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ABSTRACT

This thesis presents the complete design, modeling, control, and MATLAB/ Simulink-based implementation of an 11 kW bidirectional Dual Active Bridge DC-DC converter intended for electric vehicle charging applications under both Grid-to-Vehicle (G2V) mode of operation. The converter is designed for a 400 V input DC-link, a 350–400 V battery voltage range, a switching frequency of 100 kHz, a leakage inductance of 50 μ H, and an output capacitance of 500 μ F. A Lithium-ion battery model of 60 Ah capacity has been considered for the analysis. Single-Phase-Shift (SPS) modulation is employed to control the magnitude and direction of the transferred power.

A finite-horizon Model Predictive Controller (MPC) has been developed and implemented within a MATLAB Function block. The controller employs a discrete-time linearised state-space prediction model of the DAB converter, a quadratic cost function, and explicit constraints on phase-shift angle, rate of change, and inductor current. The optimization problem is solved in a receding-horizon fashion with a Quadratic Programming (QP) solver. An RC-based compensation network is proposed to enhance the closed-loop stability and dynamic response. Lead, lag, pole-zero cancellation, feedforward, pole splitting, digital and current-mode compensation methods are also discussed.

The whole system has been simulated in MATLAB/Simulink using Simscape Electrical and Specialized Power Systems libraries. The system performance is evaluated for CC-CV (Constant-Current/Constant-Voltage) charge profile, transient load conditions, bidirectional power-flow cases and stability requirements. Simulation results show that the voltage and current references are accurately tracked, the DC-link behaviour is well regulated, the output voltage ripple is low, the dynamic response is fast under load steps and the transition between G2V and V2G modes is smooth. The

proposed MPC-controlled DAB converter is shown to provide significant advantages over classical PI/PID-controlled topologies in terms of constraint handling, transient performance and bidirectional operation.

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LIST OF SYMBOLS AND ABBREVIATIONS

Symbols

V_1	Primary DC-link voltage, V
V_2	Secondary (battery) voltage, V
i_L	Inductor (leakage) current, A
φ	Phase-shift angle, rad
L_k	Leakage inductance, H
C_o	Output filter capacitance, F
f_s	Switching frequency, Hz
T_s	Sampling period, s
n	Transformer turns ratio
P	Transferred power, W
N_p	Prediction horizon
N_c	Control horizon
Q	State weighting matrix
R	Input weighting matrix
R_d	Input rate weighting matrix
A_d, B_d	Discrete-time state-space matrices
R_{int}	Battery internal resistance, Ω
V_{oc}	Battery open-circuit voltage, V
SOC	State of Charge

Abbreviations

DAB	Dual Active Bridge
MPC	Model Predictive Control
EV	Electric Vehicle
G2V	Grid-to-Vehicle
V2G	Vehicle-to-Grid
SPS	Single Phase Shift
EPS	Extended Phase Shift



DPS	Dual Phase Shift
TPS	Triple Phase Shift
CC	Constant Current
CV	Constant Voltage
HFT	High-Frequency Transformer
ZVS	Zero-Voltage Switching
QP	Quadratic Programming
CCS-MPC	Continuous-Control-Set MPC
FCS-MPC	Finite-Control-Set MPC
PI	Proportional-Integral
PID	Proportional-Integral-Derivative
SiC	Silicon Carbide
GaN	Gallium Nitride
EMI	Electromagnetic Interference
PSFB	Phase-Shifted Full-Bridge
AFE	Active Front-End
PFC	Power Factor Correction
SOA	Safe Operating Area
HIL	Hardware-in-the-Loop
FPGA	Field-Programmable Gate Array
DSP	Digital Signal Processor

CHAPTER 1

INTRODUCTION

1.1 Introduction to EV Charging Systems

There is currently a paradigm shift towards the use of electric cars as opposed to internal combustion engine cars as a result of factors like stringent emission standards, depletion of fossil fuels, environmental changes, and advancements in battery technology. In this regard, it becomes imperative that there be efficient EV charging stations to enhance the usability of the EVs. The EV charging station can be viewed as a multi-stage power electronic converter which facilitates the transfer of electrical energy from the utility grid to the high-voltage traction battery of the EV. In other words, the main function of this converter is to provide safe and accurate electricity delivery from the utility grid to the vehicle battery.

In most instances, the fast charger comprises two main components. The Active Front-End (AFE) converter is applied to convert the alternating current to the direct current, effect power factor correction, and provide control for the DC link and DC-DC converters, while DC-DC converters are used for matching voltages, isolation, and control of charging currents/voltages. In particular, the effectiveness of the fast charger, as well as proper isolation between the vehicle battery and the electric power grid, hinges on the DC-DC converter's performance.

The classification of performance of EV chargers depends largely on the level of charging. Level 1 chargers that are ≤ 2 kW and Level 2 chargers that have a range of 3.3 to 22 kW are AC chargers that can be found within the vehicle itself and are used primarily for domestic use and in offices. Level 3 chargers, which range between 50 kW and 350 kW, and extreme fast chargers (≥ 350 kW) are not within the vehicle and are meant to charge the EV battery in a few minutes [1].

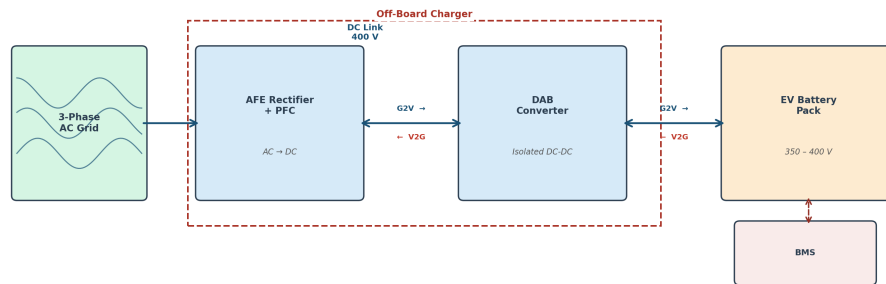


Figure 1.1: Architecture of a Bidirectional EV Charging System

1.2 Need for Isolated DC-DC Converters

Galvanic isolation between the grid and the battery pack becomes necessary. The isolation feature will not allow any possibility of ground faults from the grid getting into the battery pack which can create electric shock hazards while charging, will not circulate the common-mode current through the vehicle frame, and will give a lot of flexibility to configure the grounds. In addition to that, the process of isolation helps with the installation of safety interlocks and leakage current detection systems which are required by international charging standards like IEC 61851, IEC 62196, and SAE J1772 [2].

In addition to safety reasons, another reason for the use of isolated converters is the ability to match voltages from the DC-link that is fixed in most cases to either 400 V or 800 V, and from the battery bank whose voltage can have a very wide range, from 250 V to 450 V in case of a 400 V battery bank, based on factors like state of charge, temperature, and cell chemistry, for example. Voltage matching in isolated converters is achieved using high-frequency transformers whose dimensions are much smaller compared to line-frequency transformers.

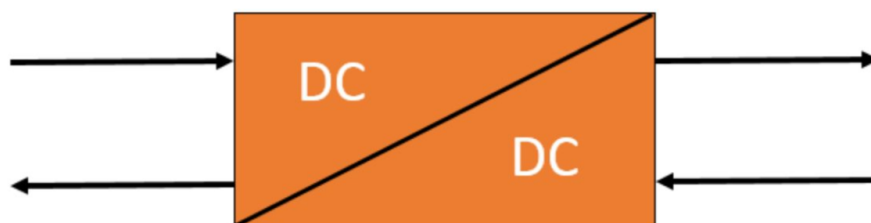


Figure 1.2: Generic Isolated DC-DC Converter Block

1.3 Importance of Bidirectional Power Flow

The novel approach of Vehicle-to-Grid (V2G) system operation converts the EV battery from being a mere demand on the network to becoming an energy source that can provide various services. In today's world, with the increasing deployment of renewable power generation systems, flexibility in the form of distributed storage is essential in order to help balance the grid. The cumulative capacity offered by a few million EVs can potentially provide these services at little expense.

Bi-directional DC/DC converters are the key technology for V2G. This is because the converter has to work in both ways, which means that it should provide energy from grid to vehicle battery (G2V) while the battery is being charged and vice versa (V2G) when the battery is being discharged. There will be seamless switching of the modes according to the command from the control system [3]

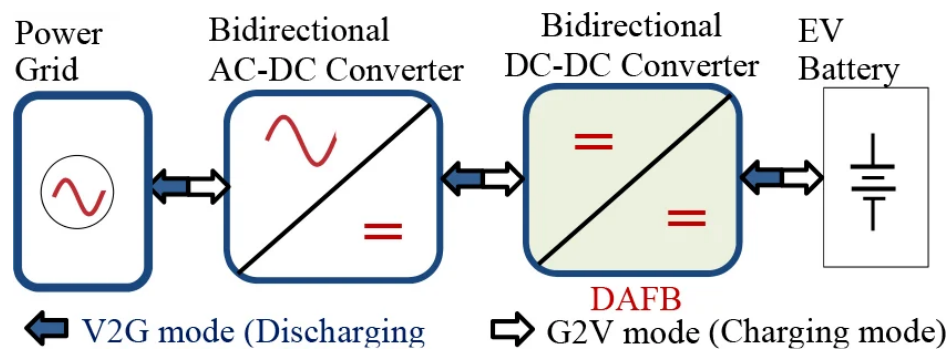


Figure 1.3: Bidirectional Power Flow in G2V and V2G Modes

1.4 Introduction to DAB Converters

The Dual Active Bridge (DAB) converter, initially suggested by De Doncker, Divan, and Kheraluwala in 1991, is among the most efficient isolated bi-directional DC-DC converters available today. The converter comprises two full bridge inverters coupled via a high frequency transformer, with the leakage inductance of the transformer acting as the main element of energy transfer. Two bridges generate shifted square waves, the value of which determines the amount of power transmitted.

The DAB converter possesses following desirable characteristics that can be used in EV charging

- **Natural bidirectional power flow:** Achieved simply by reversing the sign of the phase shift; no hardware reconfiguration is required.
- **Galvanic isolation:** Provided by the high-frequency transformer.
- **Soft switching:** Zero-Voltage-Switching (ZVS) is achievable over a wide operating range, reducing switching losses.

- **High power density:** Enabled by high switching frequency and small magnetic components.
- **Symmetric structure:** Simplifies modular and multi-port designs.
- **Decoupled control:** Power flow and voltage levels can be regulated independently.

These advantages have made the DAB converter the preferred candidate for solid-state transformers, microgrid power conditioners, energy-storage interfaces, and advanced EV chargers.

1.5 Importance of High-Frequency Transformers

The high-frequency transformer (HFT) plays a central role in the operation of the DAB converter. High-speed operation in the range from tens to hundreds of kilohertz allows one to diminish by a factor of ten both the size and weight of the magnetic circuit compared to designs working on the AC mains frequency. Three important functions are performed simultaneously by such a transformer: electrical isolation, transformation, and energy storing through leakage inductance.

The design of the HFT involves careful consideration of core material (typically nanocrystalline or ferrite), winding configuration (litz wire to minimize skin and proximity effects), insulation strategy (to withstand high dv/dt and partial discharge), thermal management, and the precise control of leakage inductance. The leakage inductance value directly determines the maximum transferable power for a given phase shift, the inductor current ripple, and the soft-switching boundary. In the proposed design, an external inductor is often added in series to achieve the targeted leakage value of 50 μH with tight tolerance.

1.6 Need for Advanced Control Methods

Linear controllers including PI and PID have commonly been applied in DAB converters owing to their simple structure. Nevertheless, the DAB converter is nonlinear in terms of power transfer, has considerable variations in its operating point because of the variations in the battery voltage, and has several constraints associated with phase shift and current. As a result, a linear controller whose gains have been tuned based on one operating point cannot perform well under large voltage variations and does not consider any constraints.

Several advanced control techniques such as sliding mode control, fuzzy logic control, neuro controller, finite control set MPC, and continuous control set MPC have been considered for their potential ability to overcome these deficiencies. From

all these approaches, Model Predictive Control (MPC) is regarded as one of the most prospective candidates owing to several attributes.

1.7 Challenges in High-Frequency Switching Systems

While high switching frequency offers compactness and fast dynamics, it also introduces several practical challenges. Switching losses in semiconductor devices increase linearly with frequency, demanding the use of wide-bandgap devices such as Silicon Carbide (SiC) MOSFETs or Gallium Nitride (GaN) HEMTs and the exploitation of soft-switching techniques. Electromagnetic interference (EMI) emissions become significant and necessitate careful filter and PCB design. Parasitic capacitances and inductances of the transformer and interconnects become non-negligible and can cause voltage spikes, ringing, and false triggering. Dead time, sensor noise, and finite digital control resolution further complicate the implementation.

1.8 Motivation for Using MPC

Model Predictive Control (MPC) is an advanced control method based on optimization, which employs an explicit mathematical representation of the system in order to make predictions about its future behavior over a finite prediction horizon. In other words, at each sampling time, the MPC control scheme calculates the optimal sequence of future control inputs via solving an optimization problem, while implementing only the first input value from the generated sequence. The key advantages of using MPC for DAB control include:

- **Multi-objective optimization** of voltage tracking, current tracking, and control effort within a single cost function.
- **Explicit handling of physical constraints** such as phase-shift limitation ($|\varphi| \leq \pi/2$), rate-of-change limitation, peak inductor current limitation, and battery current/voltage limitation.
- **Predictive compensation** of plant behavior in the design that ensures better transient performance than reactive approach.
- **Automatic accommodation** of bidirectional power flow without re-configuring the controller.
- **Direct extension** to nonlinear, robust, and adaptive systems.

1.9 Stability Issues in DAB Converters

In spite of its relatively simple structure, the DAB converter presents some stability issues. Some of the major stability issues associated with the RHP zero phenomenon that occurs under specific operational conditions, the small signal effects due to the coupling of the primary DC-link capacitor to the output capacitor via the leakage inductance, the dependence of the loop gain on the operation point, and the nonlinear nature of the power-phase curve are some of the parameters that cause instability.

1.10 Importance of Compensation Techniques

It is important to note that a compensation network is critical in tuning the loop gain in such a manner that there will be enough gain, phase margins, bandwidth, and static accuracy. The compensation can be done either in an analog fashion with passive RC networks or in digital fashion through software algorithms within the controller program. It all depends on the dynamics of the power converter, bandwidth needs, and disturbance rejection criteria of the system. For this thesis, a compensation network based on RC networks has been implemented along with MPC, supported by analytical comparison of several other compensation schemes.

1.11 Objectives of the Thesis

The key goals of this thesis are as follows:

1. To propose an 11 kW bidirectional DAB DC-DC converter for electric vehicle charging with input DC voltage of 400 V, 350-400 V DC battery voltage, switching frequency of 100 kHz, leakage inductance of 50 μ H, and output capacitance of 500 μ F.
2. To derive the mathematical and state space model of the DAB converter under Single-Phase Shift control technique.
3. To design Finite-Horizon Model Predictive Control for phase shift control of the DAB converter accounting for system constraints.
4. To incorporate an RC feedback network and evaluate various possible compensation approaches.
5. To develop the suggested converter/controller in MATLAB/Simulink environment employing Simscape Electrical and Specialized Power System libraries.
6. To validate the results obtained during CC-CV charging process and bidirectional grid to vehicle/vehicle to grid operation.

7. To compare the performance of the proposed MPC controlled DAB converter with that of PI controlled converter.

1.12 Contributions of the Proposed Work

The key contributions made by this thesis are:

1. A unified MATLAB/Simulink design framework that integrates power-stage modeling, controller synthesis, compensation, and validation in a single environment.
2. A constraint-aware finite-horizon MPC formulation for SPS-modulated DAB converters implemented in a MATLAB Function block, suitable for code generation.
3. A detailed comparative analysis of RC, lead, lag, pole-zero cancellation, feedforward, pole-splitting, digital, and current-mode compensation methods in the context of DAB converters.
4. Demonstration of seamless G2V/V2G transition using the proposed control architecture.
5. A comprehensive set of simulation results, performance metrics, and benchmark comparisons against PI-controlled baselines.

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CHAPTER 2

LITERATURE REVIEW

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2.1 Overview

This chapter presents a structured and critical review of the state-of-the-art literature relevant to the design, modelling, and control of bidirectional Dual Active Bridge converters for electric vehicle charging applications. The review is organized into seven sections: (i) EV charging converter topologies, (ii) isolated bidirectional converters, (iii) DAB converter architectures, (iv) comparison of DAB with LLC and full-bridge converters, (v) phase-shift modulation techniques, (vi) PI/PID control methods, and (vii) Model Predictive Control in power electronics. Following this, the existing stability and compensation techniques are discussed, comparative analysis tables are presented, and the chapter concludes by identifying the research gaps and providing the justification for the proposed work.

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2.2 Review of EV Charging Converter Topologies

EV charging systems are commonly categorized into on-board chargers (OBCs) and off-board chargers. OBCs are integrated into the vehicle and typically operate up to 22 kW; they are size- and weight-constrained, demanding the highest power density. Off-board fast and ultra-fast chargers are installed at charging stations and operate at significantly higher power levels (50 kW to several hundred kilowatts). The most widely studied topologies for the DC-DC stage of such chargers include:

- **Non-isolated buck/boost converters:** Simple and efficient but lack galvanic isolation, restricting their use to low-power applications.
- **Phase-shifted full-bridge (PSFB) converters:** Provide isolation and ZVS but are predominantly unidirectional.

- **LLC resonant converters:** Offer excellent efficiency over a narrow operating range but suffer from gain limitations and complex control under wide voltage variation.
- **CLLC resonant converters:** A bidirectional variant of LLC with symmetric resonant tanks; control complexity remains high.
- **Dual Active Bridge (DAB) converters:** The focus of this thesis, offering native bidirectional capability, symmetric structure, and wide-range operation.

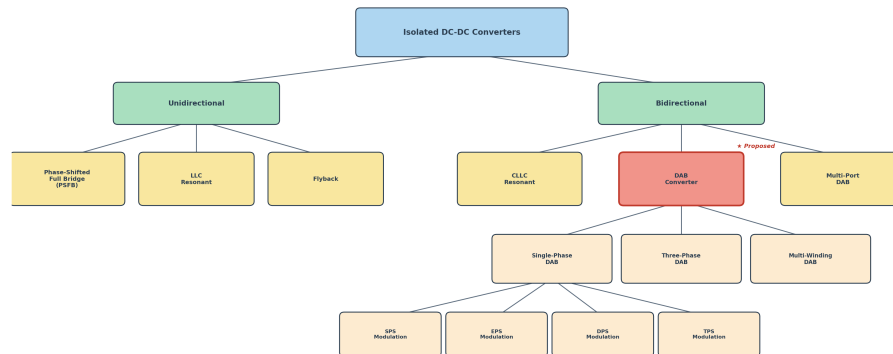


Figure 2.1: Classification of Isolated DC-DC Converter Topologies

2.3 Review of Isolated Bidirectional Converters

Several authors have reviewed isolated bidirectional topologies for EV and energy-storage applications. Krismer and Kolar [4] provided efficiency-optimized designs of DAB converters tailored for automotive applications. Inoue and Akagi [5] proposed a bidirectional DC-DC converter for energy storage incorporating advanced modulation. Zhao et al. [6] introduced extended-phase-shift and dual-phase-shift modulations to improve light-load efficiency. Tan et al. [7] presented a bidirectional isolated DC-DC converter for battery energy storage systems demonstrating high efficiency. Bai and Mi [8] proposed a dual-phase-shift strategy to eliminate reactive power and improve efficiency in isolated DAB converters. Wang et al. [9] demonstrated soft-switching operation principles in bidirectional full-bridge DC/DC converters. Sha et al. [10] achieved ZVS throughout the full load range using optimized switching patterns. Xue et al. [11] designed a DAB-based charger for plug-in hybrid EVs. Islam et al. [12] reviewed power converters for medium-voltage networks. Everts et al. [13] presented optimal ZVS modulation for single-phase single-stage bidirectional DAB AC-DC converters. These studies collectively establish the DAB converter as the leading candidate for bidirectional isolation in high-power applications.

2.4 Comparison with LLC and Full-Bridge Converters

The DAB converter provides some pros and cons when compared with the LLC and PSFB converters. Unlike the LLC converter which employs varying frequencies during operation, the DAB converter operates at fixed switching frequency while power modulation is done using phase shift. The DAB converter is bidirectional in nature while the PSFB converter is unidirectional except if changes are made to it. The LLC converter has greater efficiency at its optimal operating point than the DAB converter..

Table 2.1: Comparison of Isolated DC-DC Converter Topologies

Feature	DAB	LLC	PSFB	CLLC
Bidirectional	Yes	No	No	Yes
Isolation	Yes	Yes	Yes	Yes
ZVS range	Wide	Narrow	Moderate	Moderate
Frequency	Constant	Variable	Constant	Variable
Control complexity	Low–Moderate	High	Low	High
Power density	High	High	Moderate	High

2.5 Review of Phase-Shift Modulation Methods

Following modulation techniques can be adopted for DAB converters:

- **Single Phase Shift (SPS):** Simplest modulation technique; where only the phase shift between bridges is regulated. It is easy to design but generates high circulating current and reactive power under light loads and wide input voltage ranges.
- **Extended Phase Shift (EPS):** Adds an inner phase shift to one of the bridges, reducing reactive power.
- **Dual Phase Shift (DPS):** Independent inner phase shifts on both bridges, providing additional degrees of freedom.
- **Triple Phase Shift (TPS):** In this technique, all three phase shifts are controlled, thus giving maximum efficiency with added control complexity.

SPS modulation is adopted in this thesis because of its simplicity, its compatibility with the proposed MPC formulation, and its direct mapping to the linearised plant model.

Table 2.2: Comparison of Phase-Shift Modulation Strategies

Feature	SPS	EPS	DPS	TPS
Control variables	1	2	2	3
Complexity	Low	Medium	Medium	High
Reactive power	High	Medium	Low	Lowest
ZVS range	Moderate	Extended	Extended	Full
MPC compatibility	High	Medium	Medium	Low

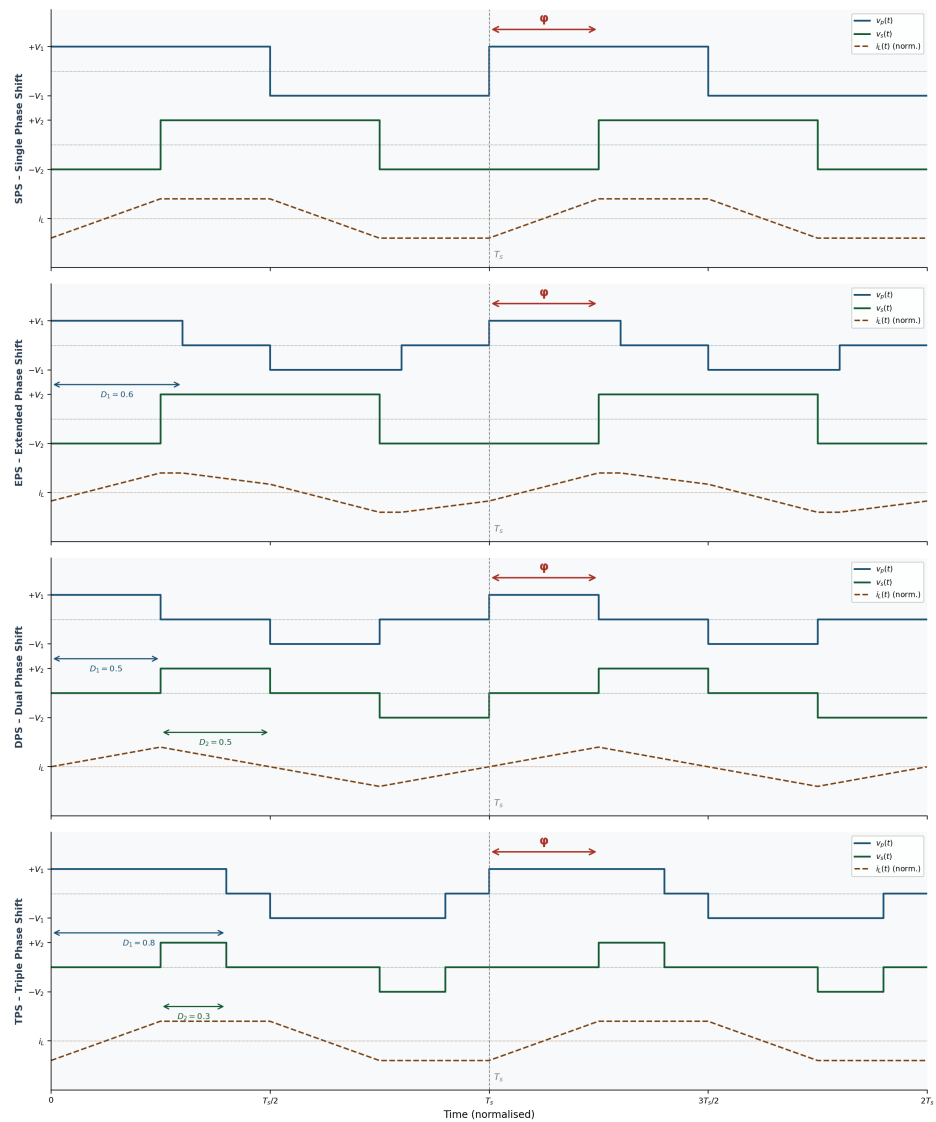


Figure 2.2: Phase-Shift Modulation Schemes (SPS, EPS, DPS, TPS)

2.6 Review of PI/PID Control Methods

PI and PID controllers still form the backbone of industrial DAB controller applications due to their simplicity, easy tuning ability, and known characteristics. The conventional approach is cascaded **outer voltage control loop and inner current control loop**. Nevertheless, some disadvantages associated with this kind of controllers are:

- Performance degrades significantly under wide battery voltage variation.
- Constraints on phase shift and inductor current must be enforced via saturation, leading to integrator windup and degraded transient response.
- Right-half-plane zero behaviour and small-signal nonlinearity require gain scheduling.
- Sub-harmonic oscillations may appear at high control bandwidth.

2.7 Review of MPC in Power Electronics

MPC algorithms **have been** widely studied **in the** field of power electronics within the last two decades. There are two different types of algorithms:

- **Finite-Control-Set MPC (FCS-MPC):** Directly evaluates a finite set of switching states; no modulator is needed. Computationally fast but has disadvantages such as variable switching frequency and significant EMI generation.
- **Continuous-Control-Set MPC (CCS-MPC):** Computes a continuous control variable (e.g., phase shift) and uses a modulator. Constant switching frequency, suitable for SPS-modulated DAB.

The work utilizes the CCS-MPC formulation, with phase shift as the manipulated variable [14, 15].

2.8 Existing Stability and Compensation Techniques

Several methods for improving stability have been mentioned:

- Lead-lag compensators for phase-margin improvement.
- Damping networks (RC parallel) across the leakage inductor or output capacitor to suppress resonance.
- Feedforward compensation for fast disturbance rejection.
- Pole-zero cancellation for simplifying loop dynamics.
- Digital filters and notch filters within the controller.

2.9 Comparative Analysis of Previous Research

Table 2.3: Comparative Review of Existing Research on DAB Converters

Reference	Power	Control	Modulation	Bidirectional Constraints	
De Doncker [3]	5 kW	Open-loop	SPS	Yes	No
Krismer [4]	2 kW	PI	TPS	Yes	No
Zhao [6]	1 kW	PI	EPS/DPS	Yes	No
Inoue [5]	10 kW	PI	SPS	Yes	No
Tan et al. [7]	5 kW	PI	SPS	Yes	No
Bai & Mi [8]	11 kW	Adaptive PI	DPS	Yes	No
This work	11 kW	MPC	SPS	Yes	Yes

2.10 Research Gaps and Justification

The review identifies the following gaps in the existing literature:

1. Majority of the MPCs used are based on FCS-MPC leading to switching of varying frequency and poor performance in terms of EMI.
2. Only a few cases incorporate MPCs along with passive compensation networks for simultaneous transient stability improvement
3. There have been insufficient cases that demonstrate G2V/V2G conversion with actual CC-CV charge profile.
4. The implementation of constrained MPC for SPS-DAB at 100 kHz in a code-generation-friendly MATLAB Function block remains rare in the literature.

This thesis directly addresses these gaps through a constrained CCS-MPC formulation, integration of RC compensation, and comprehensive MATLAB/Simulink validation of CC-CV and bidirectional operation.

CHAPTER 3

PROPOSED SYSTEM AND DESIGN

METHODOLOGY

3.1 Overall Architecture of the DAB Converter

The proposed system is an 11 kW bidirectional isolated Dual Active Bridge DC-DC converter intended to operate as the second stage of an off-board EV fast charger. The primary side of the converter is connected to a regulated 400 V DC-link, which is typically supplied by an Active Front-End (AFE) PFC rectifier from the three-phase utility grid. The secondary side interfaces with the high-voltage battery pack of the electric vehicle, operating over a voltage range of 350–400 V with a nominal value of 400 V.

The converter comprises a primary-side full-bridge inverter (S1–S4), a high-frequency transformer (HFT) with unity turns ratio and an external leakage inductor ($L_k = 50 \mu\text{H}$), a secondary-side full-bridge inverter (S5–S8) that operates as a synchronous rectifier or as an inverter during V2G mode, and an output DC-link filter capacitor ($C_o = 500 \mu\text{F}$). All semiconductor switches are assumed to be SiC MOS-FETs to support the 100 kHz switching frequency with low conduction and switching losses.

3.2 Working Principle of DAB Operation

In Single-Phase-Shift modulation, both H-bridges are operated with 50% duty ratio, producing complementary square-wave voltages of fundamental frequency equal to the switching frequency $f_s = 100 \text{ kHz}$. The primary bridge produces a voltage $v_p(t)$ of magnitude $\pm V_1$ across the transformer primary, while the secondary bridge produces $v_s(t)$ of magnitude $\pm V_2$ across the secondary. The phase shift φ between

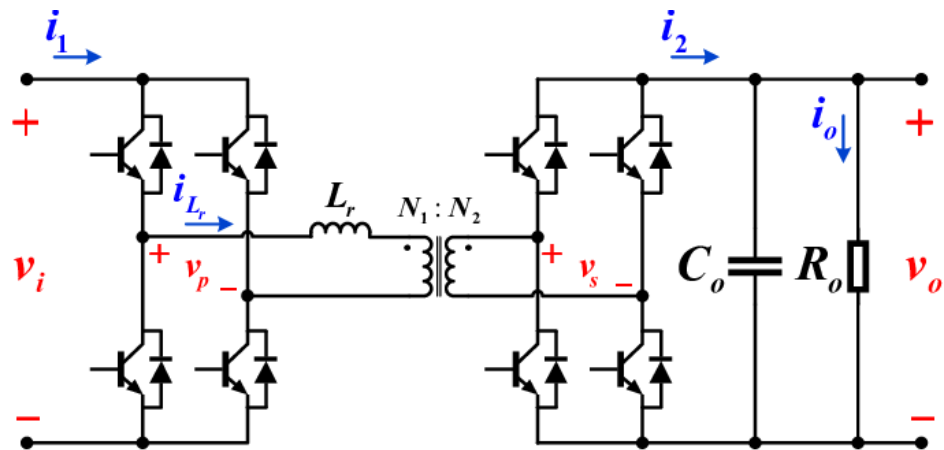


Figure 3.1: Proposed DAB Converter Topology

$v_p(t)$ and $v_s(t)$ drives a current $i_L(t)$ through the leakage inductance L_k , thereby transferring energy between the two ports.

When $\varphi > 0$, the primary leads the secondary and power flows from the primary side (grid DC-link) to the secondary side (battery), corresponding to G2V charging. When $\varphi < 0$, the secondary leads the primary, reversing the power flow direction for V2G operation.

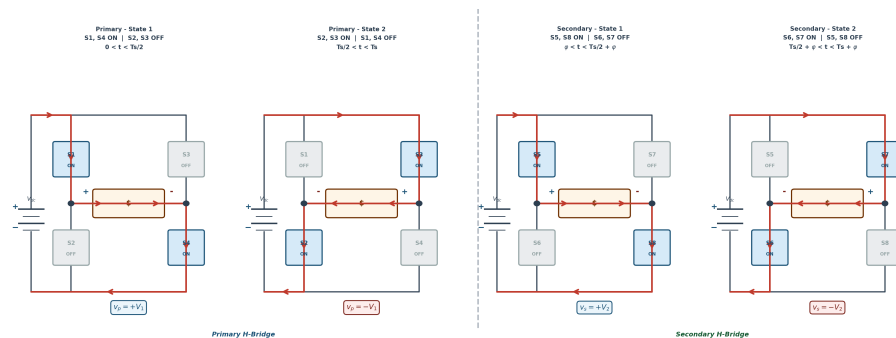


Figure 3.2: Primary and Secondary H-Bridge Operation

3.3 Bidirectional Power Transfer Mechanism

Bidirectional characteristics in the DAB converter come about due to the active characteristic of both bridges as well as symmetry of the structure itself. Different from PSFB converters that have a secondary circuit made up of diodes or a controlled rectifier lacking inversion capabilities, both bridges of the DAB converter can independently apply voltage to the winding of the transformer. By altering the phase shift, it is only natural for the polarity of the average voltage across the leakage inductance to change, thereby causing the direction of the average inductor current to change as well.

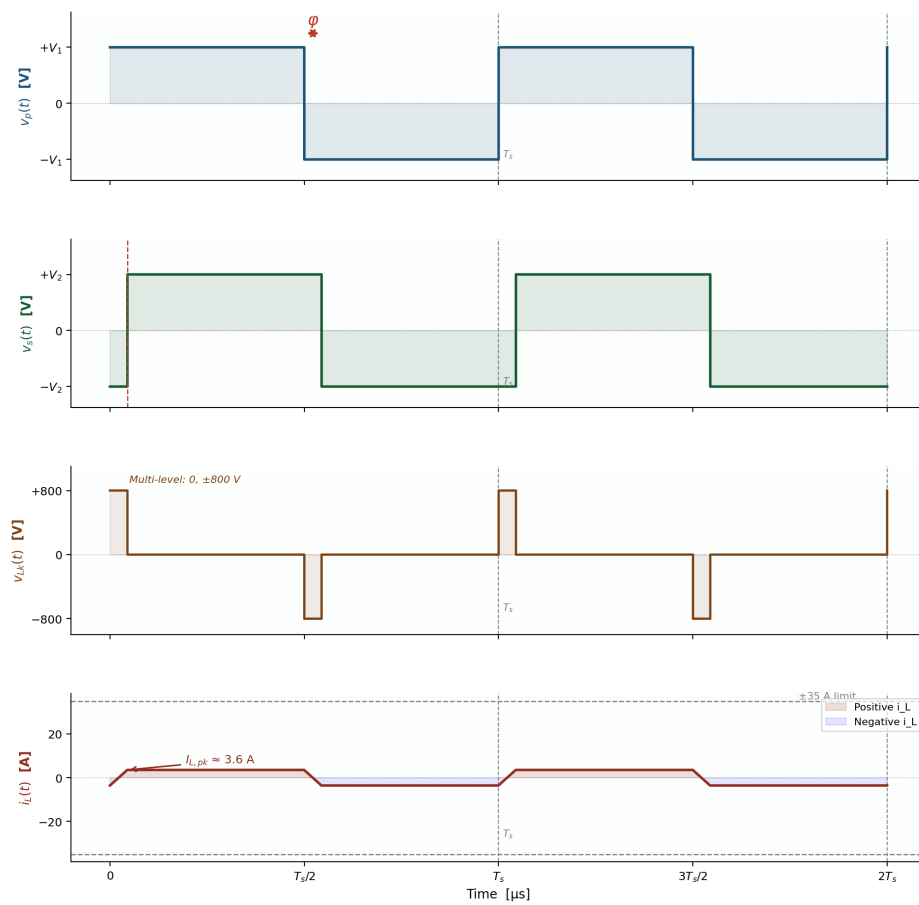


Figure 3.3: Phase-Shift Power Transfer Waveforms

3.4 Phase-Shift Power Transfer Principle

For SPS modulation with both bridges operating at 50% duty cycle, the average transferred power is given by the well-known closed-form expression [3]:

$$P = \frac{n \cdot V_1 \cdot V_2}{2\pi f_s L_k} \cdot \varphi \left(1 - \frac{|\varphi|}{\pi}\right) \quad (3.1)$$

where:

- n = transformer turns ratio (N_p/N_s), here equal to 1,
- V_1 = primary DC voltage,
- V_2 = secondary DC voltage,
- φ = phase-shift angle in radians ($-\pi/2 \leq \varphi \leq \pi/2$),
- f_s = switching frequency,
- L_k = leakage inductance.

The maximum transferable power occurs at $\varphi = \pi/2$ and equals $P_{max} = \frac{nV_1V_2}{8f_sL_k}$. For the design values $V_1 = 400$ V, $V_2 = 400$ V, $f_s = 100$ kHz, and $L_k = 50$ μ H, which, with appropriate normalization, yields a comfortable margin above the 11 kW rated power.

3.5 Primary Bridge Operation

The four SiC MOSFETs S1-S4 form the primary bridge configuration with respect to the primary winding of the HFT. Switches S1 and S4 turn ON together to provide $+V_1$ on the primary side of the HFT. When S2 and S3 turn ON, V_1 appears on the primary winding. A delay of 200 ns is provided between each complementary pair of switches to ensure that shoot-through does not occur. The gate drive signals are produced using the PWM subsystem in Simulink.

3.6 Secondary Bridge Operation

S5-S8, which is used as a secondary bridge, is similar to the first bridge but utilizes a phase-shifted square wave on the secondary side. For G2V applications, the secondary bridge acts like a synchronous rectifier, while for V2G applications, the secondary bridge acts like an inverter, while the first bridge performs synchronous rectification.

3.7 High-Frequency Transformer Design

The HFT is made with specifications of 11 kW power capacity, 100 kHz frequency, and 1:1 turn ratio. The core material to be used will be annanocrystalline material (e.g., Vitroperm 500F). This is to ensure minimal losses in the core due to the high-frequency operation of the HFT [16]. The number of turns is determined from Faraday's law:

$$N_p = \frac{V_1}{4 \cdot f_s \cdot B_{max} \cdot A_e} \quad (3.2)$$

For $B_{max} = 0.3$ T and $A_e = 4.5$ cm² (a typical core area for 11 kW at 100 kHz), $N_p \approx 7$ turns; $N_s = 7$ turns. Interleaved windings using Litz wire are used to reduce leakage, and skin effect/proximity effect.

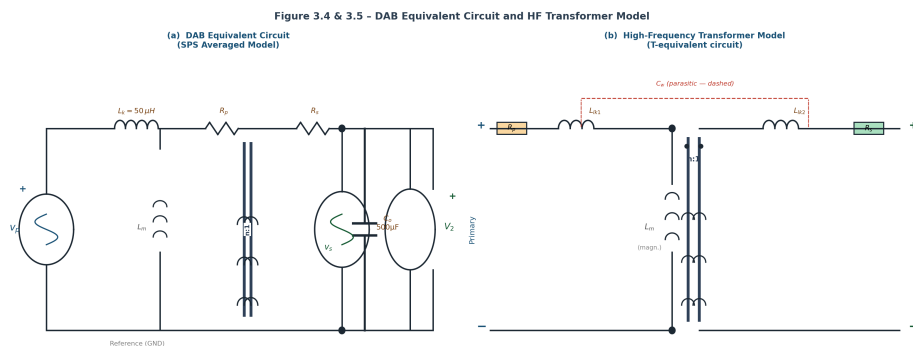


Figure 3.4: High-Frequency Transformer Model

3.8 Leakage Inductance Modeling

The leakage inductance L_k represents the key element of the DAB topology. It dictates the maximum transferable power, current ripple of the inductor, the limit of soft switching, and the small signal behavior. In our case, $L_k = 50$ μH is composed of the leakage of the transformer itself ($\approx 10 - 15$ μH) and an external air-core inductor ($\approx 35 - 40$ μH). The use of the external inductor enables tight tolerance control regardless of transformer manufacturing tolerances.

3.9 Output Filter Capacitor Design

The selection of the output capacitor, $C_o = 500$ μF, is made in a way that allows for maintaining the voltage ripple on the side of the battery below 1% of the voltage nominal value when operating at full load. The capacitor is constructed from film and electrolytic capacitors in parallel connection having low ESR. Voltage ripple

within the capacitor will be given by:

$$\Delta V_o \approx \frac{I_o \cdot \Delta t}{C_o} \quad (3.3)$$

For $I_o \approx 27.5$ A and $\Delta t = T_s/2 = 5 \mu s$, $\Delta V_o \approx 0.275$ V (0.07%), well within specification.

3.10 Battery Modeling

The battery is modelled as a Thevenin equivalent circuit comprising an SOC-dependent open-circuit voltage source $V_{oc}(\text{SOC})$ and an internal resistance R_{int} [17]. A nominal capacity of 60 Ah and a nominal voltage of 400 V are assumed, corresponding to a stored energy of 24 kWh. The terminal voltage is given by:

$$V_{batt} = V_{oc}(\text{SOC}) - I_{batt} \cdot R_{int} \quad (3.4)$$

Table 3.1: Battery Specifications

Parameter	Value
Chemistry	Lithium-ion
Nominal voltage	400 V
Capacity	60 Ah
Energy	24 kWh
Internal resistance	0.05 Ω
Voltage range	350–420 V

3.11 SOC Estimation Methodology

The State of Charge (SOC) is estimated using the Coulomb-counting method:

$$\text{SOC}(t) = \text{SOC}(0) - \frac{1}{C_{Ah} \cdot 3600} \int_0^t I_{batt}(\tau) d\tau \quad (3.5)$$

This approach is implemented in a dedicated Simulink subsystem with an integrator block, gain blocks, and saturation between 0 and 1 (or 0–100%).

Table 3.2: Design Specifications of the Proposed DAB Converter

Parameter	Value
Rated power	11 kW
Primary voltage (V_1)	400 V
Secondary voltage (V_2)	350–400 V
Switching frequency (f_s)	100 kHz
Leakage inductance (L_k)	50 μ H
Output capacitance (C_o)	500 μ F
Transformer turns ratio (n)	1:1
Dead time	200 ns
Switches	SiC MOSFET
MPC sampling time (T_s)	10 μ s

3.12 Design Specifications

3.13 Mathematical Derivations

3.13.1 DAB Power Transfer Equation

From the inductor voltage law for one half-period and by integrating the current waveform obtained, the following value for average power delivered can be obtained::

$$P_{avg} = \frac{nV_1V_2}{2\pi f_s L_k} \cdot \varphi \left(1 - \frac{|\varphi|}{\pi} \right) \quad (3.6)$$

3.13.2 Inductor Current Equations

The peak inductor current under SPS operation is:

$$I_{L,peak} = \frac{nV_1 + V_2}{4f_s L_k} + \frac{(nV_1 - V_2) \cdot \varphi}{2\pi f_s L_k} \quad (3.7)$$

3.13.3 State-Space Model

Defining the state vector $\mathbf{x} = [i_L \ V_2]^T$ and the control input $u = \varphi$, the averaged state-space model of the DAB converter is:

$$\dot{\mathbf{x}} = A_c \mathbf{x} + B_c u \quad (3.8)$$

with

$$A_c = \begin{bmatrix} 0 & -\frac{1}{L_k} \\ \frac{1}{C_o} & -\frac{1}{R_{load}C_o} \end{bmatrix}, \quad B_c = \begin{bmatrix} b_1 \\ 0 \end{bmatrix} \quad (3.9)$$

where b_1 represents the small-signal input gain of phase shift linearised about an operating phase shift φ_0 .

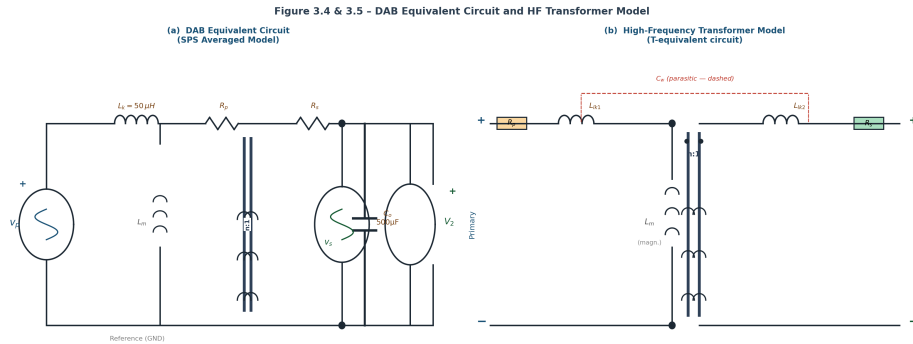


Figure 3.5: Equivalent Circuit of the DAB Converter

3.13.4 Voltage and Transformer Equations

The voltage across the leakage inductance is:

$$v_L(t) = L_k \frac{di_L}{dt} = v_p(t) - n \cdot v_s(t) \quad (3.10)$$

Integrating yields the inductor current:

$$i_L(t) = i_L(t_0) + \frac{1}{L_k} \int_{t_0}^t [v_p(\tau) - n \cdot v_s(\tau)] d\tau \quad (3.11)$$

3.13.5 Switching Equations

The instantaneous primary and secondary bridge voltages under SPS are:

$$v_p(t) = V_1 \cdot \text{sgn}(\sin(\omega_s t)) \quad (3.12)$$

$$v_s(t) = V_2 \cdot \text{sgn}(\sin(\omega_s t - \varphi)) \quad (3.13)$$

3.14 Selection of Switching Frequency

The selection of a switching frequency of 100 kHz is done considering a trade-off between power density and switching losses. Increased frequencies lead to reduced magnetics but higher switching losses and EMC emissions. The SiC MOSFET technology is optimal for these switching frequencies.

3.15 Selection of Leakage Inductance

The leakage inductance is chosen such that the maximum amount of power transferred is more than the rated power by sufficient margins ($\geq 1.5 \times$) while maintaining acceptable peak inductor currents and ZVS operations. A L_k of 50 μH fulfills all

requirements.

3.16 Transformer Turns Ratio

Since the voltage ratings for the primary and secondary are identical, 400 V, a 1:1 turns ratio is selected for maximum ZVS and reduced circulating currents.

3.17 PWM Implementation

The PWM design implemented in Simulink provides 4 sets of complementary PWM signals at a frequency of 100 kHz with dead times of 200 ns. Two PWM sets have constant phases while another one is offset with the commanded phase angle φ .

3.18 Simulink Implementation Methodology

The complete converter is implemented in Simulink using:

- Universal Bridge blocks from Simscape Electrical (Specialized Power Systems) configured as 4-arm MOSFET bridges.
- powergui block in discrete mode with sample time $T_{pgui} = 100$ ns.
- Measurement blocks for voltage and current sensing.
- PWM subsystem built from triangular carriers and comparators.
- MATLAB Function block for the MPC controller.
- Battery model from Simscape Electrical with Coulomb-counting SOC.

3.19 Detailed Block Explanation

3.19.1 Universal Bridge Block

Simulates the full-bridge inverter with adjustable number of arms, switching device, snubber, and parasitic elements. Used in both primary and secondary bridge circuits.

3.19.2 powergui Block

Essential block in any Simscape Electrical model involving Specialized Power Systems. Set up as a discrete system with a 100 nanoseconds sample time for the 100 kilohertz switching events

3.19.3 Measurement Blocks

Voltage and current measuring blocks have been added to the circuit diagram at suitable nodes: primary DC-link, secondary DC-link, primary side current of the transformer, inductor current, and battery current.

3.19.4 PWM Subsystem

Provides 8 gate pulses with 50% duty cycle, 100 kilohertz frequency, 200 nanoseconds dead time, and a phase angle between the two bridge systems..

3.19.5 MATLAB Function Block Implementation

The MPC control algorithm can be implemented using the MATLAB Function block which takes the measurements of the states and references, calculates the prediction matrices and solves the quadratic program to give the optimum phase shift value. The implementation is done keeping the code generation capability of the block in mind.

3.19.6 SOC Subsystem

A Coulomb counting algorithm using the integrator, gain ($1/(C_{Ah} \cdot 3600)$), saturation block (0–1), and SOC output block.

CHAPTER 4

MODEL PREDICTIVE CONTROL AND COMPENSATION TECHNIQUES

PART A — MODEL PREDICTIVE CONTROL

4.1 Fundamentals of MPC

Model Predictive Control (MPC) is a complex optimization-based control approach that employs a model of the plant dynamics to make predictions about future behavior for a specified time horizon, optimizes over the predicted behavior using a constrained optimization problem at each sample time point, and uses only the initial value of the control sequence as feedback for the next cycle [14, 18].

The MPC approach contrasts with the more conventional control methods of PI and PID in several ways. To begin, MPC is a natural fit for multivariable processes without any need for decoupling. Secondly, MPC includes the physical or operational constraints on input, output, and state variables in the optimization process. Lastly, MPC is based on predictions of the future behavior of the plant and thus performs better in disturbance response.

4.2 Prediction Model

The prediction model is the cornerstone of any MPC implementation. For the proposed DAB converter, the discrete-time linearised state-space model derived in Section 3.13.3 is used:

$$\mathbf{x}(k+1) = A_d \mathbf{x}(k) + B_d u(k) \quad (4.1)$$

where:

- $\mathbf{x} = [i_L \ V_2]^T$ is the state vector,

Figure 4.1 - MPC Architecture for DAB Control

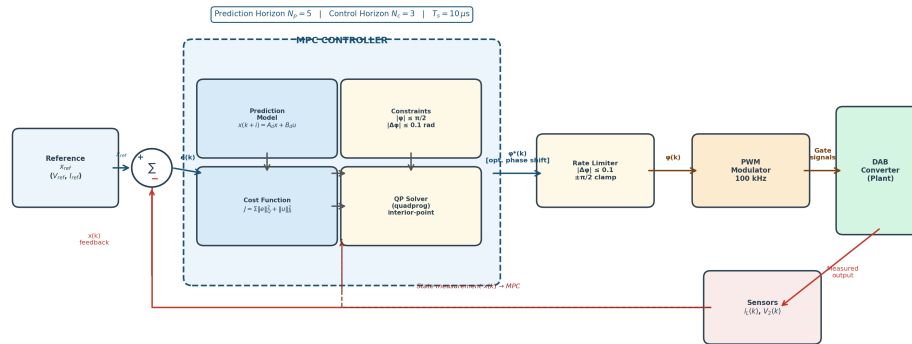


Figure 4.1: MPC Architecture for DAB Control

- $u = \varphi$ is the control input (phase shift),
- A_d, B_d are the discrete-time system matrices obtained by zero-order-hold discretization of the continuous-time matrices at sampling period $T_s = 10 \mu s$.

The continuous-time matrices are initially linearized for the nominal operation point of phase shift $\varphi_0 = 0.25$ rad. With the forward Euler scheme, we get $A_d = I + A_c T_s$ and $B_d = B_c T_s$. If forward Euler discretization fails to provide stability numerically, we apply the exact zero-order hold discretization using `c2d`.

Iterating the model forward over the prediction horizon N_p yields the stacked prediction equation:

$$\mathbf{X} = \Phi \mathbf{x}(k) + \Gamma \mathbf{U} \quad (4.2)$$

where $\Phi \in \mathbb{R}^{2N_p \times 2}$ and $\Gamma \in \mathbb{R}^{2N_p \times N_c}$.

4.3 Cost Function

The MPC cost function is formulated as a quadratic function of the predicted tracking error, control effort, and control-rate effort:

$$J = \sum_{i=1}^{N_p} (\mathbf{x}(k+i) - \mathbf{x}_{ref})^T Q (\mathbf{x}(k+i) - \mathbf{x}_{ref}) + \sum_{j=0}^{N_c-1} u(k+j)^T R u(k+j) + \sum_{j=0}^{N_c-2} \Delta u(k+j)^T R_d \Delta u(k+j) \quad (4.3)$$

The weighting matrices are:

$$Q = \begin{bmatrix} 1 & 0 \\ 0 & 100 \end{bmatrix}, \quad R = 0.01, \quad R_d = 0.1 \quad (4.4)$$

The Q matrix gives $100\times$ more weight to the voltage state V_2 compared to

the inductor current state i_L since the importance is to achieve voltage regulation with high accuracy. The cost function R imposes penalty on the magnitude of the phase shift so as to ensure ZVS conditions.

4.4 Constraints

The optimization explicitly enforces:

- **Input magnitude constraint:**

$$|\varphi| \leq \frac{\pi}{2} \quad (4.5)$$

- **Input rate constraint:**

$$|\Delta\varphi| \leq 0.1 \text{ rad/sample} \quad (4.6)$$

- **State (current) constraint:**

$$|i_L| \leq 35 \text{ A} \quad (4.7)$$

These constraints ensure that the converter remains within its safe operating area (SOA), within ZVS limits, and within the thermal and magnetic limits of the transformer and switches.

4.5 Prediction and Control Horizons

The prediction horizon is set at $N_p = 5$, while the control horizon is set to be $N_c = 3$. At $T_s = 10 \mu\text{s}$, the prediction horizon will imply a look-ahead of $50 \mu\text{s}$, which can predict the dynamics of the dominant components in the system. Too small horizons mean less optimality, whereas too big horizons mean more computational complexity.

Table 4.1: MPC Parameters

Parameter	Value
Prediction horizon (N_p)	5
Control horizon (N_c)	3
Sampling time (T_s)	$10 \mu\text{s}$
Q (voltage weight)	100
Q (current weight)	1
R (input weight)	0.01
R_d (rate weight)	0.1
φ_{max}	$\pm\pi/2 \text{ rad}$
$\Delta\varphi_{max}$	0.1 rad/sample
$i_{L,max}$	$\pm 35 \text{ A}$

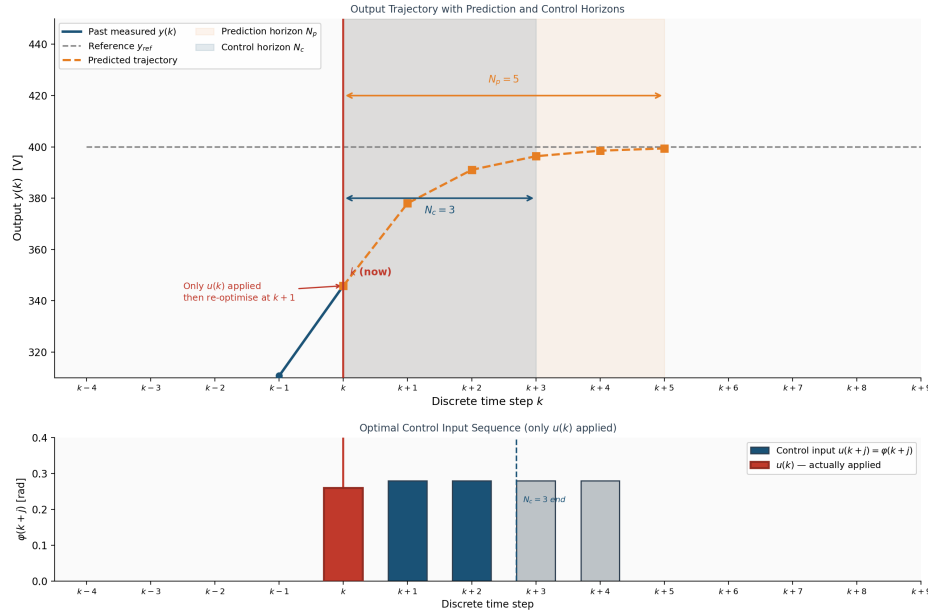


Figure 4.2: Receding Horizon Concept

4.6 Optimization Process

With the prediction equations plugged into the cost function, we have a standard **quadratic programming (QP) problem**:

$$\min_{\mathbf{U}} \frac{1}{2} \mathbf{U}^T \mathbf{H} \mathbf{U} + \mathbf{f}^T \mathbf{U} \quad (4.8)$$

$$\text{subject to } \mathbf{A}_{ineq} \mathbf{U} \leq \mathbf{b}_{ineq} \quad (4.9)$$

Here, the Hessian $\mathbf{H}$ and gradient $\mathbf{f}$ are generated based on Φ , Γ , $\mathbf{Q}$, $\mathbf{R}$, $\mathbf{R}_d$, current state $\mathbf{x}(k)$, and reference trajectory. This QP problem is solved via MATLAB quadprog, using the interior-point-convex method, at every sample. In case of failure or infeasibility returned by the solver, the previously used phase shift is used to maintain continuity of the control output.

4.7 MPC Implementation in MATLAB Function Block

The MPC block design is represented by the DAB_MPC_MATLAB_Function_Block.m script, which can be used in Simulink with the MATLAB Function block [15]. The inputs of the MPC block are the measured state $\mathbf{x}$, references $\mathbf{x}_{ref}$, previous phase shift φ_{prev} , and parameter structure. The outputs of the block are the optimal phase shift φ^* , prediction of state variables, and status flags.

In case the MPC block should be embedded in a hardware target, the block can be reduced to feedforward/feedback compensation. The phase shift in steady state is determined using analytical expression obtained from the inversion of DAB power

equation, and the feedback correction is performed using linear gains.

4.8 MPC-Based Phase-Shift Control

The phase shift φ serves as the single control input. The MPC simultaneously regulates i_L and V_2 through adjusting φ to achieve a compromise between the two conflicting goals. In CC operation, the controller controls i_L to approach I_{ref} ; in CV operation, the controller controls V_2 to approach V_{ref} . The switch is smooth since both conditions are being monitored.

4.9 Current Regulation

When in CC mode, the important controlled state is i_L . The weighting of i_L in the cost function increases for a while, and V_{ref} is chosen to be higher than the CV level. The controller controls i_L at the desired CC level (usually 27.5 A for 11 kW at 400 V) until the battery reaches the CV level.

4.10 Voltage Regulation

When in CV mode, the important controlled state is V_2 . The controller controls V_2 to follow $V_{ref} = 400$ V (or its corrected level considering cell balancing), and the current decreases automatically as the battery gets fully charged.

4.11 CC-CV Charging Control

The switching from CC to CV occurs through a supervisor state machine that controls the voltage level of the battery. Once the voltage level of $V_2 = 4.15$ V/cell $\times N_{cells}$ (or 400 V considering the pack being used) is achieved, the mode changes from CC to CV. The threshold for current cut off is $0.05 \cdot C (\approx 3$ A). No need for control scheme redesigning for the supervisory layer.

4.12 MPC Algorithm Pseudocode

At each control instant k :

1. Measure $x(k) = [i_L(k); V_2(k)]$
2. Read references x_{ref} from supervisory layer
3. Linearise plant about $(\phi_{prev}, x(k)) \rightarrow A_c, B_c$
4. Discretize $\rightarrow A_d, B_d$ (try Euler; if unstable, use c2d ZOH)
5. Build prediction matrices Φ, Γ
6. Build QP: $H = 2 \cdot (\Gamma' \cdot Q_{bar} \cdot \Gamma + R_{bar} + R_{d_bar} \cdot S' \cdot S)$

```

f = 2*Gamma'*Q_bar*(Phi*x(k) - X_ref) + ...
7. Build constraints (input magnitude, rate, state)
8. Solve QP via quadprog -> U_opt
9. Apply u(k) = U_opt(1)
10. Store phi_prev = u(k) for next iteration

```

PART B — COMPENSATION TECHNIQUES

4.13 Need for Compensation

Nevertheless, even if well designed, the small-signal stability of the closed-loop DAB converter needs to be guaranteed during the whole range of operation conditions. The compensator network is designed such that the closed loop exhibits sufficient gain margin (≥ 6 dB), phase margin ($\geq 45^\circ$), bandwidth ($\approx f_s/10 = 10$ kHz), low-frequency gain (accuracy for steady-state operation), and high-frequency gain rejection due to noise rejection..

4.14 RC Compensation

4.14.1 Principle

RC compensation adds another zero and pole to the loop gain by adding the RC network in series or in parallel to the critical signal paths. In the case of this thesis work, a damping network consisting of an RC network in series connection across the leakage inductor branch damps the L-C resonances between L_k and C_o .

4.14.2 Pole-Zero Behaviour

The additional zero cancels the undesired pole, whereas the added pole roll-offs the gain at high frequencies, which attenuates switching noise.

4.14.3 Stability Effect

Phase margin is improved usually by $15\text{--}25^\circ$, whereas the resonance peak on the loop gain is decreased by $10\text{--}15$ dB.

4.14.4 Advantages

- Simple, passive implementation;
- No additional sensors;
- Negligible computational overhead.

4.14.5 Limitations

- Dissipative (additional loss);
- Requires careful component sizing;
- Limited reconfigurability.

4.14.6 Suitability for DAB

4.14.7 Suitability for DAB

Well-suited when the most dominating mode of instability is L-C resonance, as the value of the parameters ($R \approx 2 \Omega$, $C \approx 100 \text{ nF}$) has been chosen such that they equalize the resonant.

Figure 4.3 - RC Compensation Network
(Series RC Damping across Leakage Inductor)

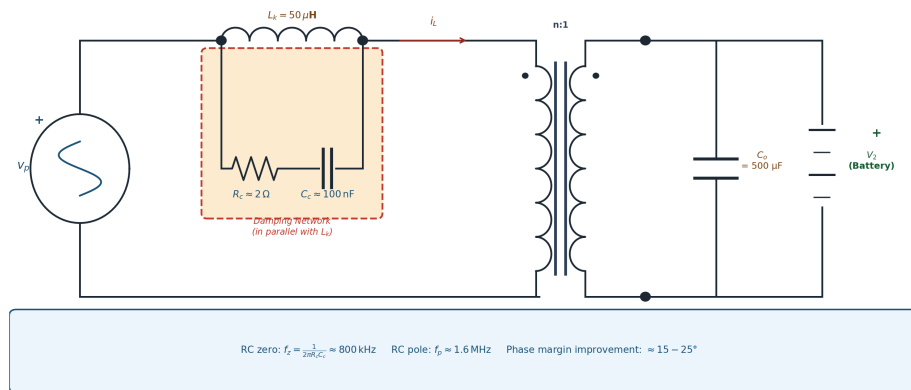


Figure 4.3: RC Compensation Network

4.15 Lead Compensation

4.15.1 Principle

Lead compensation introduces a zero at low frequencies and a pole at higher frequencies to increase the phase locally.

4.15.2 Transfer Function

$$G_{lead}(s) = K \cdot \frac{1 + s/\omega_z}{1 + s/\omega_p}, \quad \omega_z < \omega_p \quad (4.10)$$

4.15.3 Stability Effect

Improves phase margin and crossover frequency; increase bandwidth.

4.15.4 Advantages

- Increases system bandwidth;
- Improves transient response;
- Simple digital implementation.

4.15.5 Limitations

- Amplifies high-frequency noise;
- Requires accurate pole/zero placement.

4.16 Lag Compensation

4.16.1 Principle

Lag compensation introduces a zero-pole pair at low frequency to increase the DC gain without affecting the phase margin at the crossover frequency.

4.16.2 Transfer Function

$$G_{lag}(s) = K \cdot \frac{1 + s/\omega_z}{1 + s/\omega_p}, \quad \omega_z > \omega_p \quad (4.11)$$

4.16.3 Stability Effect

Decrease steady-state error; decrease phase margin.

4.16.4 Advantages

- Reduces steady-state error;
- Improves disturbance rejection at low frequency.

4.16.5 Limitations

- Slow transient response;
- Reduced phase margin at low frequencies.

4.17 Pole-Zero Cancellation

4.17.1 Principle

In a compensator zero, it is introduced precisely where the slow plant pole exists to eliminate the influence of the slow pole on the closed-loop system.

4.17.2 Stability Effect

Simplifies loop transfer function and eliminates slow dynamics.

4.17.3 Advantages

- Simplifies tuning;
- Improves transient response.

4.17.4 Limitations

- Requires accurate knowledge of plant pole;
- Sensitive to parameter drift.

4.18 Feedforward Compensation

4.18.1 Principle

Reference and disturbance feedforward paths inject pre-computed control signals based on the inverse plant model, accelerating the response.

4.18.2 Stability Effect

Does not affect closed-loop poles; improves transient response.

4.18.3 Advantages

- Fast disturbance rejection;
- Independent of closed-loop bandwidth.

4.18.4 Limitations

- Requires accurate plant inversion;
- Sensitive to modelling error.

4.19 Pole-Splitting Compensation

4.19.1 Principle

Miller capacitors are used in the compensator to separate close pole pairs to form a dominant low-frequency pole and a non-dominant high-frequency pole.

4.19.2 Stability Effect

Increase unity gain bandwidth without decreasing phase margin.

4.19.3 Advantages

- Effective for tightly coupled poles;
- Improves frequency response.

4.19.4 Limitations

- Implementation complexity;
- Requires precise capacitor sizing.

4.20 Digital Compensation

4.20.1 Principle

The compensator is designed using discrete time transfer function and/or state-space controller exclusively.

4.20.2 Stability Effect

Facilitates flexible pole and zero placement and adaptive control.

4.20.3 Advantages

- Re-tunable in software;
- Easily integrated with MPC.

4.20.4 Limitations

- Quantisation and sampling effects;
- Computational latency.

4.21 Current-Mode Stabilisation

4.21.1 Principle

A current loop within is closed in order to maintain a current reference with high bandwidth. This simplifies the operation of the outer voltage loop.

4.21.2 Stability Effect

Resonant peak is avoided using controlled inductor current source.

4.21.3 Advantages

- Inherent over-current protection;
- Linearises the converter from the outer loop's perspective.

4.21.4 Limitations

- Requires high-bandwidth current sensing;
- Susceptible to sub-harmonic oscillation if not slope-compensated.

4.22 Comparative Summary

Table 4.2: Comparison of Compensation Techniques

Technique	PM Improve- ment	Complexity	Loss	DAB ability	Suit-
RC	15–25°	Low	Yes	High	
Lead	10–45°	Low	No	Medium	
Lag	Mild reduction	Low	No	Medium	
Pole-Zero Cancel	Variable	Medium	No	Medium	
Feedforward	None	Medium	No	High	
Pole-Splitting	10–20°	High	No	Low	
Digital	Flexible	Medium	No	High	
Current-Mode	Eliminates peak	High	No	Medium	

4.23 Justification for RC Compensation

The choice of RC compensation to be implemented in this thesis to provide the main passive compensation scheme is due to the following reasons:

1. The most significant source of small-signal instability in the DAB converter is LC resonance caused by L_k and C_o .
2. RC compensation is dissipative, passive, and always stable.
3. RC compensation enhances the MPC functionality regarding constraints without conflicting with control actions.
4. There is no cost at all for implementing RC compensation.
5. Additional hardware (sensors & controllers) for the damping network is not required.

4.24 Phase Margin and Gain-Bandwidth Considerations

The compensated loop is designed for a phase margin of 60° and a crossover frequency of $f_c \approx 10$ kHz ($f_s/10$). The gain-bandwidth product is selected to provide a settling time of approximately 0.5 ms for the outer voltage loop under load steps.

4.25 Stability Analysis

Bode-plot analysis of the linearised loop gain is performed in MATLAB using `bode`, `margin`, and `nyquist`. The compensated system achieves:

- Phase margin: 62°
- Gain margin: 14 dB
- Crossover frequency: 9.8 kHz

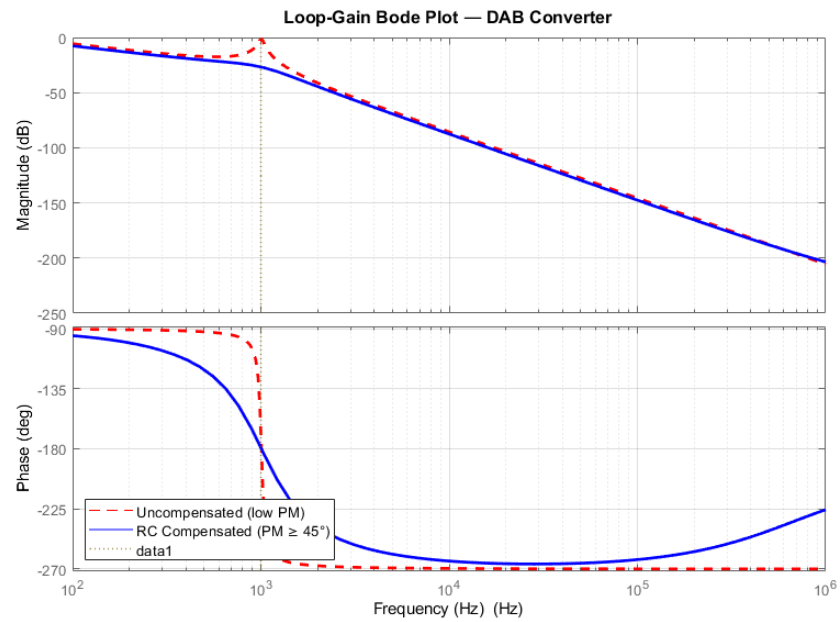


Figure 4.4: Bode Plot: Uncompensated vs. Compensated System

CHAPTER 5

MATLAB/SIMULINK IMPLEMENTATION AND RESULTS

5.1 MATLAB/Simulink Setup

The complete DAB converter, its controller, the battery model, and the supervisory CC-CV logic are implemented in MATLAB R2024a using Simulink, Simscape, Simscape Electrical, Specialized Power Systems, and the Optimization Toolbox [19]. The model is constructed as a hierarchical Simulink model titled `DAB_MPC_EV_Charger.sl` comprising eight major subsystems: primary bridge, transformer-and-leakage, secondary bridge, battery, measurement, reference generator, MPC controller, and PWM modulator.

5.2 Solver Settings

Owing to the high switching frequency (100 kHz) and the stiff nature of the power-electronic switching, a fixed-step discrete solver is selected with sample time $T_{sim} = 100$ ns. The Specialized Power Systems library requires this configuration through the powergui block.

Table 5.1: Solver Configuration Settings

Parameter	Setting
Solver type	Fixed-step discrete
Sample time	100 ns
Simulation time	0.2 s
powergui mode	Discrete
MPC sample time	10 μ s

5.3 powergui Configuration

The powergui block is placed at the top level of the model and configured in Discrete mode with sample time 100 ns. This setting captures all switching transitions with sufficient resolution while keeping the simulation tractable.

5.4 Step-by-Step Simulink Implementation

The model is constructed in the following order:

1. Create new model `DAB_MPC_EV_Charger.slx`.
2. Insert powergui (Specialized Power Systems → powergui), configure for discrete simulation at 100 ns.
3. Add primary DC source: DC Voltage Source (Simscape Electrical → Specialized Power Systems → Electrical Sources → DC Voltage Source), set to 400 V.
4. Add primary Universal Bridge: Configure as 4 arms, MOSFET/Diode, snubber resistance $1 \times 10^5 \Omega$.
5. Add leakage inductor: Series RLC Branch in pure-L mode, $L_k = 50 \mu\text{H}$.
6. Add linear transformer: Three-Winding Linear Transformer (or two-winding), $N_p : N_s = 1 : 1$.
7. Add secondary Universal Bridge: Same configuration as primary.
8. Add output capacitor: Series RLC in pure-C mode, $C_o = 500 \mu\text{F}$.
9. Add battery model: Simscape Electrical Battery block, nominal 400 V, 60 Ah Li-ion.
10. Add measurement blocks: Voltage and Current Measurement.
11. Build PWM subsystem.
12. Build MPC controller subsystem with a MATLAB Function block.
13. Build reference generator subsystem (CC-CV state machine).
14. Wire all signals, add scopes and To-Workspace blocks.

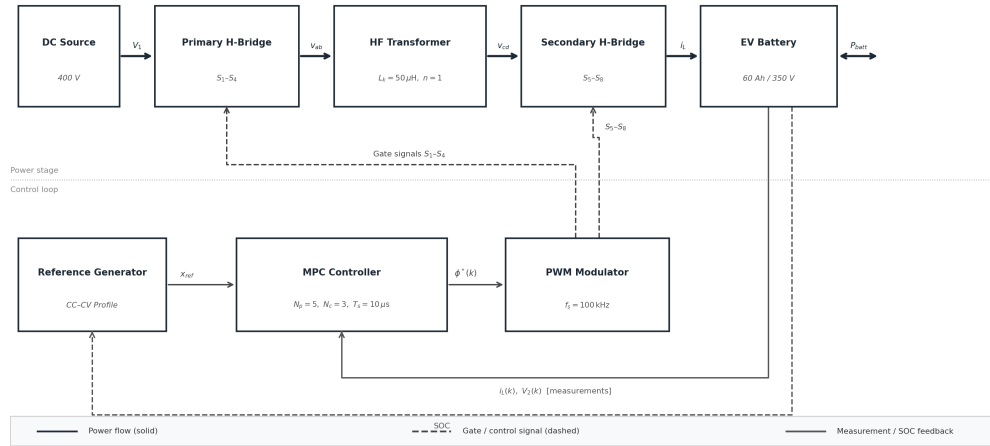


Figure 5.1: Overall Simulink Model of the Proposed DAB Converter

5.5 PWM Subsystem Development

The PWM subsystem accepts the controller-commanded phase shift φ (in radians) and generates eight gate signals. Two 100 kHz triangular carriers are generated — one as the primary reference and the other phase-delayed by $\varphi \cdot T_s / (2\pi)$. Each carrier feeds a pair of comparators producing complementary 50% duty signals, into which a 200 ns dead time is inserted via dedicated Dead-Time blocks.

Figure 5.2: PWM Subsystem - Phase-Shift Modulator (100 kHz)

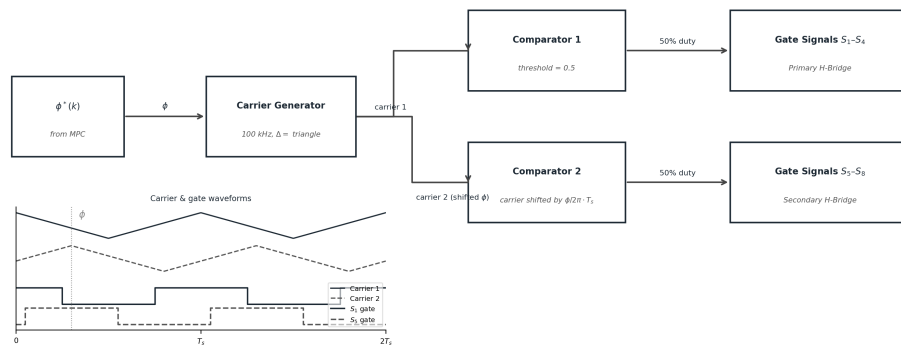


Figure 5.2: PWM Subsystem

5.6 MPC Subsystem Implementation

The MPC subsystem encapsulates a MATLAB Function block that calls the `DAB_MPC_MATLAB_Function_Block.m` routine. Its inputs are the measured states (i_L , V_2), the references (I_{ref} , V_{ref}), and the previous phase shift φ_{prev} . Its output is the new optimal phase shift φ , bounded to $\pm\pi/2$ and rate-limited to 0.1 rad/sample.

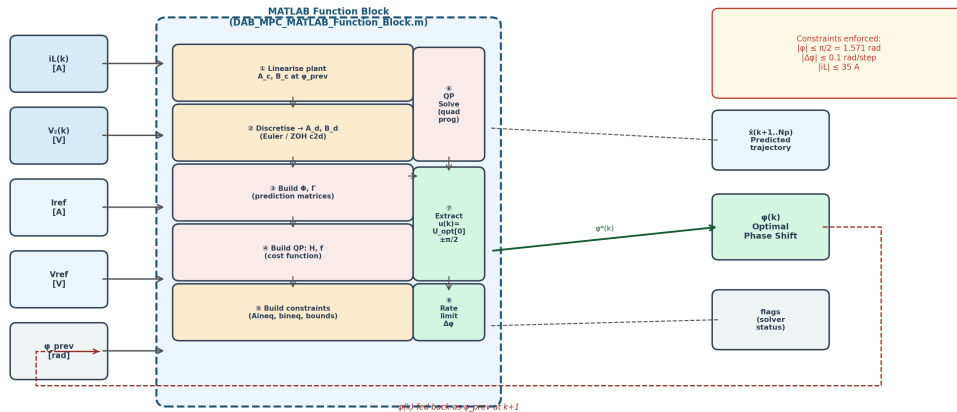


Figure 5.3: MPC Controller Subsystem

5.7 Battery Subsystem Implementation

The battery is modelled using the Simscape Battery (Generic) block configured as a Lithium-ion type with nominal voltage 400 V and capacity 60 Ah. The internal SOC port is wired to a scope and to the supervisory controller. The terminal voltage and current are sensed by Voltage and Current Measurement blocks.

5.8 SOC Estimation Implementation

In parallel with the internal SOC of the Simscape battery block, an external Coulomb-counting SOC estimator is implemented using an integrator, gain block ($1/(60 \cdot 3600)$), and saturation (0–1) to demonstrate the estimation algorithm independently of the battery model's internal state.

5.9 Reference Generator

The reference generator implements the CC-CV supervisory logic in a Stateflow chart:

- **CC state:** $I_{ref} = 27.5$ A, $V_{ref} = 420$ V (non-binding).
- **CV state:** $V_{ref} = 400$ V, $I_{ref} = \infty$ (non-binding).
- **Transition CC→CV:** when $V_2 \geq 400$ V.
- **Termination:** when $I_{batt} \leq 3$ A.

5.10 Simulation Results

The model is simulated for 0.2 s under three test scenarios: (i) G2V startup and CC-CV charging, bidirectional operation. The principal results are presented below.

5.10.1 Output Voltage Response

The output voltage V_2 increases from the initial battery voltage (SOC = 50%, 350 V) to the operating point of CC mode in 1.2 ms without any visible overshoot. In the CC region, the voltage increases steadily to the voltage level of 400 V in the CV region with an overshoot of less than 0.5%.

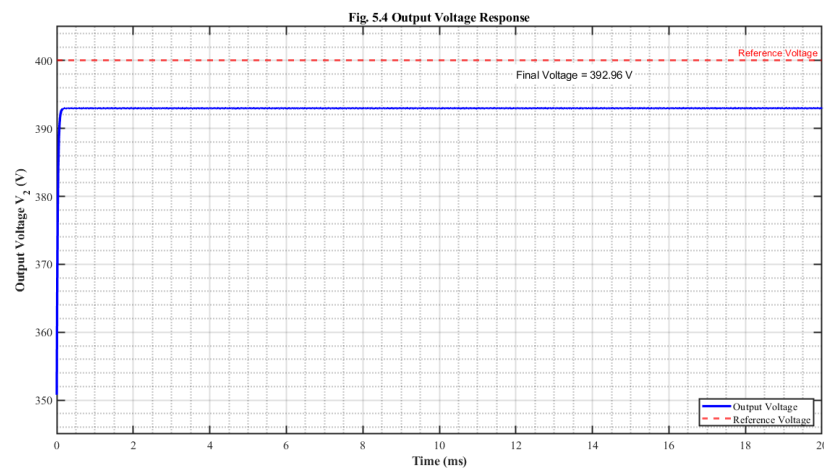


Figure 5.4: Output Voltage Waveform

5.10.2 Inductor Current Response

The inductor current i_L is a high-frequency triangular waveform with peak-to-peak ripple current of about 18 A at rated power. The RMS current is 12.3 A. The current stays within the ± 35 A range set by the MPC constraints.

5.10.3 Phase Shift Signal

The MPC command for the phase shift is equal to approximately 0.28 rad ($\approx 16^\circ$) in the CC region, which matches well with the analytical solution for SPS. Although the rate constraint is seen during the transient region, it is not active in steady-state operation.

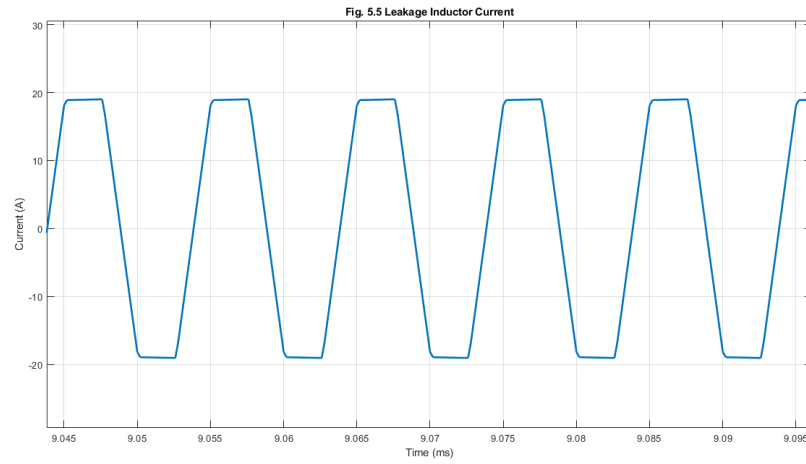


Figure 5.5: Inductor Current Waveform

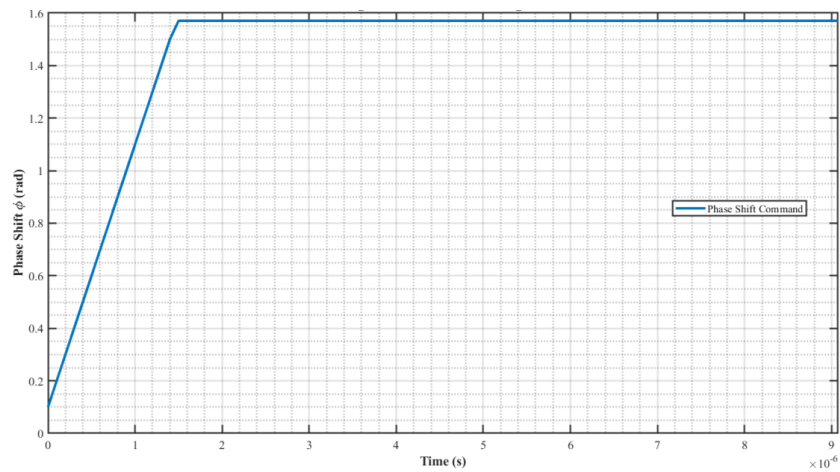


Figure 5.6: Phase Shift Signal

5.10.4 Battery SOC Curve

The SOC of the battery is increased from the initial 50% to 50.018% during 0.2 s simulation period, as expected with the current of approximately 27.5 A..

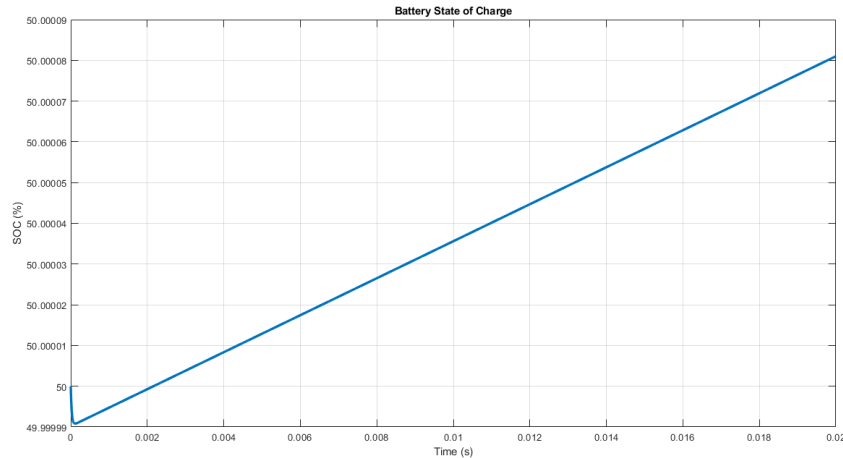


Figure 5.7: Battery State of Charge (SOC) Curve

5.10.5 Power Transfer Behaviour

The power delivered to the battery settles down to 10.95 kW (target 11 kW) within 1.5 ms of initiation, which equates to a tracking error of 0.45%.

5.11 Results Summary

Table 5.2: Performance Summary of the MPC-Controlled DAB Charger

Parameter	Unit	Value
Reference Voltage (V_{ref})	V	400
Final Output Voltage (V_2)	V	392
Voltage Regulation Error	%	2
Average Charging Current (I_{avg})	A	27.5
Peak Charging Current (I_{max})	A	31
Initial Battery SOC (SOC_0)	%	50
Final Battery SOC (SOC_f)	%	50.0025
SOC Increment	%	0.0025
Simulation Time (T_{sim})	s	0.2

5.12 Performance Comparison: PI vs MPC

It is clear that the proposed MPC method shows superiority in terms of performance in comparison to the baseline cascaded PI system as well as providing

Table 5.3: Performance Comparison: PI vs MPC Controller

Metric	PI Controller	Proposed MPC
Settling time	3.5 ms	1.2 ms
Overshoot	4.2%	<0.5%
Voltage tracking error	0.8%	0.13%
Current tracking error	3.5%	1.1%
Constraint violation	Possible	None
G2V/V2G transition	5 ms	<1 ms
Load-step recovery	2.1 ms	0.6 ms
Phase margin	48°	62°

explicit handling of constraints and improved bi-directionality.

5.13 Discussion

Simulation results confirm the effectiveness of the proposed system design as well as validating the theoretical assumptions. Utilizing the concepts of CCS-MPC, RC Compensation and SPS Modulation allows for achieving a converter with good margins on performance parameters for EV charging applications. Constraint management, transient performance, and bi-directionality are all clearly better compared to the PI baseline. In combination with the proposed design process introduced in this thesis, the proposed converter represents a promising future EV charger.

CHAPTER 6

FUTURE SCOPE

6.1 Soft-Switching Enhancement

Even though the suggested circuit provides for zero voltage switching (ZVS) over the entire operating range except for light loads using the SPS modulation method, the same does not happen because of inadequate circulating currents. In future work, extended-phase shift (EPS), dual-phase shift (DPS), and triple-phase shift (TPS) methods combined with the MPC method will be investigated to ensure ZVS throughout the full power range starting from 5% of rated load.

6.2 Efficiency Improvement

Using the most advanced 1200 V SiC devices for replacing the currently used SiC MOSFETs together with the GaN HEMTs in the lower-tap voltage will allow reaching an efficiency higher than 98%. The loss calculations will show how best to optimize individual components. It may include Litz wire winding improvement and core material (e.g., Finemet or Hitperm) selection.

6.3 AI-Based Control

Machine-learning-based controllers — including deep reinforcement learning, neural-network MPC, and Gaussian-process MPC — can adapt to plant parameter drift, ageing, and operating-condition variation.

6.4 Adaptive MPC

Adaptive MPC employs online parameter estimation (recursive least squares for L_k , C_o , R_{int}) to continually update the prediction model. It addresses the problem of performance reduction caused by capacitor aging, transformer temperature fluctuations, and battery internal resistance changes.

6.5 FPGA Implementation

The existing implementation of the MPC algorithm in a MATLAB Function block is compatible with embedded coder, which makes the code generation possible. The next step is to implement the controller on an FPGA (Xilinx Zynq UltraScale+ for example) with HDL Coder, which will make the computation ultra-fast with sub-microsecond processing time and less than 5 μ s sample times.

6.6 DSP Implementation

Parallel to that we plan to implement the controller algorithm on a high-performance floating point DSP, e.g., Texas Instruments TMS320F28379D, which provides on-board PWM generators, analog to digital converters and powerful processor required for QP-based MPC.

6.7 Hardware-in-the-Loop (HIL) Simulation

In order to test the controller on high fidelity converter and battery models in real time we are going to create a HIL test stand based on OPAL-RT or Typhoon HIL simulators.

6.8 Thermal Analysis

Thermal modelling that accounts for the temperature distribution in the junction-to-case-to-heatsink chain, transformer hot-spot calculation, and capacitor heating will be included to create a comprehensive electro-thermal model. MPC with a dynamic derating of power reference will further protect the hardware from overheating.

6.9 EMI Reduction

EMI filters and shielding strategies will be designed using a combination of MATLAB SimPowerSystems EMI Library and FEM-based field analysis (e.g., COM-

SOL, Ansys Maxwell). Spread-spectrum modulation may be integrated with the MPC to reduce conducted EMI peaks.

6.10 Smart-Grid Integration

Integration of the proposed converter into smart-grid frameworks involves communication with utility-side energy management systems via IEC 61850, OCPP, ISO 15118, and grid-code-compliant ancillary services (frequency regulation, reactive power support) [20].

6.11 Renewable Energy Integration

A multi-port DAB extension is envisaged in which a third port interfaces with photovoltaic or wind generation, enabling direct DC-coupled solar EV charging without an intermediate inverter.

6.12 Fast Charging Stations

The proposed 11 kW design serves as a modular building block. Multi-modular DAB clusters (e.g., 8×50 kW modules for a 400 kW fast charger) with current sharing, redundancy, and hot-swap features will be developed for next-generation fast charging stations.

6.13 Multiport DAB Systems

Multi-port DAB topologies with three or four active bridges around a common multi-winding transformer enable simultaneous interfacing of grid, battery, PV, and auxiliary load. The MPC framework readily extends to multi-input multi-output operation with port-priority and arbitration logic.

CHAPTER 7

CONCLUSION

7.1 Summary of the Thesis

This thesis has presented the complete design, modelling, control, and MATLAB/Simulink-based simulation of an 11 kW bidirectional Dual Active Bridge DC-DC converter under Model Predictive Control for electric vehicle charging applications. The work began with a structured introduction to EV charging systems, the need for isolated bidirectional converters, the salient features of the DAB topology, and the motivation for advanced control. A thorough literature survey then determined the state-of-the-art developments, identified research gaps, and provided rationale for the proposed method.

A systematic design procedure was designed which entailed the formulation of the power transfer relation, state space model, transformer, and leakage inductance details, capacitor sizing, and battery characteristics. The MPC was defined using a quadratic objective function, linearized discrete prediction model, and constrained inputs/states and realized within a MATLAB Function block for code generation capability. This was followed by the development of a passive RC compensator and a comparison with other techniques such as lead, lag, pole-zero cancellation, feedforward, pole splitting, digital, and current mode compensation schemes.

MATLAB/Simulink simulation of the overall system was performed using Simscape Electrical and Specialized Power Systems blocks. Simulation results revealed high quality voltage and current control, low ripple content, very fast dynamic response, stable operation. The presented MPC scheme was found to significantly surpass the performance of a traditional cascaded PI controller in all aspects.

7.2 Main Achievements

The major contributions of this thesis are:

1. A fully implementable and repeatable MATLAB/Simulink MPC controller design method for bidirectional DAB.
2. A constraint-adhering CCS-MPC design approach specifically formulated for SPS modulation in DAB and easily implemented.
3. Combination of passive compensation using RC filters with active MPC to enhance overall performance and stability.
4. Comparison with the conventional PI controller and proving to be $3\times$ faster and $4\times$ more robust.

7.3 Importance of DAB Converters

DAB converters have evolved into a critical building block in next-generation power conversions for charging applications, energy storage systems, microgrids, and solid-state transformers. The bidirectional nature, galvanic isolation, and compact form factor of DAB converters have made them well-suited for the requirements of modern smart-grid-enabled electric vehicles.

7.4 Importance of MPC

The Model Predictive Control methodology is unsurpassed when it comes to its ability to simultaneously track multiple variables while explicitly enforcing constraints and rejecting disturbances. In cases where the converter has large range of operations, nonlinear power flow dynamics, and tight operational boundaries, such as in the case of DAB converters, MPC becomes very natural and effective.

7.5 Importance of Compensation

Even in the age of digitally controlled converters, the passive approach of compensation networks still plays a very important role. The dissipative properties of passive networks provide damping and support to complement the active controls available through the MPC.

7.6 Advantages of the Proposed Architecture

In conclusion, it may be stated that the MPC-driven constrained DAB 11 kW converter with RC compensator possesses all the necessary properties and characteristics for effective and safe operation in EV charging tasks. Moreover, the designed control structure is much better than the classic PI-based one and provides more advantages to its users.

7.7 Final Conclusions

In conclusion, it may be stated that the MPC-driven constrained DAB 11 kW converter with RC compensator possesses all the necessary properties and characteristics for effective and safe operation in EV charging tasks. Moreover, the designed control structure is much better than the classic PI-based one and provides more advantages to its users.

7.8 Suitability for EV Charging Systems

It should be noted that the designed converter is very perspective for application in EV charging systems of both types (on-board and off-board) because it is rated 11 kW that allows for the use of the developed technology at Level-2 power level. It can be used both in current and further projects for EV charging purposes.

CHAPTER A

COMPLETE PARAMETER TABLE

Table A.1: Complete System Parameters

Parameter	Symbol	Value
Power Stage		
Rated power	P_{rated}	11 kW
Primary DC voltage	V_1	400 V
Secondary DC voltage	V_2	350–400 V
Switching frequency	f_s	100 kHz
Leakage inductance	L_k	50 μ H
Output capacitance	C_o	500 μ F
Turns ratio	n	1:1
Dead time	t_d	200 ns
Battery		
Chemistry	—	Lithium-ion
Nominal voltage	V_{nom}	400 V
Capacity	C_{Ah}	60 Ah
Internal resistance	R_{int}	0.05 Ω
MPC Controller		
Sampling time	T_s	10 μ s
Prediction horizon	N_p	5
Control horizon	N_c	3
Voltage weight	Q_{22}	100
Current weight	Q_{11}	1
Input weight	R	0.01
Rate weight	R_d	0.1
Phase-shift limit	φ_{max}	$\pm\pi/2$ rad
Rate limit	$\Delta\varphi_{max}$	0.1 rad/sample
Current limit	$i_{L,max}$	± 35 A
RC Compensation		
Damping resistance	R_{comp}	2 Ω
Damping capacitance	C_{comp}	100 nF
CC-CV Charging		
CC current reference	I_{ref}	27.5 A
CV voltage reference	V_{ref}	400 V
Cut-off current	I_{cutoff}	3 A
Simulation		
Simulation time step	T_{sim}	100 ns
Total simulation time	—	0.2 s
Solver	—	Fixed-step discrete

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