machine drives have gained the interest of industry for EV drivetrains [53]. Multiphase machines provide additional degrees of freedom in the converter control [54]. Since electric vehicle propulsion requires high torque density, efficiency and fault tolerance these additional degrees of freedom can be used to achieve added value for the traction drive in the traction mode of operation. These additional degrees of freedom not only provide added value in the traction mode of operation but also in integrating the battery charger to the drive [55].

IBCs that explicitly addressed the torque on the shaft of the traction motor was addressed in [56] for a five-phase machine. However, five-phase machines are rarely used in automotive applications. A charger based on a nine-phase machine was proposed in [57], on a six-phase machine with galvanic isolation in [58], and without galvanic in [59]. A torque cancelation strategy for all these machines was proposed such that the currents in the torque producing plane were either both controlled to be zero or one of them controlled to be zero. In both cases, the product of these currents results in a zero on the torque plane. The torque cancellation strategy proposed for these IBCs effectively eliminates the torque for induction motors. The strategy is effective because for induction machines the rotor is excited by the stator current itself to produce torque and is not rotor position dependent. The same is not true for synchronous machines. When applied to synchronous machines (PM and reluctance) the proposed torque cancelation strategy will be highly sensitive to the rotor position.

An IBC was proposed for a six-phase IPM machine in [60] for a nine-slot eight pole winding configuration. The torque was addressed by reconfiguring the machine into a rotating transformer such that one of the windings rotates the rotor while the other windings operate as a set of generator windings. The rotation during the charging operation requires an additional mechanical clutch on the drivetrain and added losses to the charging process.
1.2.3.3 Control challenges for three-phase integrated battery chargers with multiphase synchronous machines

1.2.3.3.1 Rotor position dependent inductances

When the windings of a synchronous motor are used as the filter inductances of a three-phase battery charger, the winding inductances are not equal for each phase. The unequal inductances cause an unbalanced power being drawn from the grid despite the grid voltage and current drawn being balanced. This unbalanced power creates a low frequency (second harmonic) oscillation on the DC link, which may require a larger DC link capacitor for the traction drive. The power balancing schemes proposed in the literature either require an active power filter [61-66] or a four-leg converter [67, 68] to compensate for the power unbalance. Both solutions are not feasible to be implemented in a traction drive, since the former requires an additional active power filter, while the latter requires a four-legged converter, which are rarely used in traction drives. Therefore, a power balancing strategy that is capable of balancing the power for the three-phases of the charger is required, such that the DC link oscillations are also reduced.

1.2.3.3.2 Circulating Current

Most IBC topologies presented are essentially that of two-level converters connected in parallel with a common DC link and the same AC grid [69]. Such converters have a circulating current path between the parallel connected converters [70]. This path is excited by the non-zero difference in instantaneous voltage between the converters. Circulating current increases the stresses on the power converter components and the system conduction losses [71] and therefore should be reduced. The magnitude of the circulating current depends on the control of the converter [72] and the modulation techniques used [73]. In the literature circulating current has been controlled in three distinct ways or their combinations: 1) using circulating current filters [74]; 2) using circulating current suppression controls [75], and 3) special modulation techniques [76].

During charging operation, IBCs can be operated with various PWM schemes, including discontinuous PWM (DPWM) [77]. With such schemes, the phase leg carrying the peak current is not switched for a period corresponding to 60° of the fundamental cycle, thus resulting in reduced switching instances per fundamental cycle and reduced switching losses by a factor of half (theoretically). The circulating current problem is aggravated when the converters are controlled using interleaved DPWM schemes because of their fundamental nature. Thus, several ways to reduce this circulating current have also been proposed in the literature.
Circulating current at high frequency was reduced in [78-80] using a common mode inductance at each converter’s output. Consequently, the size and weight of the converter are increased as bulky and heavy inductances are required. Circulating current was also reduced by modified PWM techniques in [81-84]. Several were based on space vector PWM [81, 82] while others used carrier-based PWM with third harmonic injection [83] or selective harmonic elimination [84]. These techniques cannot be used with DPWM since the modification is in the zero vector is not explicitly considered, thus rendering the opportunity of reducing switching losses as unavailable. DPWM techniques with interleaving were presented in [73] and [85]. The proposed technique in [73] increased the switching instances per line cycle to reduce the circulating current which, in turn, defeated the purpose of using DPWM. The technique in [85] added fewer switching instances per switching cycle, retaining the benefits of DPWM. However, it required accurate pulse adjustment strategies for the strategy to work.

A modulation strategy that reduces the circulating current between the IBC converters is thus required such that the converters can also be operated with DPWM strategies, allowing more efficient operation.

1.3 Problem Definition

Synchronous machines including surface mounted PM, interior mounted PM, and synchronous reluctance machines are increasingly used for traction applications due to their high power and torque density, high efficiency and ease of operation. Progressively multiphase, specially six-phase, machines have been adopted for light-duty, medium-duty, and heavy-duty automotive traction applications. Integrating a battery charger in the drives of such traction applications have a potential of providing a high-power charging opportunity on top of providing the advantages of modularity, redundancy and fault tolerance. However, integrating a battery charger in such traction drives is not straightforward, as the torque production is highly dependent on the rotor position and the rotor excitation is permanent (in the case of permanent magnet machines).

The literature review leads the following unresolved problems that this thesis attempts to resolve:

1) Three-phase integrated battery chargers based on traction drives that use three-phase and multiphase synchronous machines produce a torque on the shaft when used as chargers with
no direct way of eliminating the pulsating torque, especially when the rotor position is not aligned with the stator magnetic axis.

2) Synchronous machines have unequal and unbalanced inductances that present a challenge when used as filter inductances connected to interface the charger with the grid. Furthermore, the common inductance provided by the machine windings are conventionally small, allowing a circulating current to flow between legs if connected in parallel.

3) The impact of charging on the multiphase synchronous machine, the windings of which are used as the filter inductances for the integrated battery charger. Also, currently there are no guidelines for the design of integrated battery chargers based on traction drives with synchronous machines.

1.3.1 Research objective and methodology

The research objective as deduced from the problem formulation are listed below:

1) To analyze the nature of torque generated on the shaft of synchronous machines when used as part of integrated battery chargers and to identify a torque cancellation strategy that applies to surface mounted permanent magnet synchronous machines, interior permanent magnet synchronous machine, and the synchronous reluctance machine with three-phase and six-phase windings.

2) To identify integrated battery charger topologies based on traction drive topologies along with the control algorithms that allow effective implementation of the identified torque control strategies. The identified control algorithms should specifically address the adverse effects of the imbalance in the machine windings inductance.

3) To establish the impact of charging on the synchronous machine used as an integrated battery charger with the torque cancelation strategy. Furthermore, to determine design recommendations for the integrated battery charger based on impacts on the synchronous machines.

1.3.2 Methods and tools

The tools used to analyze the synchronous machines was based on analytical models and finite element models. Various finite element models were developed to validate the derived analytical torque models. For further analysis of the system, several simulation and hardware setups were used as mentioned below:
1.3.2.1 *Software tools*

The software tools used are:

1) *Machine design tool* – A synchronous machine design tool was used to develop various machine designs that included the previously mentioned three synchronous machine technologies. A commercial tool that allows creation of synchronous machine designs based on the required ratings and technology was used to generate the required machine designs.

2) *Finite element analysis (FEA) tool* – An FEA tool was used to analyze and co-simulate the various machine designs developed.

3) *Electro-magnetic transient program (EMTP)* – A time-domain simulation platform was required to study and validate the power balancing with LQRI control and circulating current modulation algorithms developed. A co-simulation of the machine models in the FEA tool and the power electronics in the EMTP simulator was also used to generate system results.

1.3.2.2 *Hardware tools*

Hardware tools were also used to analyze and validate the developed strategies and provided as follows:

1) *Synchronous machines* – Two types of motors synchronous motors by commercial manufacturers were used to validate the torque cancellation strategies. The first type included a pair of three-phase SPMSM, while the second type included a six-phase IPMSM.

2) *Power electronic converter* – To implement and validate the control strategies with the machines mentioned above, two-level converter stacks were used for each three-phase set of the machine windings.

3) *Digital controller* – To implement the control algorithms, including the modulated pulses for the power electronics, a digital controller or a rapid control prototyping (RCP) tool is required. The controller used for RCP had 16 analog input and 16 analog output channels, including 16 dedicated PWM channels as well.

**1.4 Claims of Originality**

1) Torque cancellation strategies for synchronous machines used in integrated battery chargers, where, the synchronous machines may include surface mounted permanent magnet, interior permanent magnet, and synchronous reluctance machines.
2) Topologies for the implementation of the integrated battery chargers based on two three-phase or six-phase synchronous machines along with the power balancing control with the linear quadratic regulator to control the currents in the unequal machine winding inductances such that the DC link ripple is minimized. Additionally, a circulating current reduction modulation strategy for the parallel connected converters forming an integrated battery charger with low winding inductance.

3) Identification of the impacts on the synchronous machine, the windings of which are used as the filter inductance of an integrated battery charger with torque cancellation, regarding saturation, internal force generation. The identified impacts provide insights that help in the design of integrated battery chargers based on traction drives with synchronous machines.

1.5 Thesis Outline

The thesis is organized into five chapters. The details of the chapters are given as follows:

**Chapter 2: Torque Control of Integrated Battery Chargers with Synchronous Machines**

This chapter presents the usage of three-phase and six-phase machines as the filter inductances of a charger. The total torque generated on the shaft of the three-phase and six-phase synchronous machines when used as a charger is empirically analyzed and validated against finite element models of those machines. A torque cancelation strategy for two three-phase and six-phase machines is then proposed and explained in detail. The effectiveness of the strategy is then demonstrated using finite element analysis. The three-phase and six-phase machines analyzed include surface mounted permanent magnet synchronous machines (SMPMSM), internal permanent magnet synchronous machines (IPMSM) and synchronous reluctance machines (SRM). The torque cancelation sensitivity to errors in the position sensing and the stator saturation are also analyzed for a six-phase machine.

**Chapter 3: Topologies and Control of Integrated Battery Chargers**

This chapter presents topologies for the practical implementation of the torque cancelation strategies presented in the previous chapter. The modeling of the machines as a part of the charger is presented. When the charger is used with synchronous machines that have winding inductances that depend on the rotor position, an unbalanced power is absorbed from the grid, resulting in a second harmonic current on the DC link. A power balancing control is presented to reduce it. The
topologies presented result in a path made available for the circulating current to flow within the windings. A modulation strategy that reduces the circulating current is also proposed and validated. Overall the chapter presents solutions to the issues that arise when using three-phase and six-phase synchronous machines as the filter inductances of integrated battery chargers.

Chapter 4: Integrated Battery Charger Analysis and Design –

This chapter presents an analysis of the machine the windings of which are used as the filter inductance of the integrated battery charger. The analysis in the chapter is done for different kinds of windings of a six-phase machine – namely full-pitched and short-pitched windings with 0°, 30° and 60° displacement between the two three-phase sets. The effect of charging on the demagnetization of the magnets is also studied for the different kinds of windings mentioned. Recommendations on the windings selection and the rating definition of the charger are presented as a result of the analysis.

Chapter 5: Summary and Conclusion –

This chapter highlighted the results of the thesis and identified the contributions as a result. The implications and the applications of the results are also discussed. The future directions of research that arise from work are then presented.
2 Torque Control of Integrated Battery Chargers with Synchronous Machines

This chapter first introduces the problem of torque production on the shaft while the battery is being charged for IBCs based on conventional three-phase synchronous machines and six-phase synchronous machines. Torque cancelation strategies are then proposed for drives with two conventional three-phase synchronous machines with a common shaft and drives with a single six-phase synchronous machine for the battery charging process. The applicability of the proposed strategy is demonstrated on the three kinds of synchronous machines (SMPM, IPM, and SRM) for both three-phase and six-phase machine based IBCs. The strategy is validated using analytical machine models, finite element analysis (FEA) machine models and hardware experiments.

2.1 Integrated Battery Chargers Based on Three-phase Synchronous Machines

The modeling conventions of the three-phase synchronous machine are presented in Appendix B. The following sections present the torque generation during the charging mode of operation of an integrated battery charger.

2.1.1 Torque in the charging mode of three-phase synchronous machines

The d-q model of the machine has an underlying assumption that the rotor rotates with an angular speed of \( \omega_r \) rad/s and therefore the transformation converts the stator voltages and currents into DC quantities. This assumption however, is not valid in the case of an IBC because the rotor is stationary and yet the stator currents are sinusoidal quantities synchronized to the grid voltage. Therefore, the d-q axes currents after transformation are given by

\[
\begin{align*}
    i_{ds} &= I_s \cos(\omega_s t + \theta_r + \theta_g) \\
    i_{qs} &= I_s \sin(\omega_s t + \theta_r + \theta_g)
\end{align*}
\]

where \( I_s \) is the peak of the three-phase currents, \( \omega_s \) is the frequency of the stator currents and \( \theta_g \) is the angle between the grid voltage and the grid current, and \( \theta_r \) is the angle between the rotor and stator d-axes.
The currents given by (2-1) can be used to calculate the torque on the shaft of a three-phase synchronous machine with its windings excited by the grid current. The following sections model the torque production for each kind of three-phase machine.

2.1.1.1 *Surface mounted permanent magnet synchronous machines*

For SPMSMs the total torque on the shaft is given by (B-5) with the fact that for SPMSMs $L_{ds} \approx L_{qs}$, we get

$$\tau_{em} = \frac{3}{2} \cdot p \lambda_{pm} I_s$$

(2-2)

Substituting (2-1) in (2-2) provides us:

$$\tau_{em} = \frac{3}{2} \cdot p \lambda_{pm} I_s \sin(\omega_s t + \theta_r + \theta_g)$$

(2-3)

Equation (2-3) shows that the torque generated on the rotor of a three-phase SPMSM with balanced sinusoidal currents in its stator windings will have a sinusoidal nature and will have a frequency same as that of the grid. From the mechanical model of the machine the change in the displacement will also be a sinusoid having the grid current frequency and an amplitude that depends on the rotor inertia $J$.

A comparison between the torque developed on the rotor for a 37kW machine calculated using (2-3) and obtained by an FEA model of the same machine (all the machine parameters are presented in Table A-1) with 80A current through its stator windings is shown in Figure 2-1. The result shown verifies the empirical prediction of the model used. The FEA results were calculated at $T=20$ °C.

2.1.1.2 *Interior permanent magnet synchronous machines*

In the case of an IPMSM, the torque produced on the shaft will be given by (B-5). Substituting the currents from (2-1) we get:

$$\tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm} I_s \sin(\omega_s t + \theta_r + \theta_g) - \frac{1}{2} (L_{qs} - L_{ds}) I_s^2 \sin(2\omega_s t + 2\theta_r + 2\theta_g) \right]$$

(2-4)
Equation (2-4) shows that the torque generated on the rotor of a three-phase IPMSM with balanced sinusoidal currents in its stator windings will consist of two components. The first component will be a sinusoid having the same frequency as of the current that excites its windings, and the second component will have twice the frequency of the current. From the mechanical model of the machine the change in rotor displacement will also be sinusoidal and will similarly consist of the fundamental and its second harmonic components. The amplitude of the displacement will depend on the rotor inertia $J$.

A comparison between the torque developed on the rotor for a 30kW machine calculated using (2-4) and obtained by an FEA model of the same machine (all the machine parameters are presented in Appendix A) with 80A current through its stator windings is shown in Figure 2-2. The result shown verifies the empirical prediction of the model used. The FEA results were calculated at $T=20$ °C.

2.1.1.3 **Synchronous reluctance machines**

In the case of an SRM, the torque produced on the shaft will be given by (B-5) with the fact that $\lambda_{pm} = 0$. Substituting the currents from (2-1) we get:
\[ \tau_{em} = -\frac{3}{2} \cdot p \left[ \frac{1}{2} (L_{qs} - L_{ds})I_s^2 \sin(2\omega_s t + 2\theta_r + 2\theta_g) \right] \] (2-5)

Equation (2-5) shows that the torque generated on the rotor of a three-phase SRM with balanced sinusoidal currents in its stator windings will have twice the frequency of the current exciting it. From the mechanical model of the machine the change in rotor displacement will also be sinusoidal having twice the grid frequency. The displacement amplitude will depend on the rotor inertia \( J \).

A comparison between the torque developed on the rotor for a 30kW machine calculated using (2-5) and obtained by an FEA model of the same machine (all the machine parameters are presented in Appendix A) with 80A current through its stator windings is shown in Figure 2-3. The result shown verifies the empirical prediction of the model used. The FEA results were calculated at \( T=20 \, ^\circ\text{C} \).
2.2 Integrated Battery Chargers Based on Six-phase Synchronous Machines

The modeling conventions of the six-phase synchronous machine are presented in Appendix B. The following sections present the torque generation during the charging mode of operation of an integrated battery charger.

2.2.1 Torque in the charging mode of six-phase synchronous machines

Like the three-phase machine d-q model, the six-phase d-q model also has an underlying assumption that the rotor rotates with an angular speed of \( \omega_r \) rad/s and therefore the transformation converts the stator voltages and currents into DC quantities. This assumption however, is not valid in the case of an IBC because the rotor is stationary and yet stator currents are sinusoidal quantities synchronized to the grid voltage. Therefore, the d-q axes currents after transformation are given by

\[
i_{d1} = I_{s1} \cos(\omega_{s1} t + \theta_{r1} + \theta_{g1})
\]  

(2-6)
\[ i_{q1} = I_{s1} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1}) \]
\[ i_{d2} = I_{s2} \cos(\omega_{s2}t + \theta_{r2} + \theta_{g2}) \]
\[ i_{q2} = I_{s2} \sin(\omega_{s2}t + \theta_{r2} + \theta_{g2}) \]  \hspace{1cm} (2-7)

where \( I_{s1} \) and \( I_{s2} \) are the peak of the three-phase currents, \( \omega_{s1} \) and \( \omega_{s2} \) are the frequency of the three-phase currents, \( \theta_{g1} \) and \( \theta_{g2} \) is the angle between the grid voltage and the grid currents, and \( \theta_{r1} \) and \( \theta_{r2} \) are the angles between the rotor and stator d-axes, of the phase-sets \( ABC \) and \( DEF \) respectively.

The currents given by (2-6) and (2-7) can be used to calculate the torque on the shaft of a six-phase synchronous machine with its windings excited by the grid current. The following sections model the torque production for each kind of six-phase machines.

2.2.1.1 Surface mounted permanent magnet synchronous machines

For six-phase SPMSMs the total torque on the shaft is given by (B-13) with the fact that for SMPMSs \( L_{d1} \approx L_{q1} \) and \( L_{d2} \approx L_{q2} \) due to the nature of the transformation, we get

\[ \tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm1}I_{s1} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1}) + \lambda_{pm2}L_{d2} \sin(\omega_{s2}t + \theta_{r2} + \theta_{g2}) \right. \]
\[ \left. + (L_{md} - L_{mq})I_{s1}L_{s2} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1} + \omega_{s2}t + \theta_{r2} + \theta_{g2}) \right] \]  \hspace{1cm} (2-8)

Substituting (2-6) and (2-7) in (2-8) provides us:

\[ \tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm1}I_{s1} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1}) + \lambda_{pm2}I_{s2} \sin(\omega_{s2}t + \theta_{r2} + \theta_{g2}) \right. \]
\[ \left. + (L_{md} - L_{mq})I_{s1}I_{s2} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1} + \omega_{s2}t + \theta_{r2} + \theta_{g2}) \right] \]  \hspace{1cm} (2-9)

Since for an SMPMSM, the air gap variation component of the inductances (\( L_{v}, L_{mv} = 0 \) in (B-9)) is negligible, therefore, \( L_{md} \approx L_{mq} \) and thus

\[ \tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm1}I_{s1} \sin(\omega_{s1}t + \theta_{r1} + \theta_{g1}) + \lambda_{pm2}I_{s2} \sin(\omega_{s2}t + \theta_{r2} + \theta_{g2}) \right] \]  \hspace{1cm} (2-10)

Equation (2-10) shows that the torque generated on the rotor of a six-phase SPMSM with balanced sinusoidal currents in its stator windings will have a sinusoidal nature and will have a frequency same as that of the stator currents. From the mechanical model of the machine the change in the displacement will also be a sinusoid having the grid current frequency and an amplitude that depends on the rotor inertia \( J \).

A comparison between the torque developed on the rotor of a six-phase 100 kW SPMSM calculated using (2-10) and obtained by an FEA model of the same machine (parameters presented
in Table A-1) with 40A current per phase is shown in Figure 2-4. The results verify the empirical model derived. The FEA results were calculated at T=20 °C.

2.2.1.2 Interior permanent magnet synchronous machines

For six-phase IPMSMs the total torque produced on the shaft will be given by (B-13). Substituting the currents from (2-6) and (2-7) and the knowledge that \( L_{d1} \neq L_{q1} \) and \( L_{d2} \neq L_{q2} \) but \( L_{d1} = L_{d2} = L_d \) and \( L_{q1} = L_{q2} = L_q \) (by design) we get:

\[
\tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm} (I_{s1} \sin(\omega_{s1} t + \theta_{r1} + \theta_{g1}) + I_{s2} \sin(\omega_{s2} t + \theta_{r2} + \theta_{g2})) + \frac{(L_d - L_q)}{2} (I_{s1}^2 \sin(2\omega_{s1} t + 2\theta_{r1} + 2\theta_{g1})) \right. \\
+ \left. I_{s2}^2 \sin(2\omega_{s2} t + 2\theta_{r2} + 2\theta_{g2})) \right. \\
+ \left. (L_{md} - L_{mq}) I_{s1} I_{s2} \sin(\omega_{s1} t + \omega_{s2} t + \theta_{r1} + \theta_{r2} + \theta_{g1} + \theta_{g2}) \right]
\]

where \( \theta_{g1} \) and \( \theta_{g2} \) are the angles between the grid voltage and the grid currents of the drives 1 and 2 respectively.
Equation (2-11) shows that the torque generated on the rotor of a six-phase IPMSM with balanced sinusoidal currents in its stator windings will consist of five components. The first two components will be sinusoids having the same frequency as of the current that excites their respective stator windings. The second two components will have twice the frequency of the current of their respective stator windings, and the fifth component will be a sinusoid having a frequency that is the sum of the frequency of the two currents that excite each set of the stator windings. From the mechanical model of the machine the change in rotor displacement will also be sinusoidal and will have the same number of components as the torque. The amplitude of the displacement will depend on the rotor inertia $J$.

A comparison between the torque developed on the rotor for a 100kW IPMSM machine calculated using (2-11) and obtained by an FEA model of the same machine (all the machine parameters are presented in) with 40A current per phase through its stator windings is shown in Figure 2-5. The result shown verifies the empirical prediction of the model used. The FEA results were calculated at $T=20$ °C.
2.2.1.3 Synchronous reluctance machines

For six-phase SRMs, the torque produced on the shaft will be given by (B-13). Substituting the currents from (2-6) and (2-7) and the knowledge that for SRMs \( \lambda_{pm1} = \lambda_{pm2} = 0 \), \( L_{d1} \neq L_{q1}\) and \( L_{d2} \neq L_{q2}\) but \( L_{d1} = L_{d2} = L_d \) and \( L_{q1} = L_{q2} = L_q \) we get:

\[
\tau_{em} = \frac{3}{2} \cdot p \left[ \frac{(L_d - L_q)}{2} \left( I_{s1}^2 \sin(2\omega s_1 t + 2\theta_r + 2\theta_g) 
+ I_{s2}^2 \sin(2\omega s_2 t + 2\theta_r + 2\theta_g) \right) 
+ (L_{md} - L_{mq})I_s I_s \sin(\omega s_1 t + \omega s_2 t + \theta_r + \theta_r + \theta_g + \theta_g) \right]
\]

Equation (2-12) shows that the torque generated on the rotor of a six-phase SRM with balanced sinusoidal currents in its stator windings will consist of three components. The first two components will be sinusoids having the same frequency as of the current that excites their respective stator windings, and the third component will be a sinusoid having a frequency that is the sum of the frequency of the two currents that excite each set of the stator windings. From the mechanical model of the machine the change in rotor displacement will also be sinusoidal and
consist of the same components as the torque. The amplitude of the displacement will depend on the rotor inertia $J$.

A comparison between the torque developed on the rotor for a 100kW machine calculated using (2-12) and obtained by an FEA model of the same machine (all the machine parameters are presented in Table A-1) with 40A current per phase through its stator windings is shown in Figure 2-6. The result shown verifies the empirical prediction of the model used. The error exists because the empirical prediction does not capture the nuances of the shape of the rotor. The FEA results were calculated at $T=20 \, ^\circ \text{C}$.

### 2.3 Torque Control During the Charging Operation

The torque analyzed in sections 2.1-2.2 will produce noise and vibration issues for the drive transmission to which the motor is connected [86]. The drivetrain for an electric vehicle is conventionally designed to meet twice the rated torque to meet shock loads and sudden accelerations [87]. The torque generated during integrated battery charging may be as high as the rated torque (Figure 2-2), which may not be detrimental to the drivetrain immediately. It should also be noted however, that the vibrations produced by the torque oscillation during charging may be damped if the total inertia on the shaft is large. However, a continuous application of the torque with a frequency of grid fundamental frequency may result in an early failure of the drivetrain.

A way to reduce the torque vibrations on the shaft can be to apply mechanical damping. Mechanical damping of the vibrations will increase the size and weight of the overall vehicle [88]. Another way of controlling the pulsating torque produced by the machine during charge is by controlling the current references for each of the stator windings such that the torque produced by each set of windings is eliminated by the torque produced by the other set. This section proposes and validates torque elimination strategies for IBC based on three-phase and six-phase drives.

#### 2.3.1 Two three-phase machine drives

Electric vehicles with multiple three-phase synchronous machine drives with shafts coupled by a torque coupler have been increasingly used for high-performance vehicles due to increased modularity and fault tolerance [89]. Some hybrid architectures also employ multiple three-phase machine drives that have their torques added on the drive shaft [2]. Such torque couplers allow the
two machines to operate at different speeds. Machine drives with directly coupled shafts are also possible [90]. A typical drive with two three-phase machines is shown in Figure 2-7. Using such machine drives it is possible for the total torque on the drive shaft to be reduced if not eliminated. This section presents the torque reduction strategy for the machine drives with two machines on the same shaft. The conceptual diagram of two machines with a common shaft is shown in Figure 2-8, showing the relation between the d-axes of the two machines. Since the shafts of the machine are coupled, and relative movement between the two may be allowed, therefore, the displacement between the rotor dq-axes of both machines is generalized by considering it to be $\theta_r$.

The torque generated by a three-phase synchronous machine drive is given by (B-5). Since the individual torques produced by the two machine drives is added on the shaft, the total torque on the shaft can be written by substituting (2-1) for machine 1 and machine 2 as:
\[ \tau = \tau_1 + \tau_2 = \frac{3}{2} \left\{ p_1 \left[ \lambda_{pm_1} I_{s1} \sin(\omega_{s1} t + \theta_{r1} + \theta_{g1}) \right. \right. \\
- \frac{1}{2} (L_{q_{s1}} - L_{d_{s1}}) I_{s1}^2 \sin(2\omega_{s1} t + 2\theta_{r1} + 2\theta_{g1}) \left. \left. \right] + p_2 \left[ \lambda_{pm_2} I_{s2} \sin(\omega_{s2} t + \theta_{r2} + \theta_{g2}) \right. \right. \\
- \frac{1}{2} (L_{q_{s2}} - L_{d_{s2}}) I_{s2}^2 \sin(2\omega_{s2} t + 2\theta_{r2} + 2\theta_{g2}) \right\} \]

where \( \tau_1 \) and \( \tau_2 \) are the torques produced by machine 1 and machine 2 respectively, the symbols with the subscript 1 belong to the machine drive 1 and the symbols with the subscript 2 belong to the machine drive 2 and \( \theta_{g1} \) and \( \theta_{g2} \) are the angles between the grid voltage and the grid currents of the drives 1 and 2 respectively.

Equation (2-13) presents the total torque on the shaft which can be eliminated if \( \tau_1 = -\tau_2 \). Meeting this condition, however, is not always possible due to the differences in the machine ratings and designs. The following subsections present the reference selection for the winding currents for machines with the same technology (SPMSM, IPMSM or SRM), however, the strategies are general and can be extended to use for machines with different technologies as well.

Figure 2-8 The relationship between the d-q axes of the two three-phase machines
2.3.1.1 Two SPMSM drives

Since for SPM machines \( L_d \approx L_q \), the total torque equation will be given by

\[
\tau = \tau_1 + \tau_2
\]

\[
= \frac{3}{2} p_1 \lambda_{pm_1} I_s 1 \sin(\omega_s t + \theta_{r1} + \theta_{g1}) + \frac{3}{2} p_2 \lambda_{pm_2} I_s 2 \sin(\omega_s t + \theta_{r2} + \theta_{g2})
\]  \hspace{1cm} (2-14)

From (2-14), the torque elimination condition gives us

\[
p_1 \lambda_{pm_1} I_s 1 \sin(\omega_s t + \theta_{r1} + \theta_{g1}) = -p_2 \lambda_{pm_2} I_s 2 \sin(\omega_s t + \theta_{r2} + \theta_{g2})
\]

For both sides to be equal and opposite, the magnitude of the current references can be set by adjusting the current magnitudes \( I_s 1 \) and \( I_s 2 \). The magnitude is selected as

\[
p_1 \lambda_{pm_1} I_s 1 = p_2 \lambda_{pm_2} I_s 2
\]

\[
I_s 1 = \frac{p_2 \lambda_{pm_2}}{p_1 \lambda_{pm_1}} I_s 2
\]  \hspace{1cm} (2-15)

The sign of the torque generated can only be controlled by adjusting the angle or frequency of the sine in the equation. One way of adjusting for the sign is to have the reference angles of the currents meet the condition \( \theta_{g1} = \theta_{g2} + \pi \). Since the machines are parallel connected electrically, this would result in active power circulation between the two machines and no power exchange with the grid, therefore, the sign can only be set by meeting the following condition:

\[
\omega_s t + \theta_{r1} + \theta_{g1} = -\omega_s t - \theta_{r2} - \theta_{g2}
\]  \hspace{1cm} (2-16)

Since the frequency of both the machine sets will be equal to the grid frequency, they can have opposite signs by connecting the second machine with the opposite phase sequence as that of the first. Second, \( \theta_{g1} \) and \( \theta_{g2} \) can be selected so that \( \theta_{r1} - \theta_{r2} = \theta_r = \theta_{g1} - \theta_{g2} \).

It should be noted that if both the machines are identical, (2-15) becomes \( I_s 1 = I_s 2 \) and similarly in (2-16) \( \theta_{r1} - \theta_{r2} = 0 \). In this case, the torque is eliminated with equal currents and there is no need of phase angle adjustment for the two machine set currents.

2.3.1.2 Two IPMSM drives

For IPMSMs, the total torque equation is given by

\[
\tau = \tau_1 + \tau_2
\]  \hspace{1cm} (2-17)
For SRMs, the total torque equation is given by

\[
\tau = \tau_1 + \tau_2
\]

\[
= -\frac{3}{4} p_1 \left[ (L_{qs1} - L_{ds1}) I_{s1}^2 \sin(2\omega_{s1} t + 2\theta_{r1} + 2\theta_{g1}) \right],
\]

\[
-\frac{3}{4} p_2 \left[ (L_{qs2} - L_{ds2}) I_{s2}^2 \sin(2\omega_{s2} t + 2\theta_{r2} + 2\theta_{g2}) \right].
\]  

(2-18)

The negative sign is again ensured by following the reference given in (2-16), however, unlike the IPMSMs, it is possible to find an exact nullifier by selecting the current reference to be chosen as

\[
p_1 (L_{qs1} - L_{ds1}) I_{s1}^2 = p_2 (L_{qs2} - L_{ds2}) I_{s2}^2.
\]  

(2-19)
$$I_{s1} = \frac{p_2(L_{q_{s2}} - L_{d_{s2}})}{\sqrt{p_1(L_{q_{s1}} - L_{d_{s1}})}} I_{s2}$$

Like the other two machine types, if both SRMs are identical in design and ratings, the torque can be eliminated by $I_{s1} = I_{s2}$ and without phase angle adjustments.

### 2.3.2 Validation of torque cancellation strategy with two three-phase machine drives

#### 2.3.2.1 Validation with identical machines

To validate the torque elimination strategies outlined in the previous sections, FEA was performed using several designs of SPMSMs, IPMSMs, and SRMs. The specifications of the machines are provided in Table A-1. The results for the charger operating at 28.8kW power exchange with the grid were generated for identical machine types and differently rated machine types and are presented in this section.

The simulation results for machine drives with identically designed machines are shown in Figure 2-9 to Figure 2-11. Figure 2-9 shows the total torque on the common shaft of two identical 30kW SPMSMs for three grid fundamental cycles. The torque on the shaft without the torque

![Figure 2-9 Simulated FEA total torque on the shaft for two identical three-phase SPMSMs with a common shaft for 28.8kW charging operation, with (w) and without (wo) torque cancellation.](image-url)
control strategy has peaks of 135Nm, and with the torque control strategy, the total torque on the shaft is eliminated. The currents of the two machines in the rotor frame of reference are shown in Figure 2-12, showing that the currents satisfy the torque minimizing condition derived from (2-14).

In Figure 2-10 three grid fundamental cycles of total torque on the common shaft of two identical 30kW IPMSMs. The total torque on the shaft without the torque control strategy has peaks of 157Nm, and with the torque control strategy, the total torque is eliminated. Similarly, in Figure 2-11 the total torque on the shaft of two identical 30kW SRMs is shown for three grid fundamental cycles. The total torque on the shaft without the torque control strategy peaks at 68Nm and with the torque control strategy the total is eliminated. For all the three cases shown, as hypothesized, the total torque on the shaft is eliminated completely.
Figure 2-11 Simulated FEA total torque on the shaft for two identical three-phase SRMs with a common shaft for 28.8kW charging operation, with (w) and without (wo) torque cancellation.

Figure 2-12 Simulated results for currents in rotor frame of reference of two identical three-phase SPMSMs for 28.8kW charging operation.
Figure 2-13 Simulated FEA total shaft torque for two differently rated three-phase SPMSMs with a common shaft for 28.8kW charging operation, with (w) and without (wo) torque cancellation.

Figure 2-14 Simulated FEA total torque on the shaft for differently rated three-phase IPMSMs with a common shaft for 28.8kW charging operation, with (w) and without (wo) torque cancellation.
2.3.2.2 Validation with machines of different ratings

The torque results for three grid fundamental cycles for two differently rated machines are shown in Figure 2-13 to Figure 2-15. Figure 2-13 shows the total torque on the common shaft of a 30kW and a 100kW SPMSM. The torque on the shaft without the torque control strategy has peaks of 93Nm, and with the torque control strategy, the total torque on the shaft is eliminated. In Figure 2-14 the total torque on the common shaft of a 30kW and a 100kW IPMSM is shown. The total torque on the shaft without the torque control strategy has peaks of 135Nm, and with the torque control strategy, the total torque peaks are reduced to 20Nm. Similarly, in Figure 2-15 the total torque on the shaft of a 30kW and a 100kW SRM is shown. The total torque on the shaft without the torque control strategy peaks at 62Nm and with the torque control strategy the total peak is reduced to 4Nm. The total torque on the shaft is eliminated only for the case of two differently rated SPMSMs, whereas for the case of different IPMSMs, the total torque cannot be eliminated but only minimized. In the displayed result, the reluctance torque was chosen to be eliminated as it resulted in a lower total torque peak than eliminating the excitation component. The case for the
differently rated SRMs should have resulted in total torque elimination; however, due to the slot effects, there is residual total-torque albeit of a very small value (6% of the uncompensated torque).

2.3.2.3 Experimental validation with identical machines

For experimental validation, a scaled-down hardware setup with two identical 37 kW SMPM machines was built, the specifications of the setup are presented in Appendix E. The power of the system was scaled down while the base impedance of the system was kept constant such that the inductance offered by the machine windings remained the same, i.e., 0.07 pu. A total of 19.2 Arms per phase was drawn from the grid, with each drive converter drawing 9.6 Arms per phase at unity power factor. The angular displacement of the rotor shaft was measured to evaluate torque cancellation. The results of the rotor oscillations are shown in Figure 2-16. When charging occurs without torque cancellation, the rotor oscillates more than when it occurs with the torque control strategy (reduced by 96%). The input currents in the rotor frame of reference of their respective machines are shown in Figure 2-17, $I_{q1}$ and $I_{q2}$ are equal and opposite as required by (2-15). Further tests were performed at fractions of rated power for unity power factor charging with different relative rotor displacement angles $\theta_r$. The results for relative rotor displacement of 0°, 65° and -75° are shown in Figure 2-18(a), (b) and (c) respectively. The torque was controlled at zero by

![Figure 2-16](image-url) Experimentally measured rotor angular displacement of the identical three-phase SPMSMs for rated charging operation, with (w) and without (wo) torque cancellation.
Figure 2-17 Experimentally measured currents in rotor frame of reference for identical three-phase SPMSMs for rated charging operation.

Figure 2-18 Experimentally measured rotor oscillation comparison for different charging powers with (a) no relative rotor displacement; (b) 65° rotor displacement; and (c) -75° rotor displacement.
in section 2.3.1.1. The results shown are normalized to the worst-case oscillations of the rotor when operated at 1pu (i.e., 9.6 Arms per converter). The rotor oscillations for all the current levels tested are considerably reduced when the proposed torque elimination strategy is used with a reduction of 96% for the worst case shown in Figure 2-16. Oscillations due to the four quadrant operation of the charger were also recorded and are presented in Figure 2-19. The modes are as follows: I – unity power factor charging; II – 0.9 leading power factor charging; III – 0.9 lagging power factor charging; IV – unity power factor discharging; V – 0.9 leading power factor discharging; VI – 0.9 lagging power factor discharging; VII – leading reactive power; and VIII – lagging reactive power. The result shows that the torque produced in the machine is independent of the grid power factor and the torque cancellation strategy is able to cancel the torque despite a non-unity power factor operation at the grid interface.

2.3.3 Six-phase machine drives

Six-phase machines provide extra degrees of freedom in their control and therefore have been gaining interest in the industry for EV traction [54]. These machines are used to provide redundancy [91] and better fault-tolerant performance for drives for increased reliability [92] and current distribution for large machines rated for high power [93]. A typical six-phase drive is shown in Figure 2-20. The extra degree of freedom provided by these drives can be used to eliminate the torque produced by the three-phase charging currents in the machine windings. This section presents the torque reduction strategy for the machine drives with six phase synchronous
machines. The conceptual diagram of a six-phase machine with the two phase-sets, ABC and DEF, was introduced in section 2.2 and is shown in Figure B-2. The figure shows a general six-phase motor with two phase-sets displaced by an angle of $2\alpha^\circ$ from each other. The rotor displacement $\theta_r$ is measured from the effective stator d-axis (mid-point of the set ABC d-axis and DEF d-axis).

The following sections present the torque cancellation strategy for IBCs based on six-phase machines of the three machine technologies.

The total torque on the shaft of a six-phase PMSM is given by (2-11) and is repeated here:

$$
\tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm} (I_{s1} \sin(\omega_{s1} t + \theta_{r1} + \theta_{g1}) + I_{s2} \sin(\omega_{s2} t + \theta_{r2} + \theta_{g2})) + \frac{(L_{d} - L_{q})}{2} (I_{s1}^2 \sin(2\omega_{s1} t + 2\theta_{r1} + 2\theta_{g1}) + I_{s2}^2 \sin(2\omega_{s2} t + 2\theta_{r2} + 2\theta_{g2})) + (L_{md} - L_{mq}) I_{s1} I_{s2} \sin(\omega_{s1} t + \omega_{s2} t + \theta_{r1} + \theta_{r2} + \theta_{g1} + \theta_{g2}) \right]
$$

(2-20)

The torque elimination condition is derived as

$$
I_{s1} \sin(\omega_{s1} t + \theta_{r} + \alpha + \theta_{g1}) = -I_{s2} \sin(\omega_{s2} t + \theta_{r} - \alpha + \theta_{g2})
$$

(2-21)

For both sides to be equal and opposite, the magnitude of the current references can be set by adjusting the current magnitudes $I_{s1}$ and $I_{s2}$. The magnitude is selected as
\[ I_{s1} = I_{s2} \] (2.22)

Like the two SPMSM drives, the sign requirement can only be met by meeting the following condition:

\[ \omega_{s1} t + \theta_r + \theta_{g1} = -\omega_{s2} t - \theta_r - \theta_{g2} \] (2.23)

Since the frequency of both the phase-sets will be equal to the grid frequency, they can have opposite signs by connecting the second phase-set with the opposite phase sequence as that of the first. Second, \( \theta_{g1} \) and \( \theta_{g2} \) can be selected as \( \theta_{g1} = -\theta_r \) and \( \theta_{g2} = \theta_r \).

The conditions derived here apply to all the three kinds of synchronous machines. For SPMSMs, \( L_d \approx L_q \) and \( L_{md} \approx L_{mq} \), therefore, the reluctance torque is negligible, and the excitation torque is directly cancelled. For IPMSMs, the reluctance torque and the excitation torque both are canceled. For SRMs, the excitation torque does not exist because \( \lambda_{pm} = 0 \), and the reluctance torque is canceled.

### 2.3.4 Validation of torque cancellation strategy with six-phase machine drives

#### 2.3.4.1 Simulation validation

The torque cancellation strategy was validated using the six-phase SPMSM, IPMSM and SRM mentioned in section 2.2.1. The parameters of these machines are presented in Table A-2. The torque on the shaft of the six-phase SPMSM with and without the torque elimination strategy is shown in Figure 2-21. Without the torque elimination strategy, the total torque on the shaft reaches a maximum of 206.9 Nm, while with the torque elimination strategy the maximum torque is reduced to 0.02 Nm. The currents of the two phase-sets in the rotor frame of reference are shown in Figure 2-22, showing that they meet the torque elimination condition in (2-21).

Figure 2-23 shows the torque on the shaft of a six-phase IPMSM with and without the torque elimination strategy. The maximum torque without the strategy is 151.7 Nm which is reduced to 2.9 Nm when the strategy is used. The residual torque is due to the reluctance torque of the machine slots. Similarly, the shaft torque of a six-phase SRM is shown in Figure 2-24. The maximum torque without the strategy is 444.8 Nm which is reduced to 0.41 Nm when the torque elimination strategy is used.
2.3.4.2 Experimental validation

For experimental validation, a hardware setup with a 100 kW IPMSM was built, the specifications and the details of the setup are presented in Appendix E, while the machine parameters are presented in Table A-3. A total of 20A per phase (7.2 kW) was drawn from the grid, with 10A per phase per drive converter at unity power factor. The rotor angular displacement results for the 7.2kW charging at unity power factor is shown in Figure 2-25. The figure shows that the rotor oscillations are eliminated when the proposed torque control strategy is used. The two phase-set currents in the rotor frame of reference are presented in Figure 2-26 showing that they meet the torque elimination condition in (2-21). The case for the rotor displaced by an angle of 15°, and -30° was also tested, and the results for the rotor oscillations at different power levels are shown in Figure 2-27. The 60Hz and 120Hz components of the oscillations in the rotor displacement are shown in Figure 2-27 (a)-(b), (c)-(d) and (e)-(f) for 0°, 15° and -30° rotor displacements respectively. The measurements are all normalized to the maximum rotor displacement caused for 60Hz at 12A per phase per phase-set current drawn from the grid. The torque elimination strategy manages to effectively eliminate the torque on the shaft and hence the resultant oscillations in the rotor displacement.
Oscillations due to the four-quadrant operation of the charger were also recorded and are presented in Figure 2-28. The modes are as follows: I – unity power factor charging; II – 0.9 leading power factor charging; III – 0.9 lagging power factor charging; IV – unity power factor discharging; V – 0.9 leading power factor discharging; VI – 0.9 lagging power factor discharging; VII – leading reactive power; and VIII – lagging reactive power. The result shows that the torque produced in the machine is independent of the grid power factor and the torque cancellation strategy is able to cancel the torque despite a non-unity power factor operation at the grid interface.
Figure 2-23 Simulated FEA total shaft torque for six-phase IPMSM for 28.8kW charging operation, with (w) and without (wo) torque cancellation.

Figure 2-24 Simulated FEA total torque on the shaft for a six-phase SRM for 28.8kW charging operation, with (w) and without (wo) torque cancellation.
2.4 Effect of accuracy and parameters on torque cancellation

The torque cancellation strategy was assessed against the error in position sensing for the six-phase machine and against stator saturation due to charging at machine rated or higher currents. Each assessment is detailed as follows.

2.4.1 Position sensor

A study was conducted to assess the sensitivity of the torque control strategy to the error in position due to sensor accuracy. Three cases with $\theta_r = 0^\circ, 15^\circ$ and $30^\circ$ were studied and a position error corresponding to the least significant bit (LSB) of different resolutions was added to the rotor position. The results are shown in Figure 2-29(a). The figure shows that if an 8-bit encoder is used to sense the rotor position, the LSB error will correspond to $11.25^\circ$ electrical and result in the residual torque going up to 20% of the rated torque. If a 14-bit encoder is used the residual torque due to the accuracy error amounts to 1% of the rated torque.
2.4.2 Effect of non-linearities

The effect of machine non-linearities to its parameter variations, such as the effect of saturation on the machine inductances, on the torque cancellation was studied. Charging at higher than rated currents (to ensure stator saturation) was tested on the FEA model. The residual torque on the shaft against the charging current is shown in Figure 2-29(b). The results show that with the current, the residual torque also increases, especially beyond the point where parts of the stator start to saturate. The residual torque increase is because of the q-axis inductance, $L_{q12}$ in (2-17) and (2-20), for both phase-sets are dependent on the stator current, and therefore the assumption that they are equal is not true anymore, resulting in a residual torque. This implies that the upper limit of the IBC charging current should be less than the machine rated current to ensure that the machine does not go into saturation during the charging operation. The residual torque for $\theta_r = 15^\circ$ is higher than when $\theta_r = 0^\circ$ and $30^\circ$. This is because the rotor d-axis is between stator teeth for $\theta_r = 15^\circ$ resulting in a detent torque.
2.5 Conclusion

The problem of torque production on the shaft, while the battery is being charged for IBCs based on three-phase synchronous machines, was first presented in this chapter. Two different torque cancellation strategies were proposed. The first was based on two synchronous machines on a common shaft and the second was based on a six-phase synchronous machine. The proposed strategies were tested on models of nine different machines (designs shown in Appendix A). The strategies were shown to effectively cancel the total torque on the shaft for all kinds of machine combinations except for when differently rated IPMSMs or SRMs are used for the strategy based on two synchronous machines. The validation included analytical machine models, finite element analysis (FEA) machine models and hardware experiments. The torque cancellation helps in reducing the unwanted noise, vibration and potential deterioration of the mechanical drivetrain.

Figure 2-27 Experimentally measured rotor angular displacement oscillation normalized to the worst-case displacement for different charging powers with (a) 0°; (b) 15°; and (c) -30° rotor position from the stator d-axis. (a), (c) and (e) show the 60 Hz component and (b), (d) and (f) show the 120 Hz component of the angular displacement.
Figure 2-28 Shaft oscillations due to the four quadrant operation of the rectifier.

Figure 2-29 Simulated residual torque on the rotor of a machine with $\alpha=0^\circ$ normalized to the rated torque against (a) position sensor resolution at rated charging power; (b) winding currents normalized to rated current in charging operation.
3 Topologies and Control of Integrated Battery Chargers

The torque cancellation strategies proposed in the previous chapter can be used only if implemented on existing EV drive topologies. The IBC topologies that result pose several challenges including the control of current with unbalanced filter inductances (common for IPMs and SRMs) with increased second harmonic oscillation on the power sent to the DC link and circulating current, and the circulating current flow between the parallel connected drive converters. This chapter proposes topologies and control algorithms to practically implement integrated battery chargers based on three-phase and six-phase machines. Control algorithms that meet the earlier mentioned challenges are also presented. The algorithms are validated using simulation and hardware experiments.

3.1 Topologies

The torque cancellation strategy can be implemented on an EV drive train that has a traction motor with two or more sets of three-phase windings using winding reconfiguration devices such as contactors. Similarly, drivetrains with two three-phase motors can also be used. A drivetrain with a single three-phase motor can be used by splitting the three-phase winding into two three-phase sets [46, 47], however, this can only be achieved on motors that have stator windings with at least two slots per phase per pole. Traction motors with one slot per phase per pole windings can also be used, if they are wound with double layers.

Contactors can be used for reconfiguration of the machine windings to split the phases, to disconnect the winding from traction mode configuration to the charging mode configuration. They can be placed on board as close to the traction motor as possible. If not placed close the traction motor, they will require additional cables that will appear in series with the motor windings, adding to the voltage drop across it and its losses. Additionally, the contactors also introduce a contact resistance in series, this will add to the voltage drop and losses. Furthermore, a redesign of the traction drive might be required to adjust the contactors in the drivetrain. Despite these drawbacks, the total added weight of the contactors is expected to be less than the weight of a dedicated high-power charger with its filters and power electronics.
Two topology implementations with reconfiguration contactors for two three-phase and six-phase machines are proposed and presented in the following sections.

### 3.1.1 IBC for EVs with two three-phase synchronous motors

The proposed topology of the three-phase charger based on two three-phase synchronous machines is shown in Figure 3-1. In the traction mode of operation, the contactors reconfigure the windings of each machine to be connected in a star or a delta configuration. In the charging mode of operation, the contactors disconnect the machine windings and connect them to the grid. The first machine, $M_1$, is connected with a positive phase sequence, i.e., ABC and the second machine, $M_2$, is connected with a negative phase sequence, i.e., ACB as required by the torque cancellation strategy presented in the previous chapter. Both motors have their shafts coupled by the torque coupler T as shown in the figure. The torque coupler allows the torque of the individual machines to be added to the drive transmission with both operating at different [2] or same speeds [90]. The machine windings are reconfigured to act as a series three-phase inductance for the traction drive to connect to the three-phase grid. The traction motor for the IBC can have star connected, delta connected [94] or open-ended windings, all of which would be reconfigured to have a topology as shown in Figure 3-1. The traction drive is a DC to AC three-phase power electronic converter with bi-directional power flow capability and controls the DC link at a constant voltage. The DC-DC converter steps down the DC link voltage to the battery voltage. This converter is needed to match variability in the battery terminal voltage and has a bidirectional power flow capability.
3.1.2 IBC for EVs with six-phase synchronous motors

The topology of the three-phase charger based on a six-phase synchronous machine is shown in Figure 3-2. In the traction mode of operation, the contactors reconfigure the windings of each phase-set to be connected in a star each or a delta each or one star and one delta. In the charging mode of operation, the contactors disconnect the machine windings and connect them to the grid. The first phase-set is connected with a positive phase sequence, i.e., ABC and the second phase-set is connected with a negative phase sequence, i.e., ACB as required by the torque cancellation strategy presented in the previous chapter. The machine windings are reconfigured to act as a series three-phase inductance for the traction drive to connect to the three-phase grid. Like the first topology, the traction motor for the IBC can have both star-connected, or delta-connected [94] or one star- and one delta-connected, or open-ended windings, all of which would be reconfigured to have a topology as shown in Figure 3-2. The traction drive is a DC to AC three-phase power electronic converter with bi-directional power flow capability and controls the DC link at a constant voltage. Similar to the first topology, the DC-DC converter steps down the DC link voltage to the battery voltage and is needed to match the variability in the battery terminal voltage and has a bidirectional power flow capability.

The topologies proposed in this section represent the majority of the traction drives for low and medium duty electric vehicles. The traction inverter, a two-level converter, has been the industry standard for low voltage applications (with DC link voltages up to 800Vdc) due to its simplicity, high switch utilization, and low cost [95]. The motor technology most used for electric vehicle applications is that of a three-phase or multiples of three-phase (for example six-phase and nine-phase motors) motors due to the simpler scalability provided by them, opposed to other topologies such as five and seven phase machines. Therefore, the topology of a two-level converter connected
to a three-phase machine is representative of most of the traction drives [2], and therefore other topologies are not considered in this thesis.

### 3.2 IBC Modeling and Control

The resultant topology of the IBC is that of two voltage source converters connected in parallel via the machine winding inductances and is shown in Figure 3-3. The topology can be operated as a grid-connected bi-directional converter. The rotor position dependence of the winding inductances of permanent magnet synchronous machines and synchronous reluctance machines results in unequal phase and mutual inductances. The position-dependent inductances make the topology of Figure 3-3 different from that of conventional parallel connected voltage source converters. This section presents the modeling of the IBC and shows why the conventional rotating dq frame modeling is not suitable for its control. It also presents the control algorithms that can control the IBC in both battery charging mode (when the IBC absorbs active power to charge the battery) and the vehicle to grid (V2G) mode (when the IBC operates in all four quadrants of the power plane).

#### 3.2.1 IBC modeling

Conventionally the interface filter inductances of a grid-connected voltage source converter are modeled in the rotating frame of reference that is synchronized with the grid voltage. Rotating dq transformation is not possible because $\theta_r$ is constant for the stationary rotor. To show this, consider

Figure 3-3 Equivalent IBC topology
the IBC with the six-phase machine windings. The transformation, synchronized to the grid angle $\omega_g t$, applied to the inductance matrix in (B-19) is shown below:

$$T_6(\omega_g t)L_{abcdef}(\theta_r, \alpha)T_6^{-1}(\omega_g t)$$

$$= \begin{bmatrix} T_3(\omega_g t) & 0 \\ 0 & T_3(-\omega_g t) \end{bmatrix} \begin{bmatrix} L_{abc}(\theta_r, \alpha) & L_{m_{12}}(\theta_r, \alpha) \\ L_{m_{21}}(\theta_r, \alpha) & L_{def}(\theta_r, \alpha) \end{bmatrix} \begin{bmatrix} T_3(\omega_g t) & 0 \\ 0 & T_3(-\omega_g t) \end{bmatrix}^{-1}$$

$$= \begin{bmatrix} T_3(\omega_g t)L_{abc}(\theta_r, \alpha)T_3^{-1}(\omega_g t) & T_3(\omega_g t)L_{m_{12}}(\theta_r, \alpha)T_3^{-1}(-\omega_g t) \\ T_3(-\omega_g t)L_{m_{12}}(\theta_r, \alpha)T_3^{-1}(\omega_g t) & T_3(-\omega_g t)L_{def}(\theta_r, \alpha)T_3^{-1}(-\omega_g t) \end{bmatrix}$$

After transformation, the first element of the matrix is given by

$$T_3(\omega_g t)L_{abc}(\theta_r, \alpha)T_3^{-1}(\omega_g t)$$

$$= \frac{3}{2}\begin{bmatrix} L_0 + L_v\cos(2\omega_g t) + L_l & -L_v\sin(2\omega_g t) & 0 \\ -L_v\sin(2\omega_g t) & L_0 - L_v\cos(2\omega_g t) + L_l & 0 \\ 0 & 0 & L_l \end{bmatrix}$$

(3-1)

The first element of the matrix, as shown by (3-1) makes the inductance a function of time due to the presence of the term $\omega_g t$ in it. Furthermore, it does not decouple the inductance matrix losing the purpose of transforming to the rotating grid frame of reference in the first place. Therefore, the IBC system must be controlled in the stationary grid frame of reference as six individual phases without transformations.

Due to the time dependence of the inductance matrix in the rotating frame, the IBC was modeled as a six-phase system as shown in Appendix B. Linear quadratic regulator with integral action (LQRI) control was used, instead of regular PI controllers, on the thus formed six-phase system as it does not require an explicit decoupling of the phases and is capable of controlling coupled multi-input multi-output systems [96-98]. The control law for the six-phase system is presented in Appendix D.

3.2.1.1 Control validation

Both topologies were tested with simulation and hardware experiments to validate the developed machine model and its control. The current references selected were such that the charger operates at unity power factor charging, 0.9 power factor charging and unity power factor discharging.
Figure 3-4 Simulated results for unity power factor charging at 28.8kW (a) AC; and (b) DC voltage and currents.

Figure 3-5 Simulated results for leading 0.9 power factor charging at 26kW (a) AC; and (b) DC voltage and currents.
The simulation results for the two three-phase machine based topology showing the AC grid and DC link voltages and currents are shown in Figure 3-4, Figure 3-5, and Figure 3-6 for unity power factor charging, 0.9 power factor charging and unity power factor discharging. In the same order, the experimental results for the topology are shown in Figure 3-7, Figure 3-8, and Figure 3-9 respectively.

The results show that the topology is controlled to follow the references provided. The parameters for the simulation setup and the hardware setup based on the two three-phase machine based IBC are provided in Appendix A.

Similarly, IBC topology based on the six-phase machine was also tested with unity power factor charging and different values of $\theta_r$. The AC grid and DC link voltages and currents for unity power factor charging and $\theta_r = 0^\circ$ are shown in Figure 3-4. Further tests were conducted with different values of $\theta_r$, and the results are shown in Figure 3-10 and Figure 3-11.
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Figure 3-7 Experimental results for unity power factor charging at rated power (a) AC; and (b) DC voltage and currents.

Figure 3-8 Experimental results for leading 0.9 power factor discharging at 0.9 pu (a) AC; and (b) DC voltage and currents.
The test was performed on machines with $\alpha = 0^\circ, 30^\circ$. Figure 3-10 (a) and (b) show the torque on the shaft of machines with $\alpha = 0^\circ$ and $\alpha = 30^\circ$ respectively for $\theta_r = 30^\circ, 15^\circ, 0^\circ, -15^\circ, -30^\circ$. The results shown in Figure 3-10 are all shown without introducing a phase difference between the currents of the two phase-sets. We see that for both machines, the torque is eliminated for $\theta_r = 0^\circ$, however, for other values of $\theta_r$, we see that there is a residual torque on the shaft. This torque is eliminated as shown in Figure 3-11 (a) when the appropriate phase difference between the currents was implemented. The currents for $\theta_r = 15^\circ$, and $30^\circ$ are shown in Figure 3-11 (b) and (c) respectively.
Hardware results for a scaled down setup of the IBC topology based on the six-phase machine was also tested, and the results of the same are presented in Figure 3-12, Figure 3-13, and Figure 3-14 for unity power factor charging, unity power factor discharging and for unity power factor charging with $\theta_r = 30^\circ$ respectively. The ratings and parameters of the hardware setup are presented in Appendix E.
The individual converter currents shown in Figure 3-14 have a significant third harmonic component. This is because once the machine windings are connected in parallel, the common mode impedance between the phase sets is very low, therefore, allowing the third harmonic to flow between the phase sets. This third harmonic, however, does not have an impact on the torque as the triplen harmonics of the current do not contribute to torque production.
Figure 3-12 Experimental results for grid and DC link voltages and currents for rated power unity power factor charging.

Figure 3-13 Experimental results for grid and DC link voltages and currents for rated power unity power factor discharging.
3.2.2 Power balancing control

The unequal phase inductances of the machines cause the power processed by the inverters to be unbalanced. Unbalance in the per phase power due to the unbalanced inductance will depend on the difference of the phase-inductances within a phase-set (which is usually up to 10%) \[92\]. The cause of this unbalanced inductance is the saliency of the rotor and the winding distribution in the stator. The spatial harmonics that depend on the type of windings, also manifest themselves as an unbalance in the three-phase inductances when the rotor is at standstill.

The unbalance in the power drawn due to these unbalanced inductances cause the DC link to have a second harmonic on it. To reduce this second harmonic, the DC link capacitor must have a large value. Such large value capacitors limit the designers to certain technologies (such as electrolytic capacitors). Reduction of the second harmonic ripple can reduce the required value of the DC link capacitor and may allow the designer to opt for more efficient capacitor technologies (such as film capacitor). The power imbalance can be reduced by altering the current reference such that the power absorbed by the six phases is balanced.
In this section an interphase power balance control of both phase-sets is proposed to reduce the second order harmonic on the DC link. The control reduces the difference in power consumption at the converters’ input for each phase-set. The power consumption is balanced by measuring the power absorbed by each phase in each phase-set, $P_{1,2}$ in Figure 3-15. This value is then averaged to create a power reference, $P_{1,2ref}$ in the figure. The power feedback is then subtracted from the reference and fed through a PI controller to create the currents that are required to be added to the current references, $I_{ref}$, created by the zero-torque control. The power balance control cannot eliminate the power imbalance and only reduces it if the machine is of interior permanent magnet type or synchronous reluctance type due to the asymmetric mutual inductance components. For surface-mounted permanent magnet machines, the power balance control can achieve true power balance because the inductances provided by this kind of a machine are symmetric and equal. The proposed overall control diagram is shown in Figure 3-15.

Since the imbalance in the current references can be modeled as a linear sum of the positive and negative sequence currents, each creating a clockwise and anticlockwise torque respectively, therefore it does not affect the torque elimination strategy. Furthermore, if the residual imbalance in the currents of phase-set 1 and 2 is equal, i.e., the positive and negative sequence currents of both phase-sets are equal, the torque generated on the shaft remains eliminated. The mean of the control signals of each phase is added as the correction input to both the phase-sets to ensure that the created imbalance is equal for each phase-set, such that the correction for phase $A$ is the same as that of phase $D$ as shown in Figure 3-15. The torque control loop is as defined in chapter 2. The operating mode selection between charging and distributed resource mode is selected by the block

![Figure 3-15 Control loop for the IBC](image-url)
Ref. Sel. that passes on the current references as required. \( I_{drefDR} \) and \( I_{qrefDR} \) are calculated from the active and reactive power references that are provided by an external entity. For the battery charging mode, the current references, \( I_{drefch} \) and \( I_{qrefch} \), are calculated to control the DC link voltage. The system only controls the active power drawn from the grid in the charging mode, therefore \( I_{qrefch} = 0 \). \( I_{drefch} \) is calculated from the DC link voltage control loop. In this mode, the charging current is controlled by the DC-DC converter according to the battery type and its charging strategy.

3.2.2.1 Validation of Power Balancing Control

The power balancing control was validated on both IBC topologies presented in section 3.1. For the topology based on two machines drive, the power balance control was tested for identical and differently rated machines. The difference between the d and q axes inductances of both machines

![Simulation results for unity power factor charging at rated power with power balance control activated at 0.05s](image)

Figure 3-16 Simulation results for unity power factor charging at rated power with power balance control activated at 0.05s (a) input voltage and currents; and (b) output DC voltage and current.
were changed from 0 to 20% of their respective base inductances, and the DC link current was monitored with and without the power balancing scheme. The results are shown in Figure 3-17 (a) and (b) for identical machines and Figure 3-18 (a) and (b) for the differently rated machines. The results show that the scheme can eliminate the second harmonic on the DC link current. The reason for the second harmonic elimination in the two-machine drive is that there is no mutual coupling between the two machines and therefore the power balancing of one phase set does not have an impact on the power balancing of the other phase set.

The active power balancing strategy was also validated on the six-phase machine based IBC. The result for the grid voltage, grid and the phase \( A \) and \( D \) current and the output DC voltage and DC current for six fundamental cycles are shown in Figure 3-16. The power balancing strategy was activated after the first three cycles and the second harmonic on the DC link current is shown to be decreased by 5%. The second harmonic is not eliminated in the case of a six-phase machine due to the mutual coupling between the two-phase sets. Unlike for the two-machine based IBC, the current of one phase-set of a six-phase machine-based IBC is coupled to the current on the other phase-set, therefore, making it impossible to eliminate the second harmonic. The strategy, however, can significantly reduce the 2\(^{nd}\) harmonic.
Figure 3-17 Simulation results for the DC link second harmonic current at rated charging power versus dq-axes inductance deviation for identical three-phase SPMSMs (a) without power balancing, and (b) with power balancing.
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Figure 3-18 Simulation results for the DC link second harmonic current at rated charging power versus dq-axes inductance deviation for different rated three-phase SPMSMs (a) without power balancing, and (b) with power balancing.
3.2.3 Circulating current control

The parallel converters of the proposed IBCs with torque elimination is repeated here as Figure 3-19. The inductances $L_A \ldots L_F$ denote the winding inductances of each phase of the machine used for the IBC. For IBC’s based on the SPMSMs, the winding impedance can be very low, such that, if used individually, they won’t meet the THD requirements of the grid. However, when operated with interleaved modulation, they can meet the 5% THD standard for the grid connection of DERs. Furthermore, grid-connected converters in parallel have the zero-sequence path between them. This path does not exist in the drives’ traction mode of operation when the machine windings are configured in either double-star or double-delta. Therefore, during interleaved operation when the zero-sequence path is excited, a zero-sequence current flows. The path of this current flow between the two inverters is as shown in Figure 3-20.

Conventionally the traction drives are operated with continuous PWM. However, in order to reduce losses, these drives can be operated under dis-continuous PWM (DPWM) [77]. With such schemes, the phase leg carrying the peak current is not switched for a period corresponding to $60^\circ$ of the line cycle, thus resulting in fewer switching instances per fundamental line cycle consequently reducing switching losses at most by half. Operation with interleaved DPWM does not impose a problem in the traction mode of operation since there is no circulating current path between the two sets of three-phase windings. However, in charging mode of operation, interleaved DPWM results in circulating current that should be mitigated if the loss reduction properties of these schemes are to be capitalized on.

![Figure 3-19 Topology of a grid-connected paralleled converter setup](image-url)
This section presents the proposed circulating current suppression strategy for interleaved discontinuous PWM for parallel connected two-level voltage source converters and presents a comparison of the circulating current produced with conventional DPWM strategies without the proposed compensation. The template used for the DPWM generation for the proposed compensation strategy is shown in Figure 3-21. A short introduction to the DPWM strategy and the conventional templates used, fixed zero-vector (FZ) and fixed active-vector (FZ) DPWM, to generate pulses for the conventional DPWM, while the modeling of the circulating current for parallel connected converters in IBCs is presented in Appendix C, and the reader is encouraged to go through the summary to understand the proposed template.

Since the circulating current is generated because of the imbalance in the active and zero vectors in a switching cycle, therefore, a compensation technique is proposed that introduces an additional switching sub-cycle that is equal to the length of the time difference from the end of the last completed switching cycle to the instant where the zero-vector crossing (ZVC) boundary occurs. This additional part of the switching cycle complements the part of the switching cycle that has already occurred before the ZVC boundary. The additional part ensures balanced common mode and phase voltage difference before the start of the next switching cycle. The next switching cycle then has balanced voltage difference because it is not interrupted by a sector change. It also ensures that the phase voltage difference, $\Delta V_x$, has opposite signs at the start and the end of the clamping period. The strategy is shown in Figure 3-21 with the ZVC boundary occurrence. The compensation technique is based on the fixed zero-vector (FZ)DPWM template, henceforth called compensated FZ-DPWM (CFZ-DPWM).
The length of the resultant switching cycle is shorter than the rest of the switching cycles if the ZVC boundary occurs before the midpoint of the switching cycle and longer if the ZVC boundary occurs after the midpoint of the switching cycle. In both cases, the number of extra switching instances is not more than two per ZVC boundary crossings. The worst case of two additional switching instances is when the ZVC boundary occurs during a zero-vector, and it changes from, either 000 to 111 or vice versa. The number of switching instances added in the worst case is the least for the FA-DPWM. When compared to FA-DPWM, the additional switching pulses added by the proposed CFZ-DPWM technique range from 1 to 6 per fundamental cycle depending on where the ZVC boundary occurs in the switching cycle (which depends on the modulation index, the carrier ratio, and the clamping region).

For comparison, the instance of ZVC occurrence for conventional DPWM templates is shown in Figure 3-22. The unbalance in ΔV for all the phases and common mode voltages are highlighted (with diagonal fill). An IBC with parameters shown in Appendix E was simulated with the conventional and the proposed DPWM templates. The results are shown in Figure 3-23, Figure 3-25 and Figure 3-24 for the proposed strategy, the FZ-DPWM and the FA-DPWM respectively. The motor used for the IBC had equal inductances.

The zero-sequence (Figure 3-23b) and the phase (Figure 3-23c) circulating current for the proposed technique do not have a low-frequency component. The phase circulating current jumps are in opposite directions at the start and the end of each clamping period. The figure shows the

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Figure 3-21 Proposed DPWM strategy to complete a short switching cycle before the start of the next switching cycle.
3.2.3.2 Effectiveness of the proposed control strategy in minimizing both the circulating current components.

Simulation results for FZ-DPWM show that neither the zero-sequence (Figure 3-25b) nor the phase (Figure 3-25c) circulating current is controlled. Results for FA-DPWM show that the zero-sequence circulating current (Figure 3-24b) only consists of the high-frequency component with a zero average. The phase circulating current (Figure 3-24c), however, has a low-frequency component because of the z-sector changes. The start and the end of the clamping period do not always have opposite direction jumps, thus resulting in a low-frequency intra-phase circulating current.

3.2.3.2 Hardware validation

A scaled-down experimental setup was built to conduct hardware experiments that verify the effectiveness of the proposed circulating current compensation technique and to compare it with the already presented technique in literature. The parameters of the setup are shown in Appendix E. The experimental setup was designed to have the passive components to be the same in per unit. The system was controlled in open loop, drawing power from the grid at unity power factor. Both the converters were provided the same reference but were modulated with a 180° interleaved pulse pattern.
The results for the DC link of 37.5V, corresponding to a modulation index of 1, and a carrier ratio of 57 are shown in Figure 3-26, Figure 3-27 and Figure 3-28 for the proposed DPWM, FA-DWPM, and FZ-DPWM respectively. The proposed strategy effectively reduces the rms of the phase circulating current when compared to the FA-DPWM technique and reduces the rms of the common-mode circulating current when compared to the FZ-DPWM technique.
Similar experiments were repeated for the DC link voltage controlled at 37.5V, 43.75V and 50V, corresponding to modulation indices of 1, 0.9 and 0.8 respectively. These specific modulation indices were selected because most grid-connected converters are designed to work in the high modulation index range. The carrier ratios were selected to test the performance of all the techniques at integer (54 – where the ZVC boundary occurs at the end of the switching cycle), non-integer multiples of six (55 and 56 – where the ZVC boundary occurs during the active vector
The rms values of the common mode and the phase circulating current, as a ratio of the FZ-DPWM results are shown in Figure 3-29. The proposed technique reduces the circulating currents for both carrier ratios and the three modulation indices. The difference, however, is not as obvious for the modulation index of 1 because the zero-vector dwell times are very small and do not create a significant circulating current.

Figure 3-25 Simulation results of FZ-DPWM strategy for rated current drawn from the grid (a) \(i_{A1}, i_{A2}, i_A\); (b) \(i_{CC}\); (c) \(i_{PCA}\).

- (a) shows the phase currents for different carrier ratios.
- (b) illustrates the common mode current.
- (c) depicts the phase circulating current.
Figure 3-26 Experimental results of the proposed strategy for rated current drawn from the grid (a) $i_{A1}, i_{A2}, i_A$; (b) $i_{CC}$; (c) $i_{PCA}$.

Figure 3-27 Experimental results of the FA-DPWM strategy for rated current drawn from the grid (a) $i_{A1}, i_{A2}, i_A$; (b) $i_{CC}$; (c) $i_{PCA}$.
3.3 Conclusion

This chapter proposed two IBC topologies that can be used to implement the torque control strategies proposed in Chapter 2. One topology was based on traction drives with two three-phase synchronous machines, while the other topology was based on traction drives on six-phase synchronous machines. The modeling for the IBC topologies was presented and the controls that address the issues that arise with topology implementation were also presented and validated. It was shown that unbalanced inductances of the machine windings disallow the use of rotating reference frames to control the IBC and therefore, stationary frame control allowing individual phase current control, like the linear quadratic regulator with integral (LQRI) action, needs to be employed. The power balancing control that reduces the second harmonic oscillation on the DC link due to the unbalance in the input power of the three phases at the input was also presented. In addition, an interleaved modulation strategy capable of controlling the circulating current was also presented along with its validation on hardware. It was shown that the control strategies successfully achieve their objectives.
Figure 3-29 Experimental results for carrier ratios showing $i_{CC}$ and $i_{PC}$ for DC link voltages of (a),(b) 450V; (c),(d) 400V; and (e),(f) 350V.
4 Integrated Battery Charger Analysis and Design

The control strategies presented in the previous chapter help meet the challenges that arise when the torque cancellation strategies are applied on the six-phase machines. However, once the control strategies are implemented the ratings of battery charger need to be defined respecting not only the ratings of the traction inverter but also the machine ratings. This chapter presents an analysis of the effects of charging operation on a six-phase machine with windings having various phase displacements between their two three-phase sets. The limits on the current through the windings to define the maximum power rating of the charging operation for a given six-phase machine are then theoretically derived and then are validated using FEA models of a IPMSM used for a commercially available passenger EV [99]. A discussion on the machine windings inductance and its impact on the charging operation is also included. The analyses were done on an FEA model of a six-phase traction motor.

4.1 Flux and Force Distribution during Torque Elimination

The torque cancellation strategy reduces the torsional stress on the rotor shaft by ensuring equal and opposite torque generation by the two three-phase sets. This cancellation results in a force distribution in the air gap that may stress the stator teeth and rotor surface. Furthermore, with the torque elimination strategy the torque is cancelled at a macroscopic level, the magnetic field produced by the windings still rotates in the air gap at the fundamental frequency. This rotating field needs to be analyzed to ensure that the magnetic field density with torque cancellation does not increase beyond the rated magnetic field density of the machine. If the flux density is allowed to increase beyond the rated flux of the machine, it will result in stator saturation, additional losses, and consequently, greater heating that requires further cooling effort. It should be pointed out that the conventional cooling apparatus for electric vehicle drives, esp. forced air cooling, relies on the rotor movement for effective cooling. A magnet temperature estimation algorithm [100, 101] can also be used to derate the charging power as a function of the magnet temperature. Further research would be required to ensure cooling with a rotor at a standstill.
4.1.1 Flux density distribution

This section analytically examines the flux and force distribution inside a six-phase synchronous machine during the operation of an IBC and compares the flux density and force density distribution for full-pitched and short-pitched windings with $\alpha = 0^\circ, 30^\circ, 60^\circ$ using FE analysis. The flux distribution of the machine without excitation is shown in Figure 4-1 and the flux density distribution in the middle of the air gap is shown in Figure 4-2. Where $B_n$ and $B_t$ denotes the normal (or radial) and tangential flux density respectively. It is worth mentioning that the unexcited flux density distribution is created only by the magnets. In the figure, the radial component of the flux density, $B_n$, is significantly larger than its tangential component, $B_t$. We can also see that $B_n$ has dips at points of the stator slot opening, where $B_t$ has spikes at those same points. It is as

Figure 4-1 Flux density for the machine without current excitation.
expected, since the reluctance to the normal flux density in the slot is higher than that in the stator tooth. Similarly, the tangential flux exists at the tooth edges.

When the windings are excited, the total radial and tangential flux densities in the air gap of the machine can be defined by (4-1).

\[
B_n = B_{ns} + B_{nPm} \quad (4-1)
\]
\[
B_t = B_{ts} + B_{tPM} \quad (4-2)
\]

where \(B_{ns}\) is the total normal flux density in the air gap, \(B_{ts}\) is the normal flux density due to the stator excitation and \(B_{nPm}\) is the normal flux density due to the permanent magnets on the rotor. The flux density due to the stator windings is the sum of the individual flux densities produced by the phase windings (a-f) in the stator and can be written as (4-3).

\[
B_{ns} = B_{na} + B_{nb} + B_{nc} + B_{nd} + B_{ne} + B_{nf} \quad (4-3)
\]

Whereby, the flux produced by each winding is given by

\[
B_{na} = i_a B_1 \left( \phi + \frac{\alpha}{2} \right) \quad B_{nd} = i_d B_1 \left( \phi - \frac{\alpha}{2} \right)
\]
\[
B_{nb} = i_b B_1 \left( \phi - \frac{2\pi}{3} + \frac{\alpha}{2} \right) \quad B_{ne} = i_e B_1 \left( \phi - \frac{2\pi}{3} - \frac{\alpha}{2} \right) \quad (4-4)
\]
\[
B_{nc} = i_c B_1 \left( \phi + \frac{2\pi}{3} + \frac{\alpha}{2} \right) \quad B_{nf} = i_f B_1 \left( \phi + \frac{2\pi}{3} - \frac{\alpha}{2} \right)
\]

where \(B_1(\phi)\) is an odd function of the angular position in the air gap \(\phi\), \(i_x\) is the phase \(x\) current \((x = a, b, ..., f)\), and \(\alpha\) is the phase displacement between the two sets of three-phase windings. The function \(B_1(\phi)\) is approximated by the first fundamental and is therefore of the form given by (4-5).

\[
B_1(\phi) = B_{n0} \sin(\phi) \quad (4-5)
\]

For the traction mode of operation, the windings are excited by the currents given by (4-6)

\[
i_a = I_0 \cos(\phi_{abc}) \quad i_b = I_0 \cos \left( \phi_{abc} + \frac{2\pi}{3} \right) \quad i_c = I_0 \cos \left( \phi_{abc} - \frac{2\pi}{3} \right)
\]
\[
i_d = I_0 \cos(\phi_{def}) \quad i_e = I_0 \cos \left( \phi_{def} + \frac{2\pi}{3} \right) \quad i_f = I_0 \cos \left( \phi_{def} - \frac{2\pi}{3} \right) \quad (4-6)
\]
where $I_0$ is the rated phase current for the traction mode, $\phi_{abc} = \theta_r + \omega_r t + \frac{\alpha}{2} + \phi_{adv}$ and $\phi_{def} = \theta_r + \omega_r t - \frac{\alpha}{2} + \phi_{adv}$, $\theta_r$ is the rotor angle, $\omega_r$ is the rotor speed, and $\phi_{adv}$ is the current advance angle.

Substituting (4-5) and (4-6) in (4-4), and the results in (4-3) give us the total flux density as a function of $\phi$ in the air gap.

$$B_{ns} = 3B_{n0}I_0\cos(\alpha)\sin(\phi + \theta_r + \omega_r t + \phi_{adv})$$  \hspace{1cm} (4-7)

A similar analysis is done for the tangential flux component, with the knowledge that the tangential flux density is an even function of $\phi$, and is approximated by its first fundamental $B_2(\phi) = B_{t0}\cos(\phi)$. The total tangential flux density is therefore given by

$$B_{ts} = 3B_{t0}I_0\cos(\alpha)\cos(\phi + \theta_r + \omega_r t + \phi_{adv})$$  \hspace{1cm} (4-8)

The six-phase machine is designed such that the rotor and stator steel does not saturate when, as per (4-1), the flux densities of $B_{nP} + B_{nmax}$ and $B_{tP} + B_{tmax}$ are sustained in the air gap.

For the torque cancellation strategy, the currents in (4-6) will have $\theta_r = 0$ and $\omega_r = 0$ because the rotor is aligned to the centre of the two stator d-axes and the advance angle will be a synchronized to the grid frequency, $\omega_g$. The currents will therefore be defined by (4-9)
\[ i_a = I_{bc} \cos(\omega_g t) \quad i_b = I_{bc} \cos(\omega_g t - \frac{2\pi}{3}) \quad i_c = I_{bc} \cos(\omega_g t + \frac{2\pi}{3}) \]
\[ i_d = I_{bc} \cos(\omega_g t) \quad i_e = I_{bc} \cos(\omega_g t + \frac{2\pi}{3}) \quad i_f = I_{bc} \cos(\omega_g t - \frac{2\pi}{3}) \] (4-9)

where, \( I_{bc} \) is the rated phase current for the charging mode. Following a similar analysis as done for the traction mode, the total normal and tangential flux in the air gap is given by (4-10) and (4-11) respectively.

\[ B_{ns} = 3B_{n0} l_{bc} \sin(\theta_r + \omega_r t) \cos \left( \omega_g t + \frac{\alpha}{2} \right) \] (4-10)
\[ B_{ts} = 3B_{t0} l_{bc} \cos(\theta_r + \omega_r t) \cos \left( \omega_g t + \frac{\alpha}{2} \right) \] (4-11)

From (4-10) and (4-11) we see that the maximum values of the flux in charging mode of operation do not depend on the phase displacement alpha. The maximum charging current should, therefore, be limited by (4-12) to avoid saturation during charging operation.

\[ I_{bc} \leq I_0 \cos(\alpha) \] (4-12)
The magnetic flux densities in the air gap of a full pitched, and a short-pitched winding, with $\alpha=0^\circ$ are shown in Figure 4-3 and Figure 4-4 respectively for eight time instants ($[t_1, t_2, t_3, ..., t_8] = [0, 2.1, 4.2, ..., 14.7]$ms). The results show the flux density for one pole pair $[-\pi, \pi]$. In the figures, z axes show the flux density in Tesla, y axes show the observer position in the air gap, and the x axes show the time instants at which they were recorded. As time progresses from $t_1$ to $t_2$ the peak radial flux density ($B_r$), starts from the rotor d-axis (0) and then collapses and re-emerges at $t_8$. The two distinctive flux density peaks by each phase set can be seen at $t_3$ and $t_4$ for Figure 4-3. While in Figure 4-4 the peaks are not as distinctive for the two-phase sets. The reason is the double layer winding that was required to split the phase-belt to create the short-pitched windings.

The flux in the airgap during charging operation will produce eddy current losses in rotor magnets in addition to the iron losses in the machine stator and the copper losses in the windings. These losses will raise the temperature of all the machine components, which may result in the demagnetization of the magnets. The demagnetization of the permanent magnets used in the machine can be assessed by monitoring the magnetic field projected in the direction of the field.

Figure 4-4 Simulated FEA results for flux density distribution for short-pitched windings and $\alpha=0^\circ$ at various time instants for a grid fundamental cycle at rated charging power.
created by the magnets. If this field reduces the magnets’ flux near its knee-point, the magnet risks permanent demagnetization. The magnet’s proximity to demagnetization can be calculated as:

\[ D_p = B_{\text{demag}} - B \cdot u_{\text{mag}} \]  \hspace{1cm} (4-13)

Where the demagnetization proximity \( D_p \) defines the proximity of the magnet to demagnetization and should always be negative, the nearer the value is to 0, the more likely the demagnetization. A positive value means the magnets are irreversibly demagnetized. \( B_{\text{demag}} \) is the demagnetization flux of the magnet and is generally defined in the magnet manufacturer’s data sheet, \( B \) is the resultant flux density inside the magnet and \( u_{\text{mag}} \) defines the magnets flux direction.

The magnetization proximity for the machine with full-pitched windings and \( \alpha = 0^\circ \) is shown in Figure 4-5 for the charger operating at half and full machine ratings with torque cancellation for one grid cycle without cooling. The results show that the charger can be operated without the risk of demagnetization if the magnet temperature remains below 180°C, while for (machine) rated operation, the magnets must be kept below 140°C.
Results of a study on machines with $\alpha = 0^\circ, 30^\circ, 60^\circ$, with full pitch and short pitched windings are shown in Figure 4-6 and in Figure 4-7 respectively. The short pitch windings were short pitched by one slot. Other windings with more than one slot short pitch exist, however are uncommon, and therefore not studied in the thesis. The figures show the maximum of the demagnetization proximity seen by the magnets in one grid cycle. The results show that for full pitched windings, having $\alpha = 30^\circ$ and $\alpha = 60^\circ$ provide a slight advantage over the $\alpha = 0^\circ$ design. The short-pitched windings on the other hand, allow the machine to be operated at a higher power without risking demagnetization for the motor studied.

### 4.1.2 Force density distribution

A similar analysis was also done on the force components produced in machine air gap. Since the stator and rotor steel can endure a certain tangential force [102] (the normal force is balanced),
the maximum force components produced in the charging mode also need to be analyzed. The force components are calculated using the Maxwell stress tensor [103] and are given by (4-14) and (4-15).

\[
f_r = \frac{1}{2\mu_0} (B_n^2 - B_t^2) \tag{4-14}
\]

\[
f_t = \frac{1}{\mu_0} B_n B_t \tag{4-15}
\]

where \(f_r\) and \(f_t\) are the radial and tangential force densities (N/m²) in the air gap and \(\mu_0\) is the permeability of free space (\(4\pi \times 10^{-7}\)). The radial and tangential force density distributions of an unexcited machine in the middle of the air gap for one pole pitch is shown in Figure 4-8. Since the rotor is at rest, the total tangential forces acting is zero. This is evident by the fact that \(f_t(\phi)\) is an odd function when the zero position is considered to be the q-axis of the rotor, as shown in the figure. Similarly, the vector sum of the radial forces, \(f_r(\phi)\), is also zero validating that the rotor is at rest. The force density distributions for rated traction operation of the same machine is shown in Figure 4-9. We can now see from the figure, that the vector sum of the radial forces still results in zero, while the tangential components produce a net positive tangential force on the rotor, which results in the rated torque on the rotor.

Analytically, the tangential force density can be expanded by substituting (4-1) and (4-2) in (4-15)

\[
f_t = \frac{1}{\mu_0} (B_{ns} + B_{nPM}) (B_{ts} + B_{tPM})
\]
\[
= \frac{1}{\mu_0} (B_{ns} B_{ts} + B_{ns} B_{tPM} + B_{nPM} B_{ts} + B_{nPM} B_{tPM}) \tag{4-16}
\]

Since the fourth term in (4-16) remains constant for both traction and charging mode, therefore, only the first three terms are analyzed. For traction mode, we substitute (4-7) and (4-8) in (4-16)

\[
f_t = \frac{1}{\mu_0} \left( \frac{9}{2} B_{n0} B_{t0} l_0^2 \cos^2(\alpha) \sin(2\phi + 2\theta_r + 2\omega_r t + 2\phi_{adv}) 
+ 3 I_0 \cos(\alpha) [B_{tPM} B_{n0} \sin(\phi + \theta_r + \omega_r t + \phi_{adv}) 
+ B_{nPM} B_{t0} \cos(\phi + \theta_r + \omega_r t + \phi_{adv})] + B_{nPM} B_{tPM} \right) \tag{4-17}
\]

Similarly, for IBC operation mode we substitute (4-10) and (4-11) in (4-16)
Comparing (4-17) and (4-18) we see that to respect the rated tangential force on the rotor, \( I_{bc} \leq I_0 \cos(\alpha) \) (derived by comparing individual terms), a limit similar to the one found for magnetic flux density.

The force density distribution for the FE models of the motor were calculated and are shown in Figure 4-10 and Figure 4-11 for \( \alpha = 0^\circ \), full-pitched and short-pitched windings respectively. The results show a similar trend as with the flux density distribution. The radial force density components are periodic per pole and therefore result in a net zero radial force on the rotor. The tangential forces on the other hand, if integrated from \(-\pi \rightarrow \pi\), equal to zero, i.e. the total torque generating forces produced on the rotor are cancelled by the torque cancelation strategy.

The peak radial and tangential force density distribution, normalized to the rated force density components, are shown in Figure 4-12 (b-c) for half rated and in Figure 4-13 (b-c) for full rated charging respectively. The force distribution for both radial and tangential components are well below the value for which the machine is designed for half rated operation, for full rated operation,
however, the maximum radial force seen by the rotor is larger than the rated force for the machine. The same is true for the tangential force density, as predicted by (4-18). The selection of $\alpha$ does not significantly impact the maximum flux or force distribution in the machine air gap. The only impact is on the maximum tangential force for when $\alpha=30^\circ$, which is expected as for $\alpha=30^\circ$ the average torque reduces along with the torque ripple. Additionally, the results for short-pitched windings show that the tangential force is below the rated force even for full rated charging, this is due to the phase flux cancellation since the phase windings are double layered and share slots with other phases.

### 4.2 Winding Inductance

The machine winding inductance is conventionally designed to be low for the traction mode of operation. For the charging mode of operation, however, it is required to be enough to keep the total harmonic injected into the grid under the limits set by the standards [104, 105] and reduce the circulating current that will flow due to the parallel connection of the phase windings in both the IBC topologies. The inductances of the machine must be measured to ascertain if they are enough to meet the requirements above.

#### 4.2.1 Total harmonic distortion

The inductance required for the charger should be provided by the machine windings to meet the total harmonic distortion (THD) limits. If the inductance provided by the windings are not enough, the parallel converters can be interleaved to reduce the THD. Furthermore, motors with the stator geometry or winding type (such as stranded with long end winding loops) that produces a larger leakage inductance can also be used [106] to provide an inductance large enough to meet the THD requirements at the given power level.

If the total inductance provided is still not enough, the THD can be further reduced by additional inductance or a filter capacitor or a combination to fashion a L-type, LC-type or LCL-type [107] input filter at the grid side of the contactors, such that the added components do not interfere with the traction mode of operation.

#### 4.2.2 Circulating current

The parallel topology of the integrated battery chargers will allow the circulating current to flow between them. This current can be common mode or differential, in both cases it can reduced by
either adding common mode and differential mode inductance [108] or by specialized modulation techniques [109] or a combination of both [110].

4.3 Conclusion

This section presented an analysis and study on the impact of the traction motor designed ratings on the charger ratings. The impact of charging on the machine was also studied concerning the magnet demagnetization, flux density and force density distribution in the air gap. It was shown that the selection of $\alpha$ does not have an impact on the maximum flux density distribution. The kind of winding, full-pitched or short-pitched however, does impact the maximum flux density distribution and is lower for the short-pitched windings. The maximum force density distribution is smaller for $\alpha = 30^\circ$ compared to other values and the short-pitched windings allow the force distribution to remain below the rated even when the charger is operated at 1 pu of the machine.

![Simulated FEA results for force density distribution in the airgap of the six-phase machine with $\alpha = 0^\circ$ and full-pitched windings at rated traction operation](image.png)
rating. This operation will, however, be limited by the demagnetization of the magnets on the rotor as shown by the demagnetization study. Impacts of the inductance on the charger operation were also discussed. The compared windings were short pitched by one slot.

Figure 4-10 Simulated FEA results for force density distribution for full-pitched windings and $\alpha=0^\circ$ at various time instants for rated charging operation.
Figure 4-11 Simulated FEA results for force density distribution for short-pitched windings and $\alpha=0^\circ$ at various time instants for rated charging operation.
Figure 4-12 Simulated FEA results for maximum value comparisons of various winding types for (a) maximum flux density; (b) radial force density; (c) tangential force density when operated at 0.5pu of machine rated power.

Figure 4-13 Simulated FEA results for maximum value comparisons of various winding types for (a) maximum flux density; (b) radial force density; (c) tangential force density when operated at rated machine power.
5 Summary and Conclusion

5.1 Summary

Battery chargers for electric vehicles should be designed for convenience such that they can be put on board of the vehicle resulting in a low power charger that takes longer to charge the battery. High power chargers charge the battery much faster but must be placed off-board of the vehicle due to their size and weight and are costly. Integrated battery chargers provide a good compromise between convenience and speed of charging by re-purposing traction drive components that are present in the vehicle. The re-purposing can be partial, i.e., only the traction inverter or the traction motor windings are used. A total re-purposing, i.e., when the traction motor and inverter are both re-purposed, is more beneficial regarding weight, size and component count reduction for the charger.

The thesis focuses on addressing the issues that arise with the use of synchronous machine windings as filter inductances of an integrated battery charger. Since the charging process excites the windings, the flux generated by the machine windings creates torque on the machine shaft. The nature of this torque depends on the machine topology. The thesis analyzes the nature of the torque produced on the shaft of three-phase and six-phase synchronous machines. The analyzed machine technologies include synchronous reluctance and permanent magnet (surface mounted and interior) synchronous machines. The torque model for the machines was analytically developed and was validated against finite element (FE) models of the machines. The analytical torque models were then used to develop the torque cancellation strategies and then were validated on the same FE models.

The traction drive topologies that implement the torque cancellation strategies were also proposed. The model of the traction motor as filter inductances of the battery charging rectifier formed by the drive inverter was also presented and the additional control strategies that address the issues that arise by the charger topology were presented. A power balancing strategy to reduce the second harmonic oscillation on the DC link output of the system due to the unequal rotor position-dependent windings of the motors was proposed. Additionally, a modulation strategy was also proposed to reduce the circulating current that otherwise flows when the charger is operated with interleaved space vector modulation.
The impact of using synchronous machine windings as filter inductances of integrated battery charger was also considered. The effects of the winding type and interphase-set displacement on the flux density distribution, and force density distribution were studied. Based on the studied results recommendations for the machine design for use in integrated battery chargers were derived.

5.2 Conclusions

The integrated battery chargers (IBC) proposed in the thesis provide a fast charging (greater than 20kW) solution. The IBC solution is lighter and smaller when compared to external battery chargers that are solely built to charge the battery. The efficiency of IBCs may not be the same as the dedicated battery chargers due to the additional sources of losses (contactors and rotor magnets), however, the drive can be optimized in the design phase to reduce these losses. The efficiencies of the integrated battery chargers were not formally studied in this thesis since the lab prototypes do not represent the designed system.

5.2.1 Synchronous machine based integrated battery chargers

A torque cancellation strategy is proposed for three-phase integrated battery chargers based on three-phase and six-phase synchronous machines. The strategy allows integration of the battery charger into a traction drive that uses multiple three-phase or six-phase synchronous machines by eliminating the pulsating torque on the rotor that results when the machine windings are excited with a stationary rotor. The strategy was shown to successfully eliminate the pulsating torque for PMSM, IPMSM, and SRM based traction drives. The strategy was shown to be effective despite the displacement of the rotor from the effective stator magnetic axis. It was shown that the effectiveness of the strategy depends on the accuracy of the rotor position sensed and on the stator saturation if the charger current is rated to be as high as the machine rated current.

5.2.2 Integrated battery charger topologies and control

The torque cancellation strategy requires the two three-phase sets to be connected in parallel. The windings inductances are a function of the rotor position causing the charger to absorb power that is not balanced, resulting in a second-order oscillation on the DC link output. A power balancing control was proposed that ensures that each three phase-set absorbs the same power as
the other. The proposed control was shown to eliminate the second harmonic on the DC link for chargers that used the windings of SMPMs, and reduce the second harmonic on the DC link for the chargers using IPMSMs and SRMs.

A modulation strategy based on discontinuous PWM (DPWM) is proposed to reduce the circulating current between the converters. The circulating current mitigation strategy was shown to effectively reduce the common mode and interphase circulating when compared to the conventional DPWM strategies for switching frequencies that are non-integral multiples of the fundamental as well. The strategy helps avoid or reduce the size of an additional common mode or differential mode filter.

5.2.3 Integrated Battery Charger Analysis and Design

The impacts on the six-phase synchronous machine when operated as part of an integrated battery charger was studied using finite element analysis. The studied impacts included flux density and force density distribution in the machine airgap during the charging operation with torque cancellation. The analysis shows that the flux density distribution causing stator saturation is the deciding factor of the current limit of the charging operation. It was shown that the charging current should be less than the rated current of the machine. It was also shown that to avoid saturation a full pitch winding is a better choice than a short pitch (by one slot) winding for a same rated six-phase synchronous motor. Several design recommendations based on the inductance provided by the motor windings were also provided.

5.3 Recommendations for Further Work

This thesis has presented torque cancellation strategies based on three-phase and six-phase synchronous machine based integrated battery chargers. The impact of charging on six-phase synchronous machines has also been studied. Some directions of research that can be further pursued in this area are as follows:

- Synchronous machines with a higher number of phases (than six) can be studied, and torque cancellation strategies for such machine can be investigated.
- Core losses can be investigated using finite element analysis and the hardware experiments for charging mode of operation. The thermal model of the machine for the charging mode of operation will be different from the traction mode of operation and therefore can be studied.
- The motor design optimization process that includes the integrated battery charger design can be researched. Combined rules for the winding inductance design, ferromagnetic material selection, and motor sizing can be provided such that the machine meets the design requirements for both, the traction and charging mode of operation.
- The thesis provides the basis of electromagnetic impact on the synchronous machine. However, the mechanical vibrational impact of charging on the machines can be further studied. Furthermore, the impact of charging on the lifetime of the machine should also be studied. The production of acoustic noise should also be analyzed.
Appendix A

Several machine designs were generated using a commercially available tool to generate synchronous machine designs based on design parameters. Table A-1 shows the machine designs used for the validation of the torque cancellation strategy using two three-phase machines. Three-phase machines were modeled to have similar characteristics as a commercially available machine by a 37kW traction drive motor for passenger EVs. While the second set of machines were modeled to have similar characteristics as of a 100kW six-phase motor for light and medium duty EVs from the same manufacturer but rewound to have three-phase winding.

Table A-1 Three-phase machine parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value (SI Units)</th>
<th>Value (SI Units)</th>
<th>Value (SI Units)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power</td>
<td>$P_{nom}$</td>
<td>37kW</td>
<td>30kW</td>
<td>33kW</td>
</tr>
<tr>
<td>Voltage Rating</td>
<td>$V_{nom}$</td>
<td>400V</td>
<td>650V</td>
<td>450V</td>
</tr>
<tr>
<td>Inductance</td>
<td>$L_{d}/L_{q}$</td>
<td>330μH / 330μH</td>
<td>1.44mH / 3.28mH</td>
<td>0.585mH / 2.02mH</td>
</tr>
<tr>
<td>Poles Pairs</td>
<td>$p$</td>
<td>5</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Magnet flux</td>
<td>$\lambda_{pm}$</td>
<td>0.0864 V.s</td>
<td>0.1398 V.s</td>
<td>-</td>
</tr>
<tr>
<td>Rated Speed$^1$</td>
<td>$\omega_{nom}$</td>
<td>2200 rpm</td>
<td>3000 rpm</td>
<td>800 rpm</td>
</tr>
</tbody>
</table>

Table A-2 shows the machines that were also modeled to have similar characteristics as a commercially available 100kW six-phase PMSM, the characteristics of which are shown in Table

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value (SI Units)</th>
<th>Value (SI Units)</th>
<th>Value (SI Units)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power</td>
<td>$P_{nom}$</td>
<td>100kW</td>
<td>100kW</td>
<td>100kW</td>
</tr>
<tr>
<td>Voltage Rating</td>
<td>$V_{nom}$</td>
<td>400V</td>
<td>650V</td>
<td>450V</td>
</tr>
<tr>
<td>Inductance</td>
<td>$L_{d}/L_{q}$</td>
<td>800μH / 800μH</td>
<td>1mH / 1.5mH</td>
<td>0.75mH / 1.4mH</td>
</tr>
<tr>
<td>Poles Pairs</td>
<td>$p$</td>
<td>8</td>
<td>8</td>
<td>8</td>
</tr>
<tr>
<td>Magnet flux</td>
<td>$\lambda_{pm}$</td>
<td>0.3446 V.s</td>
<td>0.5599 V.s</td>
<td>-</td>
</tr>
<tr>
<td>Rated Speed</td>
<td>$\omega_{nom}$</td>
<td>800 rpm</td>
<td>800 rpm</td>
<td>800 rpm</td>
</tr>
</tbody>
</table>

$^1$ The rated speeds reported are the maximum speed achievable before the flux weakening region.
A-3. It was also used in the experimental validation as well. Table A-4 shows the parameters of the traction motor of a commercially available EV rewound as a six-phase machine and was used for the analysis in Chapter 4.

Table A-2 Six-phase high torque direct-drive machine parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value (SI Units)</th>
<th>SPMSM</th>
<th>Value (SI Units)</th>
<th>IPMSM</th>
<th>Value (SI Units)</th>
<th>SRM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power</td>
<td>$P_{nom}$</td>
<td>100 kW</td>
<td></td>
<td>100 kW</td>
<td></td>
<td>100 kW</td>
<td></td>
</tr>
<tr>
<td>Voltage Rating</td>
<td>$V_{nom}$</td>
<td>450 V</td>
<td></td>
<td>450 V</td>
<td></td>
<td>400 V</td>
<td></td>
</tr>
<tr>
<td>Inductance</td>
<td>$L_{ds}/L_{qs}$</td>
<td>960 μH / 960 μH</td>
<td></td>
<td>0.69 mH / 1.57 mH</td>
<td></td>
<td>7.19 mH / 9.93 mH</td>
<td></td>
</tr>
<tr>
<td>Poles Pairs</td>
<td>$p$</td>
<td>8</td>
<td></td>
<td>4</td>
<td></td>
<td>8</td>
<td></td>
</tr>
<tr>
<td>Rated Speed</td>
<td>$\omega_{nom}$</td>
<td>800 rpm</td>
<td></td>
<td>800 rpm</td>
<td></td>
<td>800 rpm</td>
<td></td>
</tr>
</tbody>
</table>

The parameters mentioned in Table A-1 and Table A-2 were measured from their respective FEA models by using the flux linkage method [111] with the rotor at standstill for various positions of the rotor corresponding to electrical angles of 0-2$\pi$ rad. The assumptions for the measurements were that the machine was not saturated, and the temperature of operation was 20°C.

Table A-3 Six-phase high torque direct-drive IPMSM used for experimental validation

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value (SI Units)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power</td>
<td>$P_{nom}$</td>
<td>100 kW</td>
</tr>
<tr>
<td>Voltage Rating</td>
<td>$V_{nom}$</td>
<td>450 V</td>
</tr>
<tr>
<td>Poles Pairs</td>
<td>$p$</td>
<td>8</td>
</tr>
<tr>
<td>Rated Speed</td>
<td>$\omega_{nom}$</td>
<td>1000 rpm</td>
</tr>
</tbody>
</table>

Table A-4 Traction motor parameters for a commercial passenger EV

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value (SI Units)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power</td>
<td>$P_{nom}$</td>
<td>60 kW</td>
</tr>
<tr>
<td>Voltage Rating</td>
<td>$V_{nom}$</td>
<td>450 V</td>
</tr>
<tr>
<td>Poles Pairs</td>
<td>$p$</td>
<td>8</td>
</tr>
<tr>
<td>Rated Speed</td>
<td>$\omega_{nom}$</td>
<td>3000 rpm</td>
</tr>
</tbody>
</table>
The parameters for the machines presented in Table A-3 and Table A-4 were measured by locking the rotor at standstill and each winding was excited with a low voltage at the fundamental frequency and the current was allowed to reach the rated charging value. The impedance was then calculated from the voltage and current phasors. The assumptions here too are that the machine is not saturated at the rated charging current.
Appendix B

B.1 Synchronous Machine Modeling

B1.1 Modeling of the three-phase synchronous machines

The relation between the rotor d-q axes and the stator ABC axes for a three-phase synchronous machine is shown in Figure B-1. The rotor speed is \( \omega_r \) while the stator current vector rotates with a frequency of \( \omega_s \) and the initial angle between the stator current vector and rotor d-axis is \( \theta_r \). The stator quantities (voltages, currents and flux-linkages) are converted to the rotor frame of reference using Park’s transformation, \( T(\theta) \) such that:

\[
f_{dq} = T(\theta) \left[ f_{abc} \right]
\]

where, \( f_{dq} \) represents the variables in the rotating rotor d-q frame of reference and \( f_{abc} \) represents the variables in the stationary stator abc frame of reference. The angle \( \theta \) used for the transformation is given as \( \theta = (\omega_s - \omega_r)t + \theta_r \).

The three-phase machine windings are modeled as three-phase inductances which are dependent on the rotor position \( \theta_r \):

\[
L_{abc}(\theta_r) =
\begin{bmatrix}
L_0 + L_v \cos(2\theta_r) + L_l & -\frac{L_0}{2} + L_v \cos\left(2\theta_r - \frac{2\pi}{3}\right) & -\frac{L_0}{2} + L_v \cos\left(2\theta_r + \frac{2\pi}{3}\right) \\
-\frac{L_0}{2} + L_v \cos\left(2\theta_r - \frac{2\pi}{3}\right) & L_0 + L_v \cos\left(2\theta_r + \frac{2\pi}{3}\right) + L_l & -\frac{L_0}{2} + L_v \cos(2\theta_r) \\
-\frac{L_0}{2} + L_v \cos\left(2\theta_r + \frac{2\pi}{3}\right) & -\frac{L_0}{2} + L_v \cos(2\theta_r) & L_0 + L_v \cos\left(2\theta_r - \frac{2\pi}{3}\right)
\end{bmatrix}
\]

where, \( L_0 \) is the average self-inductance of each phase, \( L_v \) is the difference between the maximum and the average inductance, and \( L_l \) is the leakage inductance. Using the dq-transformation, this matrix is transformed into a constant matrix in the rotating rotor d-q frame of reference:

\[
L_{dqs} =
\begin{bmatrix}
L_{ds} & 0 & 0 \\
0 & L_{qs} & 0 \\
0 & 0 & L_l
\end{bmatrix}
= \begin{bmatrix}
L_l + \frac{3}{2} (L_0 + L_v) & 0 & 0 \\
0 & L_l + \frac{3}{2} (L_0 - L_v) & 0 \\
0 & 0 & L_l
\end{bmatrix}
\]

Conventionally, the terminal voltage of a three-phase SPMSM in the rotor frame of reference can be written as

\[
\begin{align*}
v_{qs} &= r_s i_{qs} + \omega_r L_{ds} i_{ds} + \frac{d}{dt} L_{qs} i_{qs} + \omega_r \lambda_{pm} \\
v_{ds} &= r_s i_{ds} - \omega_r L_{qs} i_{qs} + \frac{d}{dt} L_{ds} i_{ds} \\
v_{0s} &= r_s i_{0s} + \frac{d}{dt} L_{ls} i_{0s}
\end{align*}
\]

(B-4)
where $v_{dqs}, i_{dqs}$ and $L_{dqs}$ are the d-q stator voltages, stator currents and stator winding inductances respectively, $r_s$ is the stator winding resistance, $\omega_r$ is the rotor speed in electrical rad/s, $\lambda_{pm}$ is the permanent magnet flux in V/s, $v_{0s}, i_{0s}$ and $L_{ls}$ are the zero-sequence stator voltage, zero-sequence stator current and the leakage inductance of the three-phase machine windings. From (B-4) and (B-3) the electromagnetic torque of the synchronous machine is:

$$\tau_{em} = \frac{3}{2} p \left[ \lambda_{pm} i_{qs} + (L_{ds} - L_{qs}) i_{qs} \right]$$

(B-5)

where $p$ is the number of pole pairs and $\tau_{em}$ is the electromagnetic torque generated on the shaft of the machine.

The mechanical model of the machine is given by:

$$\tau_{em} - \tau_L = \frac{1}{p} \left( J \frac{d\omega_r}{dt} + B \omega_r \right)$$

(B-6)

where $\tau_L$ is the load torque in Nm, $J$ is the rotor inertia (including the mechanical drive train) in $kg \, m^2$, and $B$ is the damping coefficient of the drive train in $Nms$. Eq. (B-5) provides the general torque equation of the three kinds of three-phase synchronous machines analyzed in this thesis.

Equations (B-2)-(B-6) provide the complete machine model for a three-phase machine when operated as an electro-mechanical energy conversion transducer [112]. However, in the integrated battery charging mode, the machine is not used as an electro-mechanical energy conversion transducer, but as an inductance, with the electro-mechanical energy conversion as a side-effect. Therefore, the reviewed modeling in this section provides us the basis for the analysis of the torque production in the machine when the machine windings are used for their inductance in an IBC.
B1.2 Modeling of the six-phase synchronous machines

This section presents the modeling of a six-phase synchronous machine and demonstrates the torque production on the machine shaft when its windings are used as a charger. The six-phase machine can be looked at as a duplex machine such that the phases ABC and DEF form two phase-sets each having 120° phase displacement between the phases within the phase-set. The phase-sets (ABC and DEF) can be collectively displaced from each other by an arbitrary value, denoted by α in this thesis. Commonly six phase machines are designed for α = 0°, 30° and 60° to achieve various advantages like torque ripple reduction [91] and modularity [92].

The relation between the rotor d-q axes and the stator ABC and DEF axes for a six-phase synchronous machine is shown in Figure B-2. The rotor rotates at the speed ω_r while the stator current vector rotates with a frequency of ω_s and the initial angle between the stator current vector of the phase-set ABC and rotor d-axis is θ_r1, and the initial angle between the stator current vector of the phase-set DEF and the rotor d-axis is θ_r2. The stator quantities (voltages, currents and flux-linkages) are converted to the rotor frame of reference using a transformation, $T_6(\theta_r, \alpha)$ that converts the phase-sets ABC and DEF into two sets of decoupled d-q axes such that:

$$\begin{bmatrix} f_{dq1} \\ f_{dq2} \end{bmatrix} = \begin{bmatrix} T_1(\theta_1) & T_2(\theta_2) \end{bmatrix} \begin{bmatrix} f_{abc} \\ f_{def} \end{bmatrix} \tag{B-7}$$

where, $f_{dq}$ represents the variables in the rotating rotor d-q frame of reference and $f_{abc}$ (and $f_{def}$) represents the variables in the stationary stator abc (and def) frame of reference. The angles used for the transformation are given as $\theta_1 = (\omega_s - \omega_r)t + \theta_r1$ and $\theta_2 = (\omega_s - \omega_r)t + \theta_r2$.

The machine windings of a six-phase machine are modeled as two sets two three-phase inductances which are coupled to each other and are also dependent on the rotor position θ_r. The amount of coupling between the phases of each phase set depends on the displacement between the two sets of three-phase windings, the type of winding and the machine design.

$$L_{abcdef_s}(\theta_r, \alpha) = \begin{bmatrix} L_{abcS}(\theta_r, \alpha) & L_{m12}(\theta_r, \alpha) \\ L_{m21}(\theta_r, \alpha) & L_{defS}(\theta_r, \alpha) \end{bmatrix} \tag{B-8}$$

where $L_{abcS}(\theta_r)$ and $L_{defS}(\theta_r)$ are the 3x3 inductance matrices defining the phase self-inductance and the mutual inductance of the phases within their own phase-set, and $L_{m12}(\theta_r)$ and $L_{m21}(\theta_r)$ are 3x3 matrices defining the mutual inductance of the phases between the phase-sets. Each of the matrix elements are defined by (B-14)-(B-17). Using the d-q transformation as given by (B-7), this matrix is transformed into a constant matrix in the rotating rotor d-q frame of reference.
\[
\begin{bmatrix}
L_{dq1} = \\
L_{dq2} = \\
L_{mdq12} =
\end{bmatrix}
= \begin{bmatrix}
L_{dq1} & L_{mdq12} \\
L_{mdq12} & L_{dq2}
\end{bmatrix}
\]

\[
L_{dq1} = \begin{bmatrix}
L_{d1} & 0 & 0 \\
0 & L_{q1} & 0 \\
0 & 0 & L_{l1}
\end{bmatrix}
= \frac{3}{2} \begin{bmatrix}
(L_0 - L_v) + L_l & 0 & 0 \\
0 & (L_0 - L_v) + L_l & 0 \\
0 & 0 & L_l
\end{bmatrix}
\]

\[
L_{dq2} = \begin{bmatrix}
L_{d2} & 0 & 0 \\
0 & L_{q2} & 0 \\
0 & 0 & L_{l2}
\end{bmatrix}
= \frac{3}{2} \begin{bmatrix}
(L_0 - L_v) + L_l & 0 & 0 \\
0 & (L_0 - L_v) + L_l & 0 \\
0 & 0 & L_l
\end{bmatrix}
\]

\[
L_{mdq12} = \begin{bmatrix}
L_{md12} & 0 & 0 \\
0 & L_{mq12} & 0 \\
0 & 0 & 0
\end{bmatrix}
= \frac{3}{2} \begin{bmatrix}
(L_{m0} - L_{m2}) & 0 & 0 \\
0 & (L_{m0} + L_{m2}) & 0 \\
0 & 0 & 0
\end{bmatrix}
\]

Conventionally, the terminal voltage of a three-phase SPMSM in the rotor frame of reference can be written as:

\[
v_{q1} = r_s i_{q1} + \omega_r (L_{d1} i_{d1} + L_{md12} i_{d2}) + \frac{d}{dt} (L_{q1} i_{q1} + L_{mq12} i_{q2}) + \omega_r \lambda_{pm1}
\]

\[
v_{d1} = r_s i_{d1} - \omega_r (L_{q1} i_{q1} + L_{mq12} i_{q2}) + \frac{d}{dt} (L_{d1} i_{d1} + L_{md12} i_{d2})
\]

\[
v_{01} = r_s i_{01} + \frac{d}{dt} L_l i_{01}
\]

\[
v_{q2} = r_s i_{q2} + \omega_r (L_{d2} i_{d2} + L_{md21} i_{d1}) + \frac{d}{dt} (L_{q2} i_{q2} + L_{mq21} i_{q1}) + \omega_r \lambda_{pm2}
\]

\[
v_{d2} = r_s i_{d2} - \omega_r (L_{q2} i_{q2} + L_{mq21} i_{q1}) + \frac{d}{dt} (L_{d2} i_{d2} + L_{md21} i_{d1})
\]

\[
v_{02} = r_s i_{02} + \frac{d}{dt} L_l i_{02}
\]

where \(v_{dq12}, i_{dq12}\) and \(L_{dq12}\) are the d-q stator voltages, stator currents and stator winding inductances respectively of the phase-sets 1 and 2, \(r_s\) is the stator winding resistance, \(\omega_r\) is the rotor speed in electrical rad/s, \(\lambda_{pm12}\) is the permanent magnet flux in V/s projected on the stator d-axes of phase-sets 1 and 2, \(v_{012}, i_{012}\) and \(L_{l12}\) are the zero-sequence voltage, zero-sequence current and the leakage inductance of the phase-sets 1 and 2 of the six-phase machine windings. Additionally, due to the nature of the transformation given by (B-7), \(L_{md12} = L_{md21} = L_{md}\) and \(L_{mq12} = L_{mq21} = L_{mq}\).
The torque for these machines can be calculated by equating the mechanical and electrical power i.e.

\[ \tau_{em} = \frac{3}{2} \frac{\omega_r}{p} \left( v_{q1} i_{q1} + v_{d1} i_{d1} + 2 v_{q1} i_{q1} + v_{q2} i_{q2} + v_{d2} i_{d2} + 2 v_{q2} i_{q2} \right) \]  \hspace{1cm} (B-12)

where \( p \) is the number of pole pairs and \( \tau_{em} \) is the electromagnetic torque generated on the shaft of the machine. Substituting (B-10) and (B-11) in (B-12) and removing the loss and energy storage components we get the electromagnetic torque of the synchronous machine as:

\[ \tau_{em} = \frac{3}{2} \cdot p \left[ \lambda_{pm1} i_{q1} + (L_{d1} - L_{q1}) i_{q1} + L_{md12} i_{d1} - L_{mq12} i_{d1} i_{q2} \right] \]  \hspace{1cm} (B-13)

The mechanical model of the machine is the same as given by (B-6). Eq. (B-13) provides the general torque equation of the three kinds of six-phase synchronous machines analyzed in this thesis and Eq. (B-6). Equations (B-8)-(B-13) provide the complete machine model for a six-phase machine when operated as an electro-mechanical energy conversion transducer. However, in the integrated battery charging mode, the machine is not used as an electro-mechanical energy conversion transducer, but as an inductance, with the electro-mechanical energy conversion as a side-effect. Therefore, the six-phase machine modeling in this section provides us the basis for the analysis of the torque production in the machine when the machine windings are used for their inductance in an IBC.

---

Figure B-2 Six-phase synchronous machine model in the d-q reference frame.
Inductance matrix of a six-phase machine with rotor position dependence up to the second space harmonic.

$$L_{abc_s}(\theta_r, \alpha)$$

\[
\begin{bmatrix}
L_0 + L_v \cos(2\theta_r - 2\alpha) + L_l & -\frac{L_0}{2} + L_v \cos \left(2\theta_r - \frac{2\pi}{3} - 2\alpha\right) & -\frac{L_0}{2} + L_v \cos \left(2\theta_r + \frac{2\pi}{3} - 2\alpha\right) \\
-\frac{L_0}{2} + L_v \cos \left(2\theta_r - \frac{2\pi}{3} - 2\alpha\right) & L_0 + L_v \cos \left(2\theta_r + \frac{2\pi}{3} - 2\alpha\right) + L_l & -\frac{L_0}{2} + L_v \cos(2\theta_r - 2\alpha) \\
-\frac{L_0}{2} + L_v \cos \left(2\theta_r + \frac{2\pi}{3} - 2\alpha\right) & -\frac{L_0}{2} + L_v \cos(2\theta_r - 2\alpha) & L_0 + L_v \cos \left(2\theta_r - \frac{2\pi}{3} - 2\alpha\right) + L_l
\end{bmatrix}
\] (B-14)

$$L_{def_s}(\theta_r, \alpha)$$

\[
\begin{bmatrix}
L_0 + L_v \cos(2\theta_r + 2\alpha) + L_l & -\frac{L_0}{2} + L_v \cos \left(2\theta_r - \frac{2\pi}{3} + 2\alpha\right) & -\frac{L_0}{2} + L_v \cos \left(2\theta_r + \frac{2\pi}{3} + 2\alpha\right) \\
-\frac{L_0}{2} + L_v \cos \left(2\theta_r - \frac{2\pi}{3} + 2\alpha\right) & L_0 + L_v \cos \left(2\theta_r + \frac{2\pi}{3} + 2\alpha\right) + L_l & -\frac{L_0}{2} + L_v \cos(2\theta_r + 2\alpha) \\
-\frac{L_0}{2} + L_v \cos \left(2\theta_r + \frac{2\pi}{3} + 2\alpha\right) & -\frac{L_0}{2} + L_v \cos(2\theta_r + 2\alpha) & L_0 + L_v \cos \left(2\theta_r - \frac{2\pi}{3} + 2\alpha\right) + L_l
\end{bmatrix}
\] (B-15)

$$L_{m12}(\theta_r, \alpha)$$

\[
\begin{bmatrix}
L_{m_0} \cos(2\alpha) + L_{m_v} \cos(2\theta_r) & L_{m_0} \cos \left(2\alpha - \frac{2\pi}{3}\right) + L_{m_v} \cos \left(2\theta_r - \frac{2\pi}{3}\right) & L_{m_0} \cos \left(2\alpha + \frac{2\pi}{3}\right) + L_{m_v} \cos \left(2\theta_r + \frac{2\pi}{3}\right) \\
L_{m_0} \cos \left(2\alpha + \frac{2\pi}{3}\right) + L_{m_v} \cos \left(2\theta_r - \frac{2\pi}{3}\right) & L_{m_0} \cos(2\alpha) + L_{m_v} \cos \left(2\theta_r + \frac{2\pi}{3}\right) & L_{m_0} \cos(2\alpha) + L_{m_v} \cos \left(2\theta_r - \frac{2\pi}{3}\right) \\
L_{m_0} \cos \left(2\alpha - \frac{2\pi}{3}\right) + L_{m_v} \cos \left(2\theta_r + \frac{2\pi}{3}\right) & L_{m_0} \cos \left(2\alpha + \frac{2\pi}{3}\right) + L_{m_v} \cos \left(2\theta_r - \frac{2\pi}{3}\right) & L_{m_0} \cos(2\alpha) + L_{m_v} \cos \left(2\theta_r + \frac{2\pi}{3}\right)
\end{bmatrix}
\] (B-16)

$$L_{m_{21}}(\theta_r, \alpha) = L_{m_{12}}^T(\theta_r, \alpha)$$

(B-17)
B.2 IBC modeling

After manipulation of the Kirchoff’s voltage equations on the six phases of the circuit shown in Figure 3-3 the circuit model can be written as

\[
\frac{d\mathbf{i}}{dt} = -\mathbf{L}_s^{-1}\mathbf{R}\mathbf{i} - \mathbf{L}_s^{-1}\mathbf{V}_i + \mathbf{L}_s^{-1}\mathbf{V}_g
\]

\[
i = [i_A \ldots i_F]^T \quad \mathbf{V}_i = [V_A \ldots V_F]^T \quad \mathbf{V}_g = [V_{A_g} \quad V_{B_g} \quad V_{C_g} \quad V_{A_g} \quad V_{C_g} \quad V_{B_g}]^T
\]

\[
\mathbf{R} = \mathbf{I}_6 \times r_s
\]

where \(\mathbf{L}_s\) is the motor inductance matrix given by (B-19), \(i_A \ldots i_F\) are the phase currents, \(V_{A_g} \ldots V_{C_g}\) are the three-phase grid voltages and \(r_s\) is the winding resistance of each phase.

\[
\mathbf{L}_s = \mathbf{L}_{abcdefs}(\theta_r, \alpha) = \begin{bmatrix}
\mathbf{L}_{abc}(\theta_r, \alpha) & \mathbf{L}_{m12}(\theta_r, \alpha) \\
\mathbf{L}_{m21}(\theta_r, \alpha) & \mathbf{L}_{def}(\theta_r, \alpha)
\end{bmatrix}
\]

(B-19)

where, the elements of the matrix in (B-19) are defined by (B-14)-(B-17). The complete system model is then given by (B-20).

\[
\frac{d\mathbf{x}}{dt} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u} + \mathbf{E}\mathbf{v}
\]

\[
\mathbf{A} = \begin{bmatrix}
-\mathbf{L}_s^{-1}\mathbf{R} & 0 \\
\mathbf{I}_6 & 0
\end{bmatrix} \quad \mathbf{B} = \begin{bmatrix}
\mathbf{L}_s^{-1} \\
0
\end{bmatrix} \quad \mathbf{E} = \begin{bmatrix}
\mathbf{L}_s^{-1} \\
0
\end{bmatrix}
\]

(B-20)

\[
\mathbf{x} = [i \int i dt]^T \quad \mathbf{u} = [\mathbf{V}_i \quad 0]^T \quad \mathbf{v} = [\mathbf{V}_g \quad 0]^T
\]

where, \(\mathbf{A}, \mathbf{B}\) and \(\mathbf{E}\) are the state space matrices, \(\mathbf{x}\) is the state vector, \(\mathbf{u}\) is the input vector and \(\mathbf{v}\) is the disturbance vector.
Appendix C

The circulating current flowing between parallel connected two-level voltage source converters (VSC) was developed and is shown in Figure C-1. The switches are modeled as state dependent voltage sources on the AC side and as state dependent current sources on the DC side. The variables $S_x$ represent the state of phase ‘$x$’ pole of the converter. Its value can either be ‘1’ or ‘-1’ for a two-level voltage sourced converter. When the converters are connected in parallel, on the AC side the loop equations for phase ‘$x$’ of converter ‘$i$’ can be written as:

$$\frac{S_{xi} V_{DC}}{V_{xi}} - V_{xN} = L_{xi} \frac{di_{xi}}{dt} + r_{xi} i_{xi}$$

where $V_{DC}$ is the DC link voltage, $V_{xN}$ is the grid phase ‘$x$’ voltage with respect to the DC link neutral $N$, $L_{xi}$ is the inductance, $i_{xi}$ is the current and $r_{xi}$ is the parasitic resistance of phase ‘$x$’ connected to converter ‘$i$’. Eq. (C-1) governs the operation of the converters with the grid. There is, however, another loop formed for each phase when the converters are connected in parallel. This loop, for phase A in Figure 3-19, exists between the points A1-A-A2. The loop equation for it is written as:

$$S_{A1} V_{DC} + L_{A1} \frac{di_{A1}}{dt} + r_{A1} i_{A1} - S_{A2} V_{DC} - L_{A2} \frac{di_{A2}}{dt} - r_{A2} i_{A2} = 0$$

(C-2)

Each of the phase currents, e.g. $i_{A1}$ and $i_{A2}$, can be divided into two components: 1) the components that forms the grid current and is equal for both converters i.e. $i'_{A1} = i'_{A2} = i'_A = \frac{i_A}{2}$; and 2) the unequal and circulating current component, $i_{PCA}$. Thus, the phase currents can be written as:

$$i_{A1} = i'_A - i_{PCA} \quad , \quad i_{A2} = i'_A + i_{PCA}$$

(C-3)

Substituting (C-3) in (C-2):

$$(S_{A1} - S_{A2}) V_{DC} + L_{A1} \frac{d(i'_A - i_{PCA})}{dt} + r_{A1} (i'_A - i_{PCA}) - L_{A2} \frac{d(i'_A + i_{PCA})}{dt} - r_{A2} (i'_A + i_{PCA}) = 0$$

(C-4)
Figure C-1 Equivalent instantaneous model of parallel-connected two level voltage sourced converters. $S_i$ represents the switching state 1 or 0.

Eq. (C-4) can be simplified as

$$\Delta S_A V_{DC} = 2L_a \frac{d i_{PCA}}{dt} + 2r_A i_{PCA} \quad (C-5)$$

This voltage difference for all the phases may constitute a low frequency zero-sequence component depending on the modulation scheme. Moreover, Eq. (C-5) is a first order differential equation and therefore the intra-phase circulating current generated by the phase voltage difference is highly sensitive to the total path impedance that includes the phase inductances and the parasitic resistances.

On the DC link a nodal equation for the capacitor is:

$$i_L + i_{cap} = S_{A1}i_{A1} + S_{A2}i_{A2} + S_{B1}i_{B1} + S_{B2}i_{B2} + S_{C1}i_{C1} + S_{C2}i_{C2} \quad (C-6)$$

Substituting (C-3) in (C-6) yields:

$$i_L + i_{cap} = S_{A1}i_{A1'} + S_{A2}i_{A2'} + S_{B1}i_{B1'} + S_{B2}i_{B2'} + S_{C1}i_{C1'} + S_{C2}i_{C2'} + \Delta S_A l_{PCA} + \Delta S_B l_{PCB} + \Delta S_C l_{PCC} \quad (C-7)$$

For modeling purposes the load current, $i_L$, is considered a constant DC, i.e. the DC link capacitor is large enough to absorb the high frequency ripple current. Using Eq.(C-1) to Eq.(C-7), the instantaneous model of the parallel-connected rectifier is shown in Figure C-1.

The expression for the phase circulating current of phase $A$ of both the converters was shown in (C-5). The other two phases will also have similar equations. The multiplying factor, $\Delta S_A$, can take a value of ‘2’, ‘0’ or ‘-2’. For most applications, the phase inductances and the parasitic resistances are equal i.e. $L_{A1} \approx L_{A2} = L_A$ and $r_{A1} \approx r_{A2} = r_A$. Therefore, Eq.(C-4) can be used to solve for
$i_{PCA}$ because $\Delta t_A$, during which the switching function difference $\Delta S$ has a non-zero value, is very small:

$$i_{PCA} = \frac{V_{DC}}{2L_A} \Delta S_A \Delta t_A$$ \hspace{1cm} (C-8)

where, $\Delta t_A$ is the time-period for which $\Delta S_A$ is constant.

From (C-8), the phase circulating current is excited by the area term $\Delta S_A \Delta t_A$, therefore, for $i_{PCA}$ to have a zero average, this term needs to be balanced at both the switching frequency level and the fundamental frequency level.

To maintain a zero average at the switching frequency level, the value of $\Delta S_A \Delta t_A$ should remain balanced within each half switching cycle i.e. whenever $\Delta S_A$ takes on a value of ‘2’ for a time span $\delta t$, it must be balanced by a ‘-2’ for the same time span $\delta t$ within half of that switching cycle to keep the average zero. If $\Delta S_A$ is balanced in every half switching cycle, the circulating current that results will have a frequency of the effective switching frequency (i.e. two times the actual switching frequency of each converter) with a zero average. This high frequency circulating current is a direct consequence of interleaved modulation and can only be reduced by increasing the phase inductance $L_A$, or by decreasing $\Delta t_A$ in (C-8) for a constant $V_{DC}$. $\Delta t_A$ can be decreased by either increasing the switching frequency ($f_{sw}$) or via further dividing $\Delta t_A$ by altering the switching template of the space vector modulation scheme. Unbalance of $\Delta S_A$ in any half cycle, however, produces a jump in the circulating current. The polarity of the unbalance dictates the direction of this jump. In succession, these jumps constitute the low frequency component of the circulating current.

The zero-sequence circulating current is modeled as the average of the three-phase currents:

$$i_{CC} = \frac{i_{A} + i_{B} + i_{C}}{3} = \frac{i_{A}' + i_{B}' + i_{C}'}{3} - \frac{i_{PCA} + i_{Pcb} + i_{Pcc}}{3}$$ \hspace{1cm} (C-9)

Since the first part of (C-9) has a zero zero-sequence component by design and control, therefore the current $i_{CC}$ only comprises of the phase circulating currents and can be re-written as (C-10) from (C-8) with the assumption $r_a \approx r_b \approx r_c = r$ and $L_A \approx L_B \approx L_C = L$:

$$i_{CC} = \frac{1}{3} (\Delta S_A \Delta t_A + \Delta S_B \Delta t_B + \Delta S_C \Delta t_C) \frac{V_{DC}}{2L}$$ \hspace{1cm} (C-10)

From (C-10), if the sum of the delta terms is maintained to be zero within each switching cycle, the low frequency common mode circulating current can be reduced. This, however, is not the same as reducing the phase circulating currents. To reduce the phase circulating current each phase circulating current must be reduced individually. In case of unbalanced inductances, the assumptions made to simplify the circulating current expressions do not hold. Therefore, to reduce the common mode and the phase circulating currents the term $\Delta S_X$ must be minimized to reduce both types of circulating currents.
Appendix D

For multiple input multiple output (MIMO) systems, like the parallel connected rectifier, requires an optimal control that provides regulation for all the variable controlled. Linear quadratic regulator (LQR) is a widely used optimal controller that achieves a practical compromise between the control effort and the state feedback.

The control law for the system with LQRI control is

\[ \mathbf{u} = -\mathbf{K}\mathbf{x} = [\mathbf{K}_s \ - \mathbf{K}_i]^T \left[ \int \mathbf{i} \ dt \right] \]  

(D-1)

where \( \mathbf{K} \) is a 6 x 12 state feedback gain matrix with \( \mathbf{K}_s \) representing the proportional and \( \mathbf{K}_i \) representing the integral gains of the states. The state feedback gain vector \( \mathbf{K} \) is calculated by minimizing the LQR cost function \( J \) defined as

\[ J = \int_0^\infty (\mathbf{x}^T \mathbf{Q} \mathbf{x} + \mathbf{u}^T \mathbf{R} \mathbf{u}) \ dt \]  

(D-2)

where, \( \mathbf{Q} \) and \( \mathbf{R} \) are the state and control weighting matrices respectively which should be symmetric. The weights of state and input are chosen to prioritize one over the other. Once the appropriate weighting matrices are selected, \( \mathbf{K} \) is then calculated as

\[ \mathbf{K} = \mathbf{R}^{-1} \mathbf{B}^T \mathbf{P} \]  

(D-3)

where \( \mathbf{P} \) is the solution of the nonlinear Ricatti equation [96].

\[ \mathbf{A}^T \mathbf{P} + \mathbf{P} \mathbf{A} - \mathbf{P} \mathbf{B} \mathbf{R}^{-1} \mathbf{B}^T \mathbf{P} + \mathbf{Q} = 0 \]  

(D-4)

The Ricatti equation presented in (D-4) results from the attempt to find the optimal state feedback gain and can be solved using numerical methods. It is recommended for the states weight \( \mathbf{Q} \) should be higher than the input weight \( \mathbf{R} \) in (D-2) because state is the controlled variable in this case and the input, converter voltage, has a limit associated with it (\( V_{DC} \)).
Appendix E

E.1 Two-Three phase traction drive-based IBC

The system parameters of the of the hardware setup are shown in Table E-1. The system was controlled using a rapid control prototyping (RCP) tool. The current and voltage signals were provided as feedback to the controller by its analogue input channels. Control loops and the PLL were programmed and executed at a 20 μs time step. The contactors were emulated by a resistance of 0.1 Ω, which is greater than the typical value for high current contactors (0.4-1 mΩ), in series with the phase windings. The machines used were typical three-phase SMPM motors available for electric vehicle traction applications. The windings were measured to offer phase inductances $L_A = L_B = L_C = L_{ph} \approx 275 \ \mu H$ and the mutual inductances $M_{AB} = M_{BC} = M_{CA} = M_{ph} \approx -60 \ \mu H$ for both machines.

The DC load was emulated using an active rectifier connected to the grid while controlling the DC link voltage at 50V. The total current drawn from the input was maintained at $10A_{rms}$, thus absorbing 450W from the grid. A digital controller platform was used to implement the control and modulation strategies at a time step of 10µs.

![Hardware experiment setup for the two-three phase synchronous machine based integrated battery charger.](image-url)
Table E-1 System Parameters for Hardware Setup 1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_b$ (V)</td>
<td>50</td>
</tr>
<tr>
<td>$S_b$ (kVA)</td>
<td>1.67</td>
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<tr>
<td>$V_{DC}$ (V)</td>
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<tr>
<td>$f_{sw}$ (kHz)</td>
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</tr>
<tr>
<td>$L_{ph}$ (µH)</td>
<td>275</td>
</tr>
<tr>
<td>$C_{DC}$ (mF)</td>
<td>3</td>
</tr>
</tbody>
</table>

E.2 Six-phase traction drive-based IBC

For experimental validation, a hardware setup with the parameters shown in Table E-2 with a six-phase internal permanent magnet machine was built as shown in Figure E-2. The six-phase machine represents typical machine ratings for light and medium duty vehicle traction applications. The setup included a grid tied inverter to emulate the DC-DC converter and battery. The system was tested in its two modes of operation: 1) as a charger; and 2) as a DR. The DC link voltage is controlled at 400V in both modes. The system was controlled using a digital controller. The current and voltage signals were provided as feedback to the controller by its analog input channels. Control loops and the PLL were programmed and executed at a 20µs time step.

Figure E-2 Hardware experiment setup for the six-phase synchronous machine based integrated battery charger.
The parameters of the simulation and hardware setup used to validate the circulating current strategies is presented in Table E-3.

Table E-3 Circulating current hardware experimental setup

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
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<tbody>
<tr>
<td>$V_b (V_{LL})$</td>
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</tr>
<tr>
<td>$S_b (kVA)$</td>
<td>8.6</td>
</tr>
<tr>
<td>$V_{DC} (V)$</td>
<td>400</td>
</tr>
<tr>
<td>$f_{sw} (kHz)$</td>
<td>18</td>
</tr>
<tr>
<td>$C_{DC} (mF)$</td>
<td>3</td>
</tr>
</tbody>
</table>

### E.3 Circulating Current Hardware Experimental Setup

The parameters of the simulation and hardware setup used to validate the circulating current strategies is presented in Table E-3.

Table E-3 Circulating current hardware experimental setup

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{GLL}$</td>
<td>208 $V_{LL}$ pu</td>
</tr>
<tr>
<td>$S_b$</td>
<td>8.6 $kVA$</td>
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<tr>
<td>$I_G$</td>
<td>80 A</td>
</tr>
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<td>$C_{DC}$</td>
<td>3 $mF$</td>
</tr>
<tr>
<td>$L_{xi}$</td>
<td>1 $mH$</td>
</tr>
</tbody>
</table>

### E.4 Rotor Position Measurement for Torque Validation

A displacement measurement setup was used to monitor the rotor torque oscillations, as shown in Figure E-3. The premise of the setup was to measure the displacement on the rotor caused by the torque oscillations on it. To do this, the inertia of the setup was minimized, i.e. the tests were done without any load on the shaft. This allowed for maximum angular displacement of the rotor due to the torque. In actual application the rotor will have more inertia on the shaft. This will dampen the displacement on the rotor shaft, making it difficult to measure it. The fact that the torque has a frequency similar to the fundamental frequency also makes it difficult to measure the displacement (as its effect) with a large inertia.

In Figure E-3 the optical sensor [113] provides the value of $\Delta x$, which is then converted to angle displacement using the formula $\Delta \theta = \Delta x / R$. This is justified as $\Delta \theta$ is very small. The residual ripple on rotor position is because of the base vibrations on the whole frame on which both motors are
mounted. This was verified by repeating the same procedure at different points of the frame. To minimize the effect of the magneto-restrictive vibrations on the measured oscillations the measurement setup was placed off the frame on which the motors were mounted. This allowed isolation of the sensor from the vibrations of the frame. However, the recorded measurements did have a variation of around 5% on the results. The sources of error in the measurements can be due to the following phenomenon:

1) Displacement due to magnet restrictive vibrations on the drive setup
2) Error in the displacement between the optical sensor and the perpendicular member
3) Dependence of the optical sensors on physical parameters such as temperature, humidity etc.

Figure E-3 Rotor position measurement setup
Bibliography


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Integrated Battery Chargers based on Multiphase Synchronous Machine Drives
By Syed Qaseem Ali


"IEC 61000-3-4 Electromagnetic compatibility (EMC) - Part 3-4: Limits - Limitation of emission of harmonic currents in low-voltage power supply systems for equipment with rated current greater than 16 A," ed: IEC, 1998.


