



TÜRKİYE CUMHURİYETİ
ADANA ALPARSLAN TÜRKER SCIENCE AND TECHNOLOGY
UNIVERSITY

GRADUATE SCHOOL
ELECTRICAL AND ELECTRONIC ENGINEERING DEPARTMENT

DESIGN AND IMPLEMENTATION OF A DC-DC CONVERTER FOR
RENEWABLE ENERGY SOURCE POWERED SYSTEMS

NİHAL BAYRAMOĞLU DİŞKEN

Ph.D.



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THESIS ADVISOR

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ADANA, 2025

DECLARATION OF CONFORMITY

In this thesis study, which was prepared following the thesis writing rules of Adana Alparslan Türkeş Science and Technology University Institute of Graduate School, I declare that I provide all the information, documents, evaluations and results in accordance with scientific ethics and moral codes without resorting to any means or assistance that would be contrary to scientific ethics and traditions. I also declare that I refer to all of the articles I used in this study with appropriate references and accept all moral and legal consequences if a situation is found contrary to my statement regarding my work.

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ÖZET

YENİLENEBİLİR ENERJİ KAYNAKLI SİSTEMLER İÇİN DC-DC DÖNÜŞTÜRÜCÜ TASARIMI VE UYGULAMASI

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Danışman: Doç. Dr. Murat Mustafa SAVRUN

Aralık 2025, 112 sayfa

Elektrik sistemleri, artan talep, arz güvenliğine ilişkin belirsizlikler ve karbon azaltım hedeflerinden kaynaklanan eşzamanlı baskılarla karşı karşıyadır. Bu durum, yerinde enerji çözümlerini modern enerji sistemlerinin başlıca odak alanlarından biri haline getirmiştir. Uygun koşullar altında üretim ve depolama birimlerinin tüketim noktasına yakın konumlandırılması, şebeke yüklenmesini azaltma potansiyeli taşımaktadır. Bu bağlamda, doğru akım (DA) mikro şebekeler, yerel üretim, enerji depolama entegrasyonu ve hassas güç yönetimine uygunlukları nedeniyle giderek önem kazanmaktadır. Bu çalışmada, söz konusu gereksinimlerden yola çıkılarak, bir DA kaynağı, bir batarya ve dengelenmiş bipolar çıkışlar arasında kontrollü güç alışverişini sağlayan yeni bir batarya bağlı bipolar DA-DA dönüştürücü önerilmiş ve deneysel olarak doğrulanmıştır. Topoloji, görev oranının batarya gücünü doğrudan düzenlediği bir batarya bağlantılı DA barası oluşturan bir yükseltici kademesi kullanmaktadır. Bunu izleyen izole kademe, toplam bipolar çıkış gerilimini faz kaydırmalı kontrol ile düzenlemekte ve kendi doğal dengeleme özelliği sayesinde bipolar çıkışlar arasındaki gerilim simetrisini korumaktadır. Topoloji, enerji depolama biriminin ilave arayüz devrelerine gerek kalmadan doğrudan bağlanmasına olanak tanımakta; doğal dengeleme özelliği ise ek algılama veya dengeleme devrelerine olan ihtiyacı ortadan kaldırarak sistemin karmaşıklığını ve maliyetini azaltmaktadır. Önerilen topoloji, yapılan simülasyon ve deneysel çalışmalarla doğrulanmış olup, portlar arasında güvenilir güç alışverişini sağlarken bipolar çıkışların dengede tutulduğunu göstermiştir. Ayrıca, doğal dengeleme özelliği hem simetrik hem de asimetrik yük koşulları altında sınanmış ve başarılı sonuçlar elde edilmiştir.

Anahtar Kelimeler: DA-DA dönüştürücüler, bipolar DA mikro şebekeler, yenilenebilir enerji

ABSTRACT

DESIGN AND IMPLEMENTATION OF A DC-DC CONVERTER FOR RENEWABLE ENERGY SOURCE POWERED SYSTEMS

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Electricity systems are confronted with concurrent pressures arising from growing demand, supply security uncertainties, and carbon reduction goals. This dynamic thereby positions on-site energy solutions as a prominent focus within modern energy systems. End-use placement of generation and storage units can mitigate network loading under suitable conditions. Within this context, DC microgrids are gaining prominence due to their suitability for local generation, storage integration, and precise power management. Motivated by these needs, this study proposes and experimentally validates a novel battery-coupled bipolar DC-DC converter that enables controlled power exchange among a DC source, a battery, and regulated bipolar outputs. The topology employs a boost stage that establishes a battery-coupled DC bus, where the duty ratio directly regulates the battery power. This is followed by an isolated stage in which phase-shift control regulates the total bipolar output voltage, while its inherent balancing characteristic maintains symmetry between the bipolar outputs. The topology enables direct connection of the energy storage unit without extra interface stages, and its natural balancing property eliminates additional sensing or balancing circuits, thereby reducing system complexity and cost. The proposed topology has been validated through simulations and experimental tests, demonstrating reliable power exchange among the ports while maintaining balanced bipolar outputs. Furthermore, the inherent balancing characteristic was verified under both symmetric and asymmetric load conditions with successful outcomes.

Keywords: DC-DC converters, bipolar DC microgrids, renewable energy

Dedication

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LIST OF ABBREVIATIONS

AC	:Alternative Current
ADC	:Analog to Digital Converter
ANN	:Artificial Neural Network
CC	:Constant Current
CCM	:Continuous Conduction Mode
CV	:Constant Voltage
DAB	:Dual Active Bridge
DC	:Direct Current
EMI	:Electromagnetic Interference
EV	: Electric Vehicle
DSP	:Digital Signal Processor
EPWM	:Enhanced Pulse Width Modulator
HFT	:High Frequency TransformerS
MOSFET	:Metal Oxide Semiconductor Field Effect Transistor
MPC	: Model Predictive Control
MPP	: Maximum Power Point
MPPT	:Maximum Power Point Tracking
PCB	:Printed Circuit Board
PI	:Proportional Integral
PV	:Photovoltaic
P&O	:Perturb and Observe
SEPIC	:Single-Ended Primary Inductor Converter
SOC	:State of Charge
SISO	: Single Input-Single Output

LIST OF SYMBOLS

T_s	:Switching Period
f_s	:Switching Frequency
n	:Transformer Turns Ratio
L_{lk}	:Transformer Leakage Inductance
V_{AB}	:Primary Bridge Differential Voltage
V_{CD}	:Secondary Bridge Differential Voltage
V_{lk}	:Voltage Across Leakage Inductance
V_{in}	:Input Source Voltage
I_{in}	:Input Source Current
L_1	:Boost Inductor
ΔI_L	:Inductor Current Ripple
C_0	:DC-link Capacitor
V_{dc}	:DC-link Voltage
ESR	:Equivalent Series Resistance of Capacitor
d	:Boost Converter Duty Ratio
V_s	:Rectified Secondary Link Voltage
V_{01}, V_{02}	:Bipolar output voltages
V_{total}	:Total Bus Voltage ($ V_{01} + V_{02} $)
C_1, C_2	:Output capacitors
R_{01}, R_{02}	:Load Resistances of Bipolar Outputs
V_{bat}	:Battery Voltage
I_{bat}	:Battery Current
r_{bat}	:Internal Resistance of Battery
P_{chem}	:Chemical Power of Battery
$P_{lossbat}$	:Internal Loss Power Of Battery
P_{iso}	:Power Transferred Through Isolated Stage
I_{iso}	:Average DC-Link Current of Isolated Stage
P_{out}	:Total Bipolar Output Power
$P_{lossconv}$	:Converter Losses Excluding Battery Losses

δ	:Phase-Shift Angle Between Primary And Secondary
α	:Effective Excitation Factor Of Isolated Stage
e_v	:Voltage Control Error
e_i	:Current Control Error
V_{bat}^*	:Reference Battery Voltage
I_{bat}^*	:Reference Battery Current
d_{pi}	:Duty Ratio Command From PI Controller
d_{min}, d_{max}	:Minimum And Maximum Duty Ratio Limits
V_{pv}, I_{pv}	:Photovoltaic Voltage And Current
P_{pv}	:Photovoltaic Power
S_0	:Boost Converter Switch
D_0	:Boost Converter Diode
S_1-S_4	:Primary Bridge Switches
S_5-S_6	:Secondary Bridge Switches
D_1-D_4	:Secondary Diodes

1. INTRODUCTION

Rising global electricity demand, the spatial decentralization of generation, and the global pursuit of energy system decarbonization accelerate the adoption of modular and distributed power systems where energy is consumed close to its source [1], [2]. Direct current (DC) microgrids have emerged as a suitable organisational paradigm for such systems, as they natively interface renewable power source and electrochemical storage within compact, scalable architectures. By coupling DC sources, storage units, and DC-native loads without unnecessary transitions, DC microgrids reduce conversion stages, elevate end-to-end efficiency, and simplify supervisory control; critically, they dispense with frequency synchronisation and reactive-power management that are intrinsic to alternating current (AC) networks [3],[4].

Within DC distribution, bus topology is typically classified as unipolar or bipolar [5],[6],[7]. The unipolar arrangement is attractive for its straightforward protection and wiring practice, yet it furnishes a single voltage level and therefore tends to proliferate point-of-load converters, inflating losses and design complexity. In contrast, bipolar architectures simultaneously supply multiple voltage levels and furnish a neutral reference that facilitates fault localisation and selective reconfiguration, enhancing availability and operational flexibility [8],[9],[10],[11]. When aligned with the inherently DC nature of photovoltaics, batteries, and fuel cells, these attributes can curtail conversion overheads and streamline distribution design, contingent on operating constraints and component choices [12], [13] [14],[15].

Realising these benefits in practice imposes stringent requirements on the intervening DC–DC interfaces. Converters must (i) maintain tight output voltage regulation with low ripple against source and load variability [16], [17], [18] ; (ii) include a power interface that facilitates controlled charging and discharging of storage elements [19], [20], [21] ; and (iii) achieve high power density with modest device count and tractable control to reduce bill-of-materials and promote predictable electromagnetic behaviour [22]. Conventional single input-single output (SISO) stages, such as unidirectional boost or buck converters, satisfy these requirements only partially; this limitation has motivated extensive research into multiport, hybrid, and both isolated and nonisolated converter topologies tailored to bipolar DC microgrids [23], [24], [25].

Across these proposals, a persistent technical concern is sustaining voltage symmetry between the positive and negative buses under dynamic or asymmetric operating conditions [26]. Deviation from symmetry induces neutral shift, degrades power quality, and exacerbates component stress, prompting recent research to integrate balancing mechanisms with primary regulation and bidirectionality while avoiding auxiliary hardware or layered control schemes [27], [28].

In this context, the present study introduces a novel converter architecture for bipolar DC microgrids that integrates voltage boosting, controlled energy exchange for battery charging and discharging, and intrinsic bipolar voltage balancing within a compact and efficient design. The proposed system maintains voltage symmetry under both symmetric and asymmetric load conditions without additional balancing circuitry, ensuring stable operation. By enabling efficient battery charge/discharge control alongside intrinsic voltage regulation, the design enhances overall system reliability and facilitates the scalable integration of renewable energy and storage into DC microgrids.

1.1. Research Motivation

Maintaining balance between the positive and negative buses under varying operating conditions remains the central challenge in bipolar DC distribution [26], [29], [30]. Divergence of the two outputs causes neutral shift and pole mismatch, degrading power quality and increasing device stress; therefore many studies target balance while avoiding additional hardware or elaborate supervisory layers [27], [28],[29], [31] . Although recent efforts embed balancing alongside tight regulation, storage interaction, and compact hardware, these solutions often introduce multi stage control, dedicated balancing circuits, or auxiliary storage paths that increase design overhead and impede scalable deployment [32], [33], [34], [35].

Within this landscape, nonisolated converters with active balancing improve symmetry during dynamic loading but typically depend on multiple sensors, nested feedback loops, or high speed digital platforms [24], [36]. Advanced techniques like average current mode control and model predictive control (MPC) have shown high regulation and fast transient response, as demonstrated in several studies [37], [12], [38], [39]. Despite their effectiveness, these methods

impose complexity in terms of processing requirements, synchronization, and loop design, which can limit their adoption in compact or cost-sensitive platforms. Hybrid realizations based on SEPIC–Cuk combinations or voltage multiplying rectifiers can reduce footprint yet add magnetic complexity [40], [41]. Center tapped transformer approaches with equal turns are simple but offer limited flexibility, often requiring manual tuning or passive filtering, and they restrict power exchange among generation, storage, and loads [42], [11]. Moreover, a considerable portion of the literature remains simulation heavy with limited validation under realistic dynamics [13], [43], [9], [15]. Where hardware implementations are explored, factors such as higher switch counts, multiple control loops, and additional processing stages can pose challenges for compact or cost-sensitive applications.[44], [45].

Motivated by gaps in prior work, this study presents a compact converter architecture that inherently maintains symmetry between the bipolar buses while coordinating energy exchange with the battery. Experimental validation under both symmetric and asymmetric loading confirms stable regulation and reliable power flow control. The proposed platform integrates a boost front end and a phase shift controlled isolated stage to regulate power among the source, battery, and bipolar outputs. Control is coordinated through a unified scheme that governs battery side and bus level states, preserving bus stability and pole symmetry without per pole sensing or dedicated balancing hardware.

1.2. Scope of the Thesis

This thesis aims to develop a compact and efficient power conversion platform tailored for bipolar DC microgrid applications, integrating energy storage and renewable energy sources within a simplified yet functionally rich topology. To this end, a novel converter architecture is proposed that enables both battery energy management and photovoltaic (PV) maximum power point tracking (MPPT) through a single active switch, representing a significant reduction in circuit complexity compared to conventional solutions in the literature.

The converter facilitates regulated bipolar DC output without requiring per-pole voltage sensing or dedicated balancing loops, thereby minimizing control effort and hardware overhead. Unlike traditional designs relying on multi-stage architectures, the proposed system leverages an

integrated structure capable of coordinating power exchange among the source, battery, and bipolar bus under both symmetric and asymmetric load conditions.

To evaluate the effectiveness of the proposed topology, a comprehensive design and verification process has been carried out. A 1 kW system is modeled and simulated using MATLAB/Simulink to analyze steady-state and dynamic behavior under various operating scenarios. Subsequently, a laboratory-scale hardware prototype is developed to experimentally validate key performance metrics including voltage balance, transient response, and power flow regulation. The system is digitally controlled using a DSP-based controller, enabling real-time implementation of the proposed regulation strategy with minimal computational burden.

Through both simulation and experimental investigations, the converter's ability to maintain stable operation and effective power management is demonstrated, highlighting its potential as a compact, scalable, and cost-effective solution for emerging bipolar DC microgrid infrastructures.

1.3. Contributions of the Thesis

The core scientific and technical contributions are outlined below.

- A novel DC-DC converter topology is proposed for bipolar DC microgrids, in which the battery is directly coupled to the power stage without requiring a separate interface converter. This integration reduces the total number of circuit components and simplifies the overall system design compared to conventional multi-stage or hybrid architectures.
- Due to its inherent design, the proposed converter naturally maintains voltage balance between the positive and negative buses in bipolar DC microgrids, eliminating the need for dedicated voltage sensing on each pole or additional active balancing circuitry. This feature ensures output voltage symmetry under varying load conditions without introducing additional control complexity.

- A simple yet effective control strategy is employed, consisting of two independent regulation loops: one that controls the battery current and another that regulates the total bus voltage. The control algorithm is implemented on a DSP platform and avoids computationally intensive methods, which contributes to easier implementation in practical applications.
- The operational performance of the proposed converter topology and control strategy is validated through simulation studies conducted in MATLAB/Simulink, as well as experimental results obtained from a hardware prototype developed and tested under laboratory conditions.

1.4. Outline of the Thesis

The thesis is structured into seven chapters that collectively present the theoretical framework, control design, and experimental validation of the proposed converter system.

Chapter 1 introduces the background, motivation, and objectives of the study.

Chapter 2 provides a detailed literature review, summarizing previous work on converter topologies, control strategies, and system integration while identifying key research gaps.

Chapter 3 develops the power circuit modeling, formulating the steady-state and dynamic equations that describe the system behavior.

Chapter 4 develops the overall control framework, defines the primary objectives and architecture.

Chapter 5 reports the simulation study carried out in MATLAB/Simulink to evaluate the converter performance under diverse operating conditions.

Chapter 6 describes the experimental validation based on a laboratory scale hardware prototype, confirming the accuracy of the theoretical and simulation results.

Finally, Chapter 7 concludes the study by summarizing the main findings and suggesting directions for future research.



2. LITERATURE REVIEW

2.1. Topology Based Review

The design of DC-DC converters is primarily defined by its topology, which determines how power is processed, transferred, and managed across multiple inputs and outputs. The selection of a suitable topology directly impacts conversion efficiency, power density, isolation requirements, and system scalability [46],[47]. Different topological structures are employed to optimize power-sharing flexibility and minimize losses. Additionally, factors like component count, electromagnetic interference (EMI), thermal management, and integration with renewable energy sources influence the choice of topology.

2.1.1. Non-Isolated DC-DC Converter Topologies

Non-isolated DC-DC converters have emerged as a fundamental building block in modern power electronics due to their structural simplicity and suitability for a broad range of low and medium power applications [48], [49]. Unlike isolated converters, they do not incorporate galvanic separation between input and output, which results in reduced component count, improved power density, and lower manufacturing costs. These characteristics make them particularly attractive for integration into electric vehicles (EVs), photovoltaic (PV) energy systems, energy storage interfaces and DC microgrids [50], [51], [52].

Advantages:

- Higher power density due to the absence of bulky isolation transformers [53].
- Lower cost and compact size, making them ideal for space-constrained systems [54].
- Increased efficiency in single-stage energy conversion [53], [55].
- Simpler control strategies compared to isolated counterparts [56], [57].

Disadvantages:

- Lack of galvanic isolation may pose safety and noise immunity concerns [58].
- Limited protection against ground loops and electrical faults [58].
- Less suitable for applications requiring strict isolation between input and output [58].

One of the most extensively studied groups in the literature is the boost derived topologies, which are modifications of the conventional boost converter to enable high voltage gain without operating at extreme duty ratios. These include voltage lift techniques, energy recycling paths, and auxiliary passive networks [59], [60], [61], [62], [63]. The study [64] introduced a voltage lift enhanced quadratic converter achieving over $16\times$ gain at moderate duty cycles, while [65] proposed a high gain hybrid boost converter integrated with switched capacitor cells, achieving over $10\times$ gain and 97.1% efficiency with reduced semiconductor stress. In [66] a dual input Z quasi resonant boost derived converter delivers high gain and lower switching stress, at the cost of more components and control complexity. The study in [67] presents a bidirectional multiport boost derived converter that enables flexible power flow, though multiple ports can reduce efficiency and complicate coordination. In [68], a single-switch trans-inverse boost derived converter is proposed, which achieves high voltage gain with low switch stress. However, its complex magnetic structure and precise control requirements limit practicality. [55] introduced a dual-input triple-output high step up converter offering flexible power sharing and reduced voltage stress. On the other hand its multiport design increases component count and control complexity. In [69], a non isolated dual input boost derived high step up converter with a bidirectional port and continuous input current is presented. It achieves high voltage gain and continuous input current without coupled inductors; however, the multi switch control increases circuit complexity and switching losses.

Quadratic, super lift, and cascaded converters enhance voltage gain by employing multi stage energy transfer and voltage lift mechanisms [70], [49]. In [71], a quadratic converter with integrated voltage multiplier cells is introduced to further elevate gain and reduce switch count. Additionally, in [72], a single inductor multi input multilevel high step up converter based on

switched diode capacitor cells is proposed. This design achieves high voltage gain and compact structure with fewer inductors. In [73], a quadratic converter using a coupled inductor and two cascaded semi stages is proposed. It achieves high voltage amplification and maintains continuous input current. In the study [74], a hybrid high step up converter integrating quadratic and switched capacitor networks is presented. The design enhances voltage boosting capability and minimizes semiconductor stress. Common to Sharma [72], [73], and [74] are increased component counts from quadratic or diode-capacitor cells that raise size and cost, plus more involved control/coordination (e.g., multi input timing and capacitor voltage management), which complicate practical implementation. In another study in [75], a quasi-Z-source multiport converter is proposed, achieving high voltage gain through multi-stage energy transfer and step-up cascaded operation while enabling bidirectional power flow for battery charging and discharging. In [76] a P type modified quadratic buck boost converter provides high gain and high efficiency with fewer passive components. However, higher stress on the output diode limits its long term reliability. Switched-capacitor (SC) and switched-inductor (SI) converters presents notable gain improvements with minimized magnetic components. In the study [77] , a non isolated multi input high step up converter that integrates coupled inductors, a voltage multiplier, and switched capacitor cells is presented. It delivers high voltage gain and soft switching with fewer parts, yet the structure is more complex and the physical size is larger. In [78], a non isolated high step up converter employing the switched capacitor method is proposed. It achieves high voltage gain with a single switch and simple structure while maintaining low voltage stress on semiconductor devices. But the multiple capacitor network increases component count and limits scalability for higher power applications.

Among non-isolated multiport converters, coupled-inductor based architectures have gained substantial attention due to their ability to achieve high voltage conversion ratios without resorting to extreme duty cycles. The magnetic coupling between windings enables energy transfer through mutual inductance, which effectively multiplies the voltage gain while maintaining compactness. The study in [79] showed this principle with a single coupled-inductor three port converter providing bidirectional flow and soft switching. The studies [80] and [81] extended this concept by combining multi-winding coupled inductors with voltage-multiplier cells to further enhance voltage gain and reduce conduction losses. In [82], a three

port high gain converter employing a coupled inductor was proposed to achieve soft switching and lower device stress, though at the expense of added magnetic design complexity. However, these designs increase magnetic complexity, leakage related spikes, and parameter sensitivity, making control and optimization more challenging.

An integrated view of recent non-isolated bipolar DC-DC converter research reveals that authors approach the same design goals, including bus balancing, bidirectional energy exchange, and efficiency, from markedly different angles [83]. In [84], pioneered multi-input multi-output converters that route energy among several sources and a backup battery, achieving flexibility but suffering from numerous switches and intricate coordination. Building on the need for stability, study [85] proposed a three port multidirectional buck-boost converter emphasizing pole-voltage balancing through active control, a structure that improves symmetry yet offers only moderate voltage gain. In [86], this concept was refined through single switch variants that reduce gate drive complexity while restricting modulation flexibility. In [87], a dual output high gain interface linking photovoltaic and fuel cell sources directly to bipolar buses without transformer isolation was introduced, marking an efficiency milestone but requiring bulky passive components for wide range operation. In [88], a bipolar bidirectional converter with an integrated energy storage port was presented to stabilize bus voltage under fluctuating renewable input. This configuration improved dynamic performance and voltage regulation but required additional sensing and more complex bidirectional control loops.

Capacitive gain cells with simple gating elevate the conversion ratio at low cost, as in [53], which favors compact hardware yet limits agility during rapid transients. Network level interconnection appears in a Cuk based bidirectional link in [89] that rebalances voltages between two bipolar networks but introduces many operating modes and nontrivial drive coordination, while the quadratic interlink in [90] unifies unipolar and bipolar buses with inherent symmetry yet remains mainly supported by simulation because of its five switch complexity. Efficiency oriented designs employ coupled magnetics and related gain cells to reduce ripple and device stress across wide operating ranges. The single switch common ground design in [91] combines a coupled inductor with a voltage multiplier to cut device stress and reaches about ninety four percent efficiency, whereas the bipolar coupled inductor converter in

[92] secures continuous input current and soft switching but requires careful clamp design and precise winding construction. For residential scale systems, a wide gain single switch buck boost realization in [93] deliberately trades higher current stress at the extremes and several inductors in the passive set for low cost and straightforward control. At low voltage distribution levels, a switched inductor SEPIC Cuk hybrid in [12] demonstrates practical energy sharing with manageable control overhead. In [94], a dual input dual output multiport converter is introduced that routes power among multiple sources and delivers several output levels. It attains high voltage gain at small duty ratios and keeps device voltage stress low, which improves efficiency and suits hybrid systems with diverse source and load requirements. In [95], the four port architecture raises gain by inserting inductor-capacitor-diode and diode-capacitor cells, enabling wide conversion ratios without extreme duty commands. However, the component count is relatively high. In contrast to the designs in [94] and [95], the converter in [96] achieves similar objectives with fewer parts.

A broad view of non isolated converter research shows a diverse set of options that trade cost, control simplicity, power density, and efficiency in different ways. There is no single best solution; the appropriate choice depends on the source and load mix, required bidirectionality, allowable component count, and target voltage range. Remaining challenges include managing mode transitions, limiting leakage related transients, and maintaining stable control under asymmetric loading.

2.1.2. Isolated DC-DC Converter Topologies

Isolated DC-DC converters represent a critical class of power conversion technologies that ensure galvanic isolation between the input and output stages, primarily through high frequency transformers [97],[98],[99]. This isolation enhances safety, improves electromagnetic compatibility, and enables modular design in complex power architectures [100], [101], [102]. These features are essential for systems where voltage domains must be physically separated for operational reliability and user protection. As such, isolated converters are widely employed in grid-tied renewable energy systems, electric vehicle (EV) onboard chargers, industrial power supplies, and telecommunication infrastructure, where isolation is either required by regulation or necessitated by design constraints [103], [104], [105].

Advantages

- Galvanic isolation enhances safety in high-voltage applications, protecting users and sensitive circuits from faults and electric shock [56],[106].
- Improved electromagnetic compatibility, making isolated converters more robust in noisy environments such as industrial or renewable energy systems [107], [106].
- Support for multi-port, bidirectional energy transfer, allowing integration of PV, batteries, and grid with flexible power flow control [106].
- Capability for high voltage gain, which is critical in renewable energy systems requiring large step-up ratios [106].

Disadvantages

- Larger size and higher cost due to the inclusion of transformers and isolation circuitry [107], [108].
- Lower power density, making them less suitable for compact or portable applications [108], [109].
- Higher conduction and switching losses, especially without soft-switching techniques, reducing overall system efficiency [108].
- More complex control algorithms are required for managing multiple ports, isolation states, and bidirectional flow [108].

Isolated topologies often use high frequency transformers and voltage multiplier cells to realize high voltage gain. Full bridge variants in [110], [111], [112] target higher power transfer, higher output voltage, and a broader soft switching range. Study [110], employs a boost interleaved full bridge with two inputs to smooth input currents and widen the soft switching window. Study [111], adopts a similar structure and introduces a six switch active rectifier that lowers

secondary device stress. Study [112], integrates voltage multiplier cells into a buck or boost interleaved full bridge to raise gain. The use of several discrete magnetic elements limits the overall power density even under soft switching conditions. High frequency transformers provide galvanic isolation, enable high gain, and help mitigate leakage currents [113]. In [114], a quasi Z source converter with a high frequency transformer is proposed for interfacing small renewable sources with the grid, reporting low semiconductor stress, a high boost factor, and reduced losses, with a passive output rectifier that lowers cost and size. In [115], a three port current fed converter uses a secondary voltage multiplier and a single source inductor to improve gain while reducing size and cost. The dual active bridge underpins many isolated converters and enables bidirectional power flow with soft switching and high power density [116],[117]. In [118], a three port isolated converter based on this architecture targets high gain, galvanic isolation, and efficient photovoltaic battery interfacing. In [119], an active clamp with two boost stages forms a three port converter in which transformer leakage contributes to soft switching, while the use of two transformers increases size. Study [120] presents non isolated and partly isolated three port three level variants that achieve few switches, modest magnetic core requirements, and high gain. In [121], adding two switches to a conventional forward converter enables bidirectional operation in a compact structure. In the study [122], a series capacitor and dual transformer configuration is introduced on the primary side to reduce switching voltage stress and achieve zero voltage switching over the entire operating range. On the other hand, the presence of two magnetic components increases implementation complexity and design effort. In [123], inductors placed on both the input and output ports, together with direct current control, reduce high frequency current harmonics and inherently support short-circuit protection. In [124], a dual-switch quasi-Z-source network combined with a voltage-multiplier cell achieves high step-up gain. In [125], a primary full bridge with phase shifted control regulates power transfer, while a bridgeless Luo arrangement on the secondary reduces output ripple and limits switch voltage stress.

The recent literature on isolated DC-DC converters places strong emphasis on bipolar DC microgrids, seeking automatic regulation of the positive and negative buses while preserving high conversion efficiency [126], [127], [128]. In [129], a phase shifted full bridge fed by a three phase active front end is combined with a buck boost balancing stage to preserve voltage

symmetry under load asymmetry in bipolar buses. In [130], a full bridge converter is combined with a bipolar voltage multiplier to realize inherent interpolar self balancing in an isolated architecture. The arrangement preserves conventional drive and attains soft switching with zero voltage and zero current commutation. The drawback is more passive parts and higher capacitor stress. In [131], the current fed bipolar converter shifts the balancing function into the magnetic domain via appropriate coupling and magnetizing inductance design. This ensures continuous input current and maintains full range ZVS (Zero Voltage Switching) even under unbalanced loading. The topology in [132] yields symmetric $\pm V$ in one stage through a switched capacitor assisted dual active bridge. In [133], independent pole power regulation is realized with a T type dual active bridge. This inherently balances the positive and negative buses and lowers magnetic requirements. Its main limitation is sensitivity to modulation errors that can induce circulating currents and reduce efficiency. In [134], cross cycle modulation within a semiactive bridge preserves supply during monopolar faults and widens the soft switching window. However, its dual-parameter control structure adds tuning complexity. In [135], a voltage balancing dual active bridge regulates pole voltages within the isolated stage, improving response to load asymmetry without external balancers. In [136], a center tapped isolated four port converter yields symmetric bipolar outputs and sustains inter pole balance under unequal loading. Both [135] and [136], aim for natural bipolar balance through isolated multiport architectures, but this comes at the cost of increased system complexity in transformer design and control accuracy.

In conclusion, the literature on isolated DC to DC converter topologies outlines multiple credible pathways to high conversion gain, galvanic isolation, and bidirectional power transfer, including quasi Z source networks, dual active bridge families, and multiport architectures. Reported advances include higher efficiency, extended soft switching capability, and improved compatibility with photovoltaic and battery systems. Nonetheless, persistent issues remain in reducing magnetic volume, managing thermal and electromagnetic interactions, coordinating multiple ports, and sustaining stable performance over wide variations in input and load. Consequently, topology selection should be guided by application context, balancing attainable performance with size, control effort, and permissible component count.

2.2. Controller Based Review

In DC–DC converters, control algorithms are essential for maintaining stable and efficient power conversion. They regulate output voltage, control current flow, manage power sharing among multiple sources, and ensure reliable operation under varying load or input conditions [128].

To meet these requirements, researchers have developed and refined various control techniques such as conventional, nonlinear, and artificial intelligence-based approaches. Conventional methods, such as proportional integral derivative (PID) remain widely used due to their simplicity, robustness, and ease of implementation [116], [137]. Nonlinear control strategies, including sliding mode control, model predictive control, and passivity-based control, provide enhanced performance under parameter uncertainties and fast transient conditions. More recently, artificial intelligence-based controllers such as fuzzy logic, neural networks, and evolutionary algorithms have emerged to offer adaptive and data driven control capabilities, improving converter efficiency and stability in complex energy systems. Each control category presents distinct advantages and trade offs depending on the converter topology, power rating, and application environment.

2.2.1. Conventional Control Methods

PID control remains one of the most widely used strategies for DC-DC converters due to its simple structure, broad applicability, and dependable performance across changing operating conditions. In a typical arrangement the controller compares the measured output with a reference and adjusts the control parameter of the converter such as duty ratio, phase shift or reference current to minimize the error. The proportional term reacts to the present error, the integral term removes steady state offsets, and the derivative term anticipates error variations, which together enable accurate and responsive voltage or current regulation. When tuned properly, PID control achieves fast settling, limited overshoot, and acceptable steady state quality with modest implementation complexity. Its main challenges include parameter tuning over wide operating ranges, sensitivity to measurement noise, and performance limitations in strongly nonlinear regimes.

Classical PI and PID controllers are still used extensively in the control of DC to DC converters and remain a mainstay of contemporary practice due to their simple structure. Recent studies illustrate this continued prevalence. In [138], a standard digital PI regulator is implemented for a buck converter, showing low output ripple and rapid recovery during load steps. In the study [139], a conventional PID controller is applied on an FPGA based buck platform, achieving accurate voltage regulation without added algorithmic complexity. In [140], a simple PI loop is used in a SEPIC interface to stabilize hybrid renewable sources with straightforward tuning and robust operation. In [141], a basic PI controller is employed to regulate the DC bus of a Z source converter, providing fast voltage restoration under disturbances. In [142], a classical PID is compared with a fuzzy controller in a boost converter, demonstrating comparable steady state accuracy with lower computational burden. In [143], a PID controller tuning methodology is proposed for boost converters exhibiting right half-plane zeros, which cause an initial inverse response between the control action and the output voltage. In [144], a dual active bridge based self balancing converter with a coupled inductor uses a PI loop to regulate one pole and achieves automatic bipolar voltage balancing under asymmetric loads. In [145], an isolated high step down converter regulates its output with a classical PI loop and demonstrates stable operation with low ripple under load variations.

Beyond their conventional use, recent studies investigate methods to enhance PI and PID through refined tuning, digital realization, and hybrid integration. In [146], a robust PID combined with a feed forward path stabilizes non minimum phase boost converters and shortens settling time. In the study [147] a complex order PI controller is introduced for both buck and boost converters. The controller performs better than the standard PI by achieving reduced overshoot and faster settling for voltage control under dynamic conditions. Similarly, in [148], a proposed fractional PI control method achieved improved performance over the conventional PI. In the study [149], an Internal Model Control based PID maintains the simplicity of the classical design while improving disturbance rejection and stability across wide operating ranges in step up converters. Optimization driven approaches have been explored, where [150], [151], and [152] apply metaheuristic algorithms such as the Sine Cosine method, eagle inspired search, and other evolutionary searches to adapt PID gains under variable operating conditions. These studies suggest that while PI and PID controllers remain conventional in form, they can

be integrated with digital platforms, adaptive tuning, and hybrid frameworks to address requirements in renewable energy converters.

While recent studies broaden tuning practice and implementation detail, classical PI and PID controllers are still widely used in the literature for the control of DC-DC converters and frequently serve as a practical baseline. Their mathematical simplicity, predictable performance, and proven reliability make them suitable for a wide range of operating conditions and converter topologies. In conclusion, PI and PID controllers continue to provide stable, accurate, and practical control performance, preserving their relevance and usefulness in modern converter applications.

2.2.2. Artificial Intelligence Based Control Methods

Recent literature demonstrates that artificial intelligence (AI) methods are being increasingly adopted in DC-DC converter control to address nonlinear behavior, dynamic variability, and parameter uncertainty. AI algorithms are used for diverse objectives, including output voltage regulation, current sharing, fault detection, adaptive gain tuning, and MPPT in renewable energy interfaces [153]. Neural networks and adaptive neuro-fuzzy inference systems are commonly employed for their ability to approximate nonlinear converter dynamics and provide self-adjusting control laws under load and source disturbances. Reinforcement learning and evolutionary algorithms such as genetic and particle swarm optimization further enable data-driven tuning of controller parameters, improving transient response and robustness without explicit small-signal models [154], [155], [156]. Additionally, hybrid architectures integrating ANNs with conventional PI or PID controllers have been reported for PV-based boost converters and microgrid applications, yielding improved efficiency and disturbance rejection compared to traditional control schemes [157]. In [154], deep reinforcement learning agents learn duty selection for voltage regulation without an explicit plant model and improve tracking under changing load and source conditions. In [158], an adaptive neural network regulates a buck converter voltage loop with online weight updates that reduce steady-state error across conduction mode changes, while in [159], a feedforward neural controller enhances voltage regulation and fault tolerance of a boost converter in microgrid applications. For photovoltaic interfaces, [160] artificial neural network based MPPT strategies are evaluated for a solar

photovoltaic system modeled with a boost converter and a battery in MATLAB Simulink. Compared with perturb and observe and incremental conductance, the neural methods achieve faster convergence, lower oscillation near the maximum power point, and higher tracking efficiency. In [161], an Adaptive Neuro-Fuzzy Inference System (ANFIS) combined with Particle Swarm Optimization (PSO) is employed for MPPT control in grid-connected photovoltaic systems. The approach adaptively tunes PWM signals to maximize power extraction with minimal distortion, achieving around 99 % efficiency and fast dynamic response even under partial shading. In [162] an artificial neural network MPPT tuned by a Horned Lizard Optimization algorithm was developed for grid connected photovoltaic systems. The method adjusts the converter's PWM to follow the maximum power point accurately. In the study [163], dual artificial neural networks estimate solar irradiance and temperature from PV voltage and current, then an adaptive block computes the maximum power point to drive a boost converter in a DC microgrid. The tracker achieves up to 99.9 % efficiency and settles within 30 milliseconds under changing irradiance or temperature without using sensors or causing power oscillations.

In the integration of photovoltaic sources with DC-DC converters, many MPP tracking methods have been proposed, tested under varied operating conditions, and continue to develop as an active research area. Nevertheless, studies that explicitly model the interaction between the photovoltaic array and the battery within artificial intelligence based MPPT schemes remain relatively scarce. In this thesis, a complementary study has been conducted to address this gap, focusing on AI assisted control strategies that account for the coupled behavior of the PV and battery system.

3. POWER CIRCUIT MODELING

This section presents the analytical modeling and design considerations of the proposed battery-coupled bipolar DC-DC converter. The modeling is developed under continuous conduction conditions, assuming negligible switch and diode voltage drops and ideal magnetic coupling. This modeling framework forms the basis for the control design and the hardware implementation discussed in later sections.

3.1. Power Circuit Design of the Battery-Coupled Bipolar DC-DC Converter Prototype

The overall topology of the proposed converter is presented in Fig. 3.1. The system integrates a non-isolated boost stage, a directly coupled lithium-ion battery, and an isolated DC-DC stage that generates symmetric bipolar outputs.

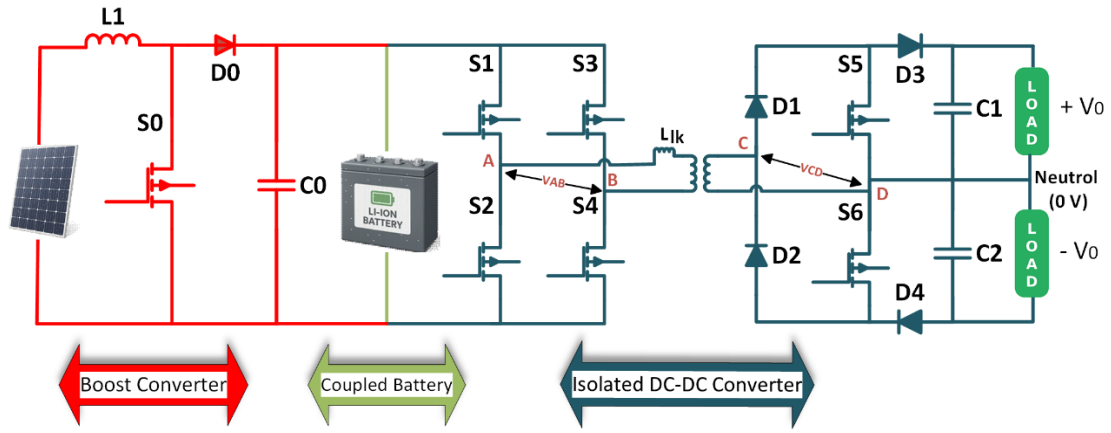


Figure 3.1. Circuit topology of the proposed battery-coupled bipolar DC-DC converter

The boost stage includes the switch S_0 , diode D_0 , and inductor L_1 , which elevate the input voltage and regulate the intermediate DC bus across the link capacitor C_0 . The battery is connected in parallel with the bus to exchange energy according to the power balance of the converter. The isolated stage contains a full bridge on the primary side (S_1 – S_4) and a clamp-assisted half bridge on the secondary side (S_5 , S_6 , D_1 – D_4), coupled through a high-frequency transformer with leakage inductance L_{lk} . The secondary network rectifies and distributes power to the bipolar outputs ($+V_0$, $-V_0$) through capacitors C_1 , C_2 and loads R_{01} , R_{02} .

This configuration provides voltage step-up, galvanic isolation, and symmetric bipolar output generation. The modeling and design of each functional block, including the boost converter, isolated DC–DC stage, coupled battery interface, and bipolar balancing circuit, are presented in the following subsections.

3.1.1. Design of Boost Converter

Boost converter operation alternates between two subintervals within each switching period T_s as seen in Figur 3.2. When the switch S_0 is turned on, the inductor L_1 stores energy from the input source, and diode D_0 is reverse-biased. When S_0 is turned off, the stored energy in L_1 is released through D_0 to the DC bus, providing a continuous current to the bus capacitor C_0 and the connected stages.

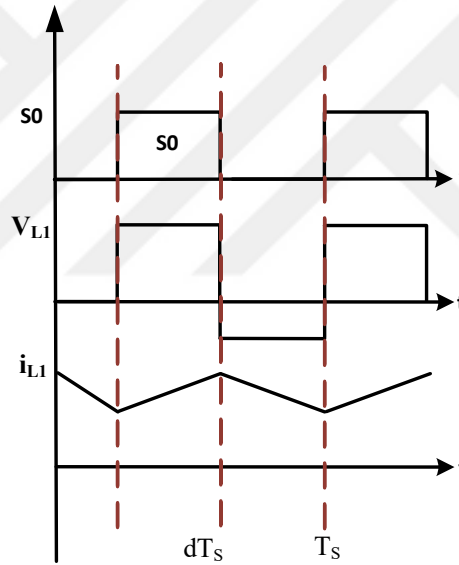


Figure 3.2. Key waveforms of the boost converter

The inductor L_1 is the key energy storage element; it determines current ripple, dynamic response, and the converter's ability to maintain continuous conduction. The instantaneous relations over one switching period are as follows:

When switch S_0 is turned on ($0 < t < dT_s$), diode D_0 becomes reverse-biased and the boost inductor is decoupled from the bus, charging directly from the source:

$$L_1 \frac{di_{L1}}{dt} = V_{in} \quad (3.1)$$

When switch S_0 is turned off ($dT_s < t < T_s$), the inductor releases its stored energy to the bus through diode D_0 :

$$L_1 \frac{di_L}{dt} = V_{in} - V_{dc} \quad (3.2)$$

where V_{dc} denotes the instantaneous voltage across the link capacitor C_0 , which represents the DC-bus voltage supplying the isolated stage and interfacing with the coupled battery. Averaging over one switching period yields,

$$V_{dc} = \frac{V_{in}}{1-d} \quad (3.3)$$

where d is the duty ratio of S_0 .

For the boost stage, the inductor is dimensioned from the inductor current ripple requirement at the worst case operating point, typically defined by $(V_{in,min}, V_o, f_s)$. In CCM, the inductor current ripple is,

$$\Delta I_L = \frac{V_{in}D}{Lf_s}, \quad D = 1 - \frac{V_{in}}{V_o} \quad (3.4)$$

The corresponding average inductor current is,

$$I_{L,avg} = \frac{I_o}{1-D} \quad (3.5)$$

A commonly used CCM design guideline is to select the inductor such that the ripple satisfies $\Delta I_L \approx 0.2-0.4 I_{L,avg}$. Imposing this constraint leads to the required inductance,

$$L = \frac{V_{in}D}{\Delta I_L f_s} \quad (3.6)$$

The peak inductor current, relevant for magnetic saturation checks and current limit design, is then,

$$I_{L,pk} = I_{L,avg} + \frac{\Delta I_L}{2} \quad (3.7)$$

The magnetic core and winding must withstand $I_{L,pk}$ without saturation, meet the RMS current requirement $I_{L,rms}$, and keep copper and core losses within acceptable thermal limits at the switching frequency f_s .

For completeness, the critical inductance at the boundary between continuous and discontinuous conduction is,

$$L_{crit} = \frac{V_{in}D(1-D)}{2f_sI_o} \quad (3.8)$$

and the selected inductance L is designed to satisfy $L > L_{crit}$ at the worst case operating point.

The boost output (DC-link) capacitor is sized by allowable ripple and transient sag. A common CCM design rule is,

$$C_o \geq \frac{I_oD}{f_s \Delta V_o} \quad (3.9)$$

3.1.2. Design of Isolated DC-DC Converter

The isolated stage transfers power from the common DC bus to the bipolar outputs through a high-frequency transformer that provides galvanic isolation and voltage scaling. The primary full-bridge (S_1 – S_4) generates the differential voltage $v_{AB}(t)$, while the secondary bridge (S_5, S_6, D_1 – D_2) synthesizes the differential voltage $v_{CD}(t)$ across the transformer secondary.

With the transformer turns ratio $n = N_s/N_p$, the leakage dynamics referred to the primary side are expressed as,

$$v_{lk}(t) = v_{AB}(t) - \frac{v_{CD}(t)}{n}, \quad \frac{di_{Lk}}{dt} = \frac{V_{lk}}{L_{lk}} \quad (3.10)$$

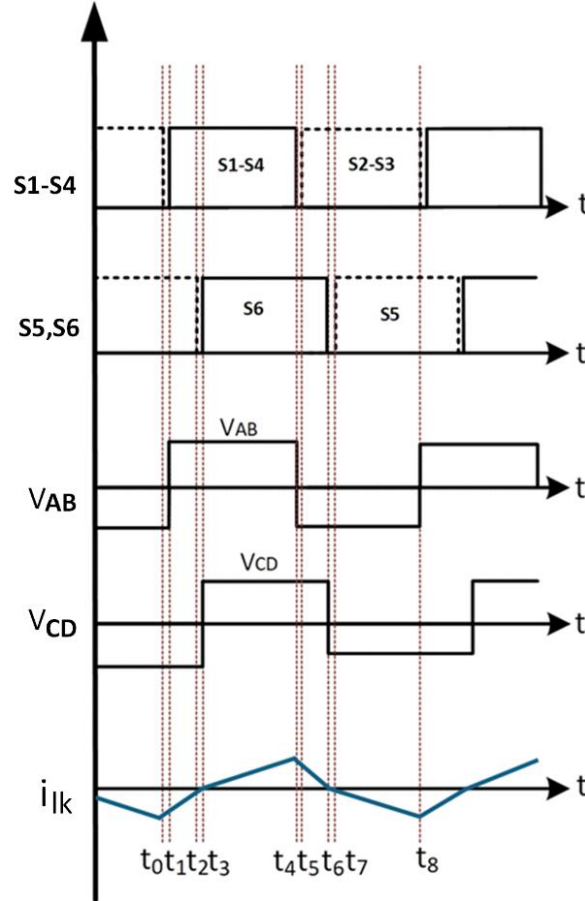


Figure 3.3. Key waveforms of the Isolated converter

The switching period T_s consists of alternating conduction stages and short commutation intervals as illustrated in Figure 3.3. DT-1, Stage 1, DT-2, Stage 2, DT-3, Stage 3, DT-4, and Stage 4. The figure explicitly shows the four conduction stages over $t_0 \rightarrow t_8$ in 3.4(b), 3.4(d), 3.4(e), and 3.4(f), together with the two commutation windows DT-1 on the primary side (Fig. 3.4(a)) and DT-2 on the secondary side (Fig. 3.4(c)). Dead times DT-3 and DT-4 are not plotted since they are time-shifted symmetric counterparts of DT-1 and DT-2.

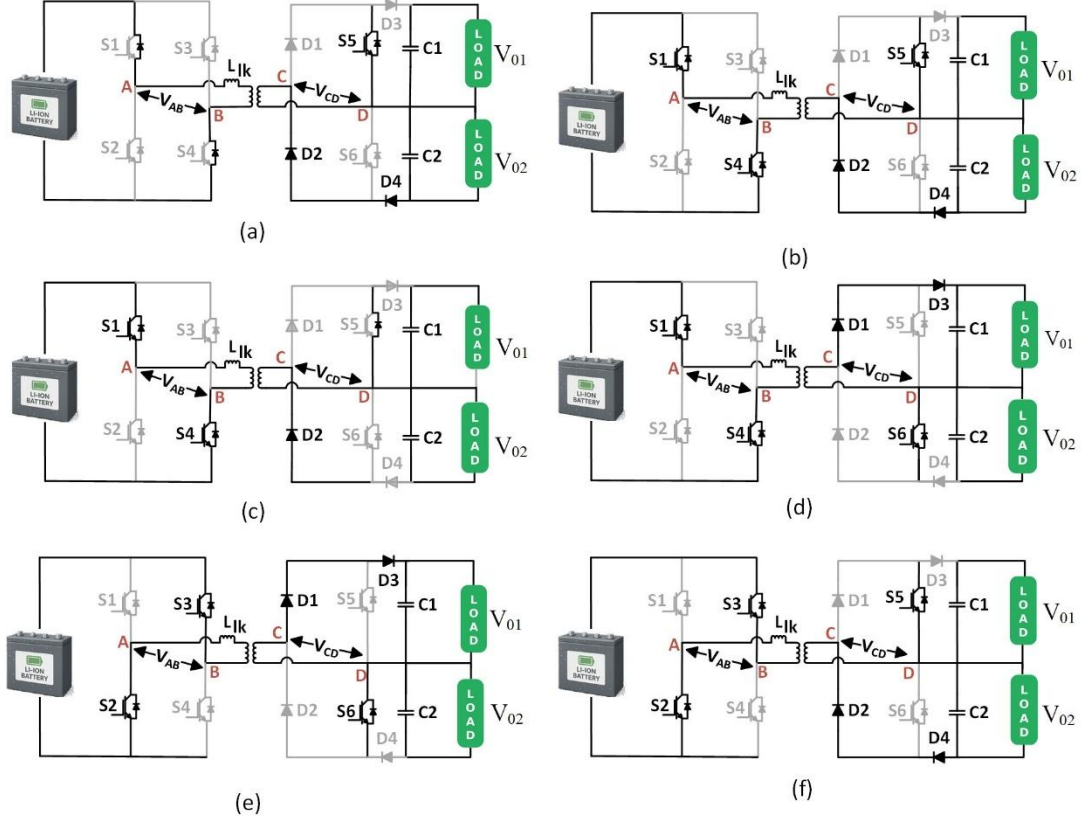


Figure 3.4. Switching states of the phase shift isolated converter over one switching period (T_s)

DT-1 [t_0, t_1]: The outgoing primary diagonal turns OFF; both primary legs are disabled during dead time. The secondary keeps its previous orientation, so $v_{CD}(t)$ keeps its sign while $v_{AB}(t)$ slews to the next level through device capacitances. Current continuity is maintained via diode paths; the instantaneous slope follows,

$$\frac{di_{lk}}{dt} = \frac{v_{lk}}{L_{lk}} \quad \text{with} \quad v_{lk}(t) = v_{AB}(t) - \frac{v_{CD}(t)}{n} \quad (3.11)$$

Stage 1 [t_1, t_2]: The primary enables the S1-S4 diagonal, establishing $v_{AB}(t) = +V_{AB}$. The secondary remains in the negative orientation (D2 and D4 forward, S5 ON), hence $v_{CD}(t) = -V_{CD}$. With opposite signs between $v_{AB}(t)$ and $v_{CD}(t)/n$, the leakage-inductance voltage satisfies $v_{lk} > 0$; i_{lk} ramps up and energy is delivered to the negative bipolar output ($-V_0$), charging C2, supplying the load.

DT-2 $[t_2, t_3]$: The secondary reverses orientation with a short dead time. During this interval nodes C and D momentarily land on the same rail, thus $v_{CD}(t) \approx 0$. Consequently $v_{lk}(t) \approx v_{AB}(t)$ and $\frac{di_{lk}}{dt} \approx \frac{v_{AB}}{L_{lk}}$ while conduction is maintained by the clamp/switch network.

Stage 2 $[t_3, t_4]$: The primary still holds $v_{AB}(t) = +V_{AB}$. The secondary is now in the positive orientation (D₁ and D₃ forward, S₆ ON), giving $v_{CD}(t) = +V_{CD}$. With same polarity between $v_{AB}(t)$ and $v_{CD}(t)/n$, $|v_{lk}|$ is reduced; i_{lk} is nearly flat (or only weakly sloped), so net incremental transfer in this sub-interval is minimal.

DT-3 $[t_4, t_5]$: The primary diagonal turns OFF; both legs enter dead time. The secondary holds its positive orientation, so $v_{CD}(t)$ preserves its sign while $v_{AB}(t)$ slews to $-V_{AB}$. The leakage current remains continuous through the clamp/diode network, with, $di_{lk}/dt = v_{lk}/L_{lk}$.

Stage 3 $[t_5, t_6]$: The primary enables the S₂-S₃ diagonal, establishing $v_{AB}(t) = -V_{AB}$. The secondary remains in the positive orientation (D₁ and D₃ forward, S ON), hence $v_{CD}(t) = +V_{CD}$. With opposite signs between $v_{AB}(t)$ and $v_{CD}(t)/n$, the leakage-inductance voltage becomes $v_{lk} < 0$; consequently i_{lk} ramps down and energy is delivered to the positive bipolar output ($+V_0$), charging C₁ and supplying the load.

DT-4 $[t_6, t_7]$: The secondary flips back with dead time; $v_{CD}(t) \approx 0$, thus $v_{lk}(t) \approx v_{AB}(t)$. The current transitions smoothly under clamp control.

Stage 4 $[t_7, t_8]$: The primary holds $v_{AB}(t) = -V_{AB}$. The secondary is back in the negative orientation (D₂ and D₄ forward, S₅ ON), giving $v_{CD}(t) = -V_{CD}$. With same polarity, $|v_{lk}|$ is reduced; i_{lk} is nearly flat (or weakly sloped). The period closes at t_8 .

Over one switching period the primary and secondary sequences synthesize piecewise constant $v_{AB}(t)$ and $v_{CD}(t)$, producing a trapezoidal leakage current $i_{lk}(t)$ whose average over T_s sets the transferred energy. Because the secondary alternates the rectified path each half cycle, the delivered charge is partitioned between the $+V_0$ and $-V_0$ branches. Under these conditions and

with small ripple on the secondary link, the rectified link voltage V_s equals the magnitude of each bipolar output, which leads to the relations,

$$\begin{aligned} |V_{o1}| &= |V_{o2}| = V_b, \\ V_{total} &= |V_{o1}| + |V_{o2}| = 2V_b, \end{aligned} \quad (3.12)$$

$$V_s = V_b = \frac{V_{total}}{2}$$

The instantaneous power transferred through the isolated stage is denoted as P_{iso} , representing the average energy per switching cycle delivered from the DC bus to the secondary through the transformer. Using the standard energy transfer model of a phase shifted isolated converter [164], [165], [166],

$$P_{iso} = \alpha \frac{V_{dc} V_s}{n\omega L_{lk}} \varphi(\delta), \quad \varphi(\delta) = \delta \left(1 - \frac{\delta}{\pi}\right) \quad (3.13)$$

where $\alpha \in (0,1)$ accounts for the semi active rectifier behavior and effective excitation reduction across the leakage inductance, and $\omega = 2\pi f_s$. The corresponding average DC bus current is,

$$I_{iso} = \frac{P_{iso}}{V_{dc}} = \alpha \frac{V_s}{n\omega L_{lk}} \varphi(\delta) \quad (3.14)$$

At equilibrium the input power of the isolated stage equals the delivered load power,

$$P_{iso} = P_{out} = \frac{V_{o1}^2}{R_{o1}} + \frac{V_{o2}^2}{R_{o2}} \quad (3.15)$$

which for the symmetric case $|V_{o1}| = |V_{o2}| = V_b$ and $R_{o1} = R_{o2} = R_b$ reduces to

$$P_{out} = \frac{2V_b^2}{R_b} = \frac{V_{total}^2}{2R_b} \quad (3.16)$$

These relations connect the phase shift command δ , the transformer parameters (n, L_{lk}) , and the secondary link V_s to the output magnitudes and load power.

3.1.3. Integration of Coupled-Battery

The battery is connected directly in parallel with the DC-bus node, providing an auxiliary energy interface that participates in bus regulation and power sharing. It is represented by its open-circuit voltage V_{bat} and internal resistance r_{bat} . The instantaneous battery current i_{bat} is defined positive when discharging into the bus and negative when charging from it. The interaction between the battery, the DC bus, and the isolated stage determines the overall power balance of the converter.

Under steady state the battery chemical power, internal loss power, and bus delivered or absorbed power satisfy,

$$\begin{aligned} P_{chem} &= V_{bat}I_{bat}, \\ P_{loss,bat} &= I_{bat}^2 r_{bat}, \\ P_{bat \rightarrow bus} &= V_{dc}I_{bat} \end{aligned} \quad (3.17)$$

from which the internal battery balance follows,

$$P_{chem} = P_{bat \rightarrow bus} + P_{loss,bat} \quad (3.18)$$

On the DC bus the power supplied by the input source through the boost path and the power exchanged with the battery equal the power absorbed by the isolated stage plus the converter losses. Under steady state with $\dot{V}_{dc} = 0$

$$\begin{aligned}
P_{in \rightarrow bus} &= V_{dc} I_{in} \\
P_{in \rightarrow bus} + P_{bat \rightarrow bus} &= P_{out} + P_{loss, conv}
\end{aligned} \tag{3.19}$$

Eliminating $P_{bat \rightarrow bus}$ via (18) yields the explicit system level power equilibrium,

$$P_{in \rightarrow bus} + P_{chem} = P_{out} + P_{loss, conv} + P_{loss, bat} \tag{3.20}$$

where $P_{loss, conv}$ includes transformer copper and core losses, semiconductor conduction and switching losses, passive ESR dissipation, and auxiliary circuits. These definitions are kept consistent with bus referred shunt losses to avoid double counting.

Two immediate corollaries follow.

(i) When the input source is disconnected $P_{in \rightarrow bus} = 0$,

$$P_{chem} = P_{out} + P_{loss, conv} + P_{loss, bat} \tag{3.21}$$

so the battery alone sustains the loads and all losses.

(ii) During charging with $i_{bat} < 0$ therefore $P_{chem} < 0$. The bus supplies both the prescribed charging power and the internal battery dissipation while the isolated stage continues to regulate the bipolar outputs through phase shift control.

3.1.4. Design of Bipolar Balancing Circuit

This section details the balancing mechanism that maintains symmetry between the positive and negative outputs with respect to the common ground. The bipolar balancing mechanism is described using averaged branch dynamics for the positive and negative rails. Let $C_{eq,1}$ and $C_{eq,2}$ denote the equivalent output capacitances, and let $i_{2,1}$ and $i_{2,2}$ be the average rectifier currents injected into nodes V_{01} and V_{02} . The per-branch dynamics are,

$$C_{eq,1} \frac{dv_{01}}{dt} = i_{2,1} - \frac{V_{01}}{R_{01}} - i_{c01,loss} \quad (3.22)$$

$$C_{eq,2} \frac{dv_{02}}{dt} = i_{2,2} - \frac{V_{02}}{R_{02}} - i_{c02,loss} \quad (3.23)$$

where $i_{c01,loss}$ and $i_{c02,loss}$ collect output-referred parasitic losses (capacitor leakage, ESR dissipation, and other shunt-type losses).

Defining the sum and difference voltages $V_{\Sigma} = |V_{01}| + |V_{02}|$ and $V_{\Delta} = |V_{01}| - |V_{02}|$, and assuming closely matched components ($C_{eq,1} \approx C_{eq,2} = C$, $R_{01} \approx R_{02} = R$), the difference behaviour is expressed as,

$$\frac{dV_{\Delta}}{dt} \approx \frac{1}{C} (i_{2,1} - i_{2,2}) - \frac{1}{CR} V_{\Delta} \quad (3.24)$$

Equation (3.24) shows that mismatches between the bipolar output voltages decay naturally. When one output becomes slightly weaker, its effective conduction interval increases. It then receives a larger portion of the average rectifier current, represented by the term $(i_{2,1} - i_{2,2})/C$ in the model. The return paths provided by D_3 and D_4 during the opposite half cycle and during dead time guarantee safe current closure and allow this limited cross flow to occur in practice. The load related term $-V_{\Delta}/(CR)$ introduces additional damping to the voltage difference. As a result, the converter exhibits a passive tendency to restore symmetry in V_{01} and V_{02} without a dedicated balancing loop.



4. CONTROLLER MODELING

The control modeling of the proposed battery-coupled bipolar DC–DC converter aims to establish a mathematical and structural framework that links the converter’s electrical dynamics with the designed control objectives. The converter operates as a two-stage hybrid system comprising a non-isolated boost and an isolated stage. The boost stage regulates the power transferred between the input source and the battery, while simultaneously supplying the isolated stage, which ensures regulated bipolar outputs at the load terminals.

4.1. Controller Design of the Battery-Coupled Bipolar DC–DC Converter

The system consists of two main stages:

- (i) a non-isolated boost stage that regulates the input power, controls the battery charge/discharge current, and establishes the DC-link voltage that supplies the isolated stage;
- (ii) an isolated stage that converts the DC-link voltage into bipolar output voltages.

Each stage is governed by its own local control loop, with the boost stage regulated through duty cycle control and the isolated stage governed by phase-shift (δ) modulation.

The primary control objectives are as follows:

- to maintain a stable DC-link voltage,
- to regulate battery charging and discharging conditions under CC and CV modes,
- to ensure symmetrical bipolar output regulation under various loading conditions.

The MPPT and CV control loops defined in this section were validated exclusively through simulation studies due to laboratory limitations. Experimental validation was conducted under the CC mode in the boost stage and bipolar output regulation in the isolated stage.

4.1.1. Boost Converter Control Design

The boost converter stage regulates both the PV input power and the battery current through the single active switch S_0 . The control variable is the duty ratio d , which determines the

average inductor voltage and hence the power transferred from the PV source to the DC bus. Under continuous-conduction operation, the averaged inductor and capacitor dynamics given in (2.6),(2.7) form the basis of the boost control design. The steady-state conversion ratio,

$$V_{dc} = \frac{V_{in}}{1 - d} \quad (4.1)$$

demonstrates the nonlinear coupling between the duty ratio and the DC-link voltage, indicating that increasing d leads to a higher input current and therefore greater power extraction from the source.

At equilibrium, the power balance on the DC bus is expressed as,

$$P_{in} + P_{chem} = P_{out} + P_{loss,conv} + P_{loss,bat} \quad (4.2)$$

which links the boost converter dynamics with the global energy exchange between the source, the converter, and the battery. Hence, by appropriate tuning of the PI parameters, the boost control loop ensures efficient power flow management. In the proposed architecture, the boost converter controller addresses two tasks: (i) battery management via CC/CV regulation to ensure safe charging and (ii) photovoltaic energy maximization via MPPT.

A PI-based CC/CV controller was adopted for battery management, while MPPT was developed and assessed using two alternative controllers: the conventional Perturb & Observe (P&O) method and a data-driven ANN-based approach. Owing to laboratory constraints the PV panel was replaced with an external DC source in experimental setup. MPPT was evaluated only in simulation, and the experimental validation on the boost converter was limited to the CC controller.

4.1.1.1. CC/CV Controller

The CC/CV charging protocol is the most widely used method for lithium-ion and other rechargeable battery systems in both research and industry [167],[168]. During the CC stage, the battery is charged with a fixed current until its terminal voltage reaches a predefined

threshold. Once this limit is attained, the control switches to the CV stage, where voltage is held constant and the current gradually decreases [169]. This two-stage approach enables fast bulk charging while preventing over-voltage stress and ensuring safe saturation of the battery [170]. The CC/CV method is commonly applied in portable electronics, electric vehicles, and stationary energy-storage systems, providing an effective balance between charging speed, energy efficiency, and cell lifetime [171].

The battery charging and discharging processes are regulated by a combined CC/CV controller implemented using cascaded proportional integral (PI) loops. This dual-loop architecture ensures fast dynamic response during the current-regulated phase and accurate voltage tracking at the constant-voltage phase.

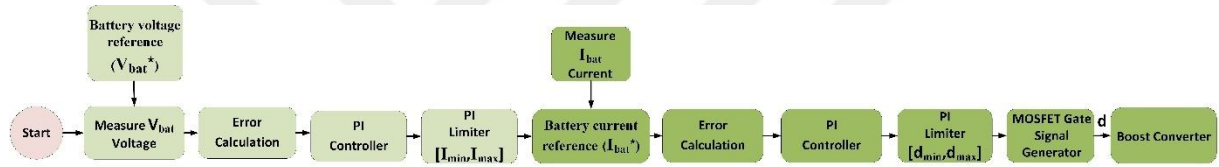


Figure 4.1. CC/CV control structure for battery charging in the boost converter stage

In the CV loop, the battery voltage V_{bat} is compared with its reference V_{bat}^* . The resulting voltage error,

$$e_V = V_{bat}^* - V_{bat} \quad (4.3)$$

is processed by a PI controller that generates a reference current I_{bat}^* for the CC current loop [172], [173], [174]:

$$I_{bat}^* = K_{p,V} e_V + K_{i,V} \int e_V dt \quad (4.4)$$

This output is passed through a limiter that constrains I_{bat}^* within the safe charging current range $[I_{bat,min}, I_{bat,max}]$.

In the CC loop, the measured battery current I_{bat} is compared with the reference I_{bat}^* . The current error,

$$e_I = I_{\text{bat}}^* - I_{\text{bat}} \quad (4.5)$$

is then fed into another PI controller:

$$d_{\text{PI}} = K_{p,I} e_I + K_{i,I} \int e_I dt \quad (4.6)$$

whose output corresponds to the duty ratio command d applied to the PWM generator. A saturation block limits d to the feasible range $[d_{\min}, d_{\max}]$ and prevents integrator wind-up. This hierarchical control structure enables seamless transition between the constant-current and constant-voltage modes:

- When the battery voltage is below V_{bat}^* , the current loop dominates, operating in CC mode.
- As the voltage approaches V_{bat}^* , the voltage loop output gradually limits the current reference, transitioning smoothly to CV mode without overshoot.

The combined PI control scheme thus ensures stable operation throughout the entire charging process while respecting battery safety constraints.

In the experimental setup, the same PI-based control strategy as in the simulation was implemented, operating exclusively in the CC mode as seen in Fig. 4.2. The duty ratio d is computed from the error e_I and generated by the DSP's ePWM module configured at the switching frequency f_s .

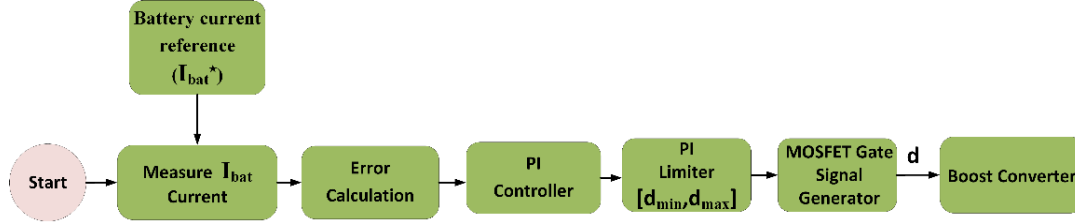


Figure 4.2. Battery current control for the boost converter

4.1.1.2. *MPPT Controller*

Two different MPPT strategies were investigated to evaluate the converter's dynamic and steady-state performance: a conventional P&O (Perturb and Observe) algorithm and an artificial intelligence based artificial neural network (ANN) controller.

- **P&O Based MPPT**

P&O method, also known as the hill-climbing algorithm, is one of the most widely implemented MPPT techniques in photovoltaic (PV) converters due to its simplicity and minimal sensing requirements.

The algorithm operates by periodically perturbing the converter duty cycle d and observing the resulting change in PV output power P_{pv} , expressed as,

$$P_{pv}(k) = V_{pv}(k)I_{pv}(k) \quad (4.7)$$

where V_{pv} and I_{pv} denote the instantaneous PV voltage and current, respectively.

At each iteration k , a small perturbation Δd is applied to the duty ratio, and the change in power is computed as

$$\Delta P_{pv} = P_{pv}(k) - P_{pv}(k - 1) \quad (4.8)$$

The slope of the power-voltage curve determines the next adjustment:

$$\text{If } \frac{dP_{pv}}{dV_{pv}} > 0 \Rightarrow V_{pv} < V_{MPP} \Rightarrow d \downarrow,$$

$$\text{If } \frac{dP_{pv}}{dV_{pv}} < 0 \Rightarrow V_{pv} > V_{MPP} \Rightarrow d \uparrow,$$

$$\text{If } \frac{dP_{pv}}{dV_{pv}} = 0 \Rightarrow V_{pv} = V_{MPP}.$$

In discrete form, this logic can be expressed as:

$$d(k+1) = \begin{cases} d(k) + \Delta d, & \text{if } P_{pv}(k) > P_{pv}(k-1) \\ d(k) - \Delta d, & \text{if } P_{pv}(k) < P_{pv}(k-1) \end{cases} \quad (4.9)$$

Here, a positive ΔP_{pv} indicates that the perturbation has moved the operating point closer to the maximum power point (MPP), and thus the same perturbation direction is maintained; otherwise, the perturbation direction is reversed. The algorithm iteratively adjusts d until the steady-state oscillation around the MPP is achieved.

- **ANN Based Controller**

Artificial neural networks (ANNs) have emerged as data-driven alternatives to classical MPPT algorithms by learning the nonlinear mapping from readily measurable electrical variables to the duty ratio that places the operating point near the maximum power point. Unlike perturbative or model-based methods, which rely on iterative searches and thus induce steady-state oscillations and delay, an ANN infers the operating point directly from instantaneous measurements, improving transient tracking under rapid irradiance and temperature variations while mitigating ripple around the MPP [160],[175].

Fig. 4.3 illustrates the structure of the proposed ANN-based MPPT controller integrated into the converter. The network consists of an input layer with four neurons corresponding to the measured quantities I_{pv} , V_{pv} , I_{bat} , and V_{bat} ; a single hidden layer containing multiple neurons labeled $H_1, H_2, H_3, \dots, H_n$, which are responsible for extracting nonlinear features and internal representations; and an output layer that produces the control signal d_{ANN} , representing the

optimal duty ratio for the boost converter. All neurons in each layer are fully connected to those in the subsequent layer, ensuring that the model can approximate complex nonlinear relationships between the photovoltaic and battery variables. Through this structure, the ANN learns to generate control commands that maximize PV power transfer while maintaining stable interaction between the source, the battery, and the DC-link bus.

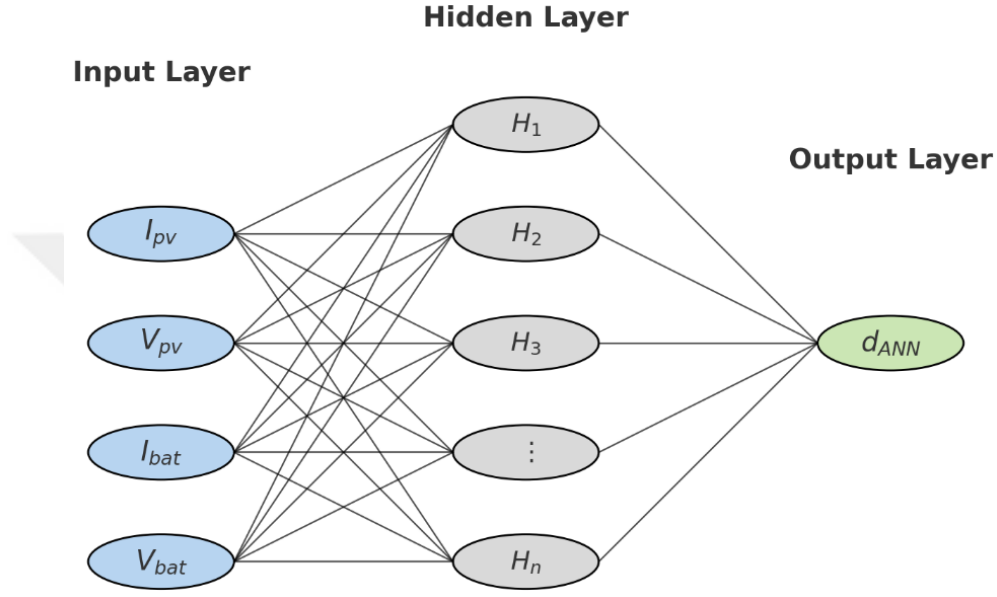


Figure 4.3. ANN-based MPPT structure integrated into the converter

In this study, the ANN is posed as a supervised regression model approximating $\mathbf{x} \mapsto d^*$, where $\mathbf{x} = [I_{bat}, V_{bat}, I_{pv}, V_{pv}]^T$ and d^* denotes the duty ratio that maximizes $P_{pv} = V_{pv}I_{pv}$ under converter constraints. Labeled pairs $(\mathbf{x}_i, d_i^*)$ are generated offline using a high-fidelity MATLAB/Simulink model by sweeping solar irradiance and battery state of charge; at each steady operating point, $(I_{pv}, V_{pv}, I_{bat}, V_{bat}, d^*)$ is recorded, with d^* supplied by a benchmark P&O routine. From a physical standpoint, minimizing the mean-squared error between the predicted and target duty ratios tunes the network so that small changes in the electrical state produce output commands aligned with increased PV power transfer; backpropagation distributes this sensitivity across layers, allowing the ANN to act after training as a compact nonlinear surrogate of the implicit control law.

A compact multilayer perceptron is adopted to balance approximation accuracy and implementation feasibility; hidden units use smooth activations for expressive yet stable

gradients, and the output layer employs a bounded activation with linear scaling to the admissible interval $[d_{\min}, d_{\max}]$. After training, the ANN output d_{ANN} is saturated and rate-limited and then applied to the PWM generator that drives the boost switch. In this architecture, the ANN replaces the iterative search layer of classical MPPT and provides a one-shot estimate of the near-optimal duty ratio, while surrounding control/protection layers enforce converter limits and ensure safe operation.

4.1.2. Isolated Converter Control Design for Bipolar Outputs

The phase-shifted isolated stage governs the power transferred from the DC bus to the bipolar outputs. By modulating the phase difference between the primary and secondary bridge gating signals, the effective excitation across the transformer leakage inductance is varied, thereby controlling the power delivered to the load. A larger phase shift increases the energy transferred per switching period, whereas a smaller shift reduces it, enabling smooth bidirectional power control within the allowed range $\delta \in [0^\circ, 90^\circ]$.

Leveraging the converter's inherent bipolar voltage self balancing property established in Section 2, the voltage loop is simplified by regulating the total bus magnitude rather than each polarity separately. The measured variable is,

$$V_\Sigma \triangleq |V_{o+}| + |V_{o-}| \quad (4.10)$$

which is compared with the reference V_Σ^* to form the error $e_\Sigma = V_\Sigma^* - V_\Sigma$. This total-voltage error is processed by the PI controller to determine the required phase-shift angle:

$$\delta_{PI} = K_{p,\Sigma} e_\Sigma + K_{i,\Sigma} \int e_\Sigma dt \quad (4.11)$$

The phase-shift command is constrained to $\delta \in [0^\circ, 90^\circ]$ in both simulation and hardware. This enforces forward power flow and preserves a monotonic relation between δ and transferred power [176], [177].

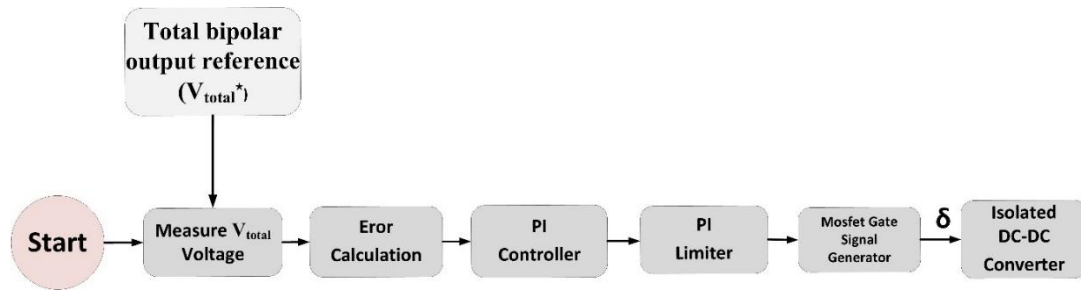


Figure 4.4. Control architecture for bipolar bus voltage regulation

In simulation, δ is applied to a MATLAB Function block that produces the gating pulses for the isolated stage. In hardware, the DSP calculates δ and sends it to the ePWM module to generate the same complementary gate signals. All switches in the isolated stage operate at 50% duty with adequate dead time.



5. SIMULATION STUDY

The simulation study plays a crucial role in validating the proposed novel battery-coupled bipolar DC–DC converter concept before experimental implementation. It enables precise observation of facilitates parameter selection and provides an effective platform to assess the efficiency and functional reliability of the proposed topology under controlled and repeatable conditions.

The main objectives of the simulation studies are:

- to verify the steady-state and dynamic performance of the proposed battery-coupled bipolar DC–DC converter,
- to determine optimal design parameters for passive components and control loops,
- to examine the converter behavior under three distinct operating cases (Case I: battery-only operation, Case II: simultaneous load supply and charging, Case III: source-deficit operation),
- to evaluate MPPT, CC/CV control strategies applied to different operating conditions,
- to investigate the system response under symmetric and asymmetric load variations,
- to confirm the converter's ability to maintain bipolar voltage regulation and stable power sharing among the input, battery, and outputs.

To accomplish these objectives, a comprehensive simulation model of the proposed system was developed in MATLAB/Simulink. The model integrates the complete converter topology, control architecture, and dynamic references. Within this framework, the system was tested under the three operational cases defined above, and its performance under MPPT, CC/CV, symmetric and asymmetric load scenarios was analyzed in detail. The overall model of the simulated system is illustrated in Fig. 5.1.

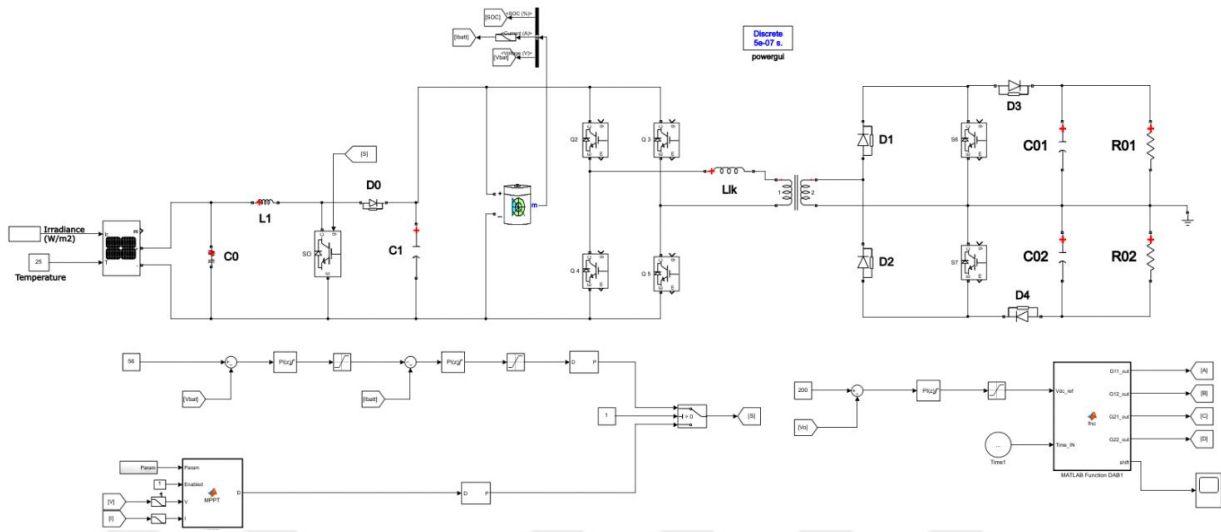


Figure 5.1. MATLAB/Simulink model of the proposed battery-coupled bipolar DC–DC converter

5.1. MATLAB/Simulink Model of the System

This study develops a MATLAB Simulink model of a battery-interfaced boost stage cascaded with an isolated DC-DC converter that delivers balanced bipolar outputs. The system is designed for approximately 1.1 kW, and all component and control parameters used in the simulations are those reported in Table 5.1. Simulations were carried out in MATLAB/Simulink R2021b with a fixed-step discrete solver; the electrical network time step was set to 100 ns.

Table 5.1. Simulation parameters of the proposed converter topology

System	Parameter	Value
Transformer	L_{lk}	2.5 μ H
	N_s/N_p	2
Isolated DC-DC converter	Switching components	MOSFET
	C_{01}, C_{02}	680 μ F
	Switching Frequency	50 kHz
Boost Converter	Switching components	MOSFET
	L_1	1mH
	C_0	2200 μ F
	Switching Frequency	50 kHz
Pv	Maximum Power (1000 W/m ²)	1.16 kW
	Maximum Voltage (1000 W/m ²)	36.5 V
Battery	Nominal Voltage	48 V
	Nominal Capacity	24 Ah
	Maximum Current	12 A
General	Bipolar Outputs (V_{01}, V_{02})	100 V, -100 V
	Output Power	450 W

5.1.1. Case I

In Case I, the converter operates entirely with the battery as the only energy source since the input power from the PV side is nearly zero. This condition represents evening or nighttime operation when solar generation is unavailable. The simulation shows that the battery delivers around 8 A of discharge current, continuously supplying energy to the DC bus and the bipolar outputs (Fig. 5.2 and Fig 5.3). As expected, the battery voltage gradually decreases, accompanied by a decline in the state of charge (SOC), which reflects normal discharge dynamics under constant load.

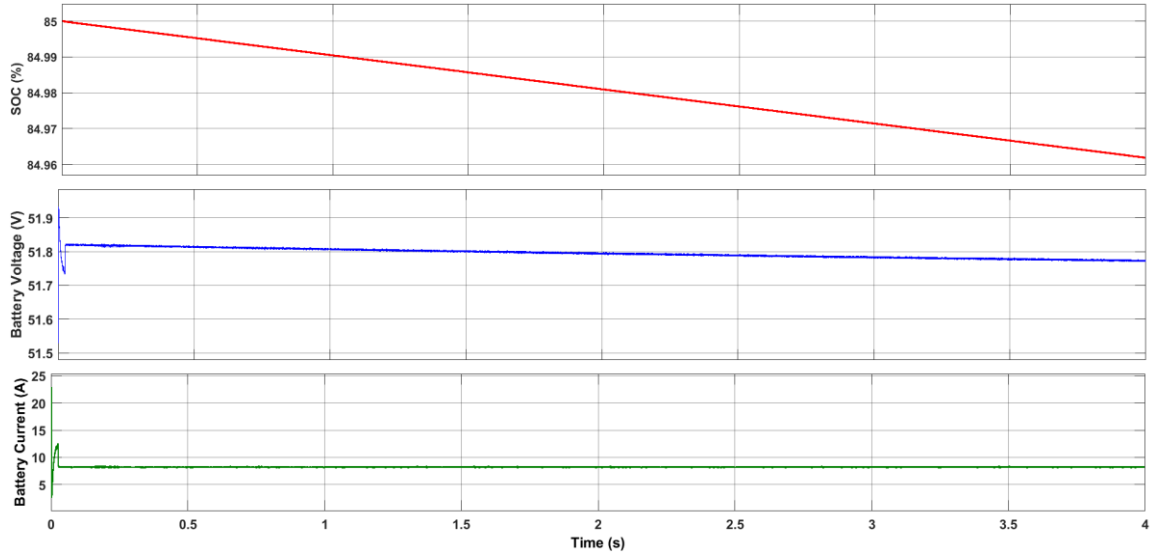


Figure 5.2. Case I simulation results: battery SOC, voltage, and current

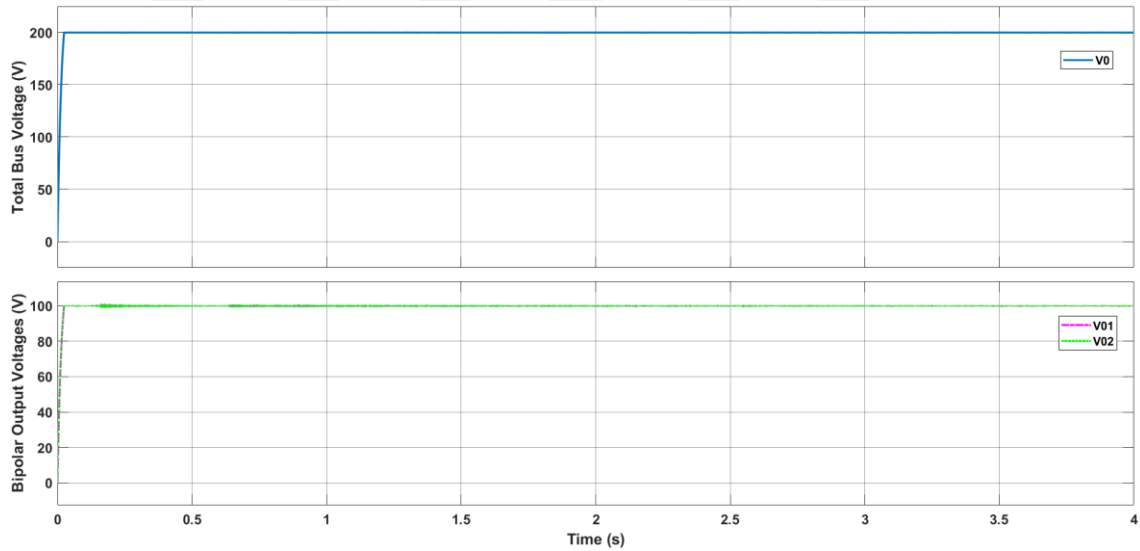


Figure 5.3. Case I simulation results: total bus and bipolar output voltages

Despite the absence of input power, the converter maintains a stable total DC-bus voltage of about 200 V and balanced bipolar outputs at ± 100 volts throughout the simulation. This outcome confirms the system's ability to ensure continuous and balanced energy delivery under complete source-deficit conditions, relying solely on the battery as the power provider.

5.1.2. Case II

In Case II, the system operates under conditions where the input source (PV) is active and capable of providing sufficient energy. In this mode, the PV array simultaneously supplies the

output load and charges the battery, representing typical daytime operation. The simulation studies include the implementation of CC/CV charging algorithms to ensure proper and safe control of the battery, while the MPPT control is employed to extract the maximum available power from the PV source. This configuration allows the converter to achieve efficient energy utilization, maintain regulated bipolar output voltages, and ensure stable battery charging behavior.

5.1.2.1. Battery Charging with CC/CV Control

Fig. 5.4 shows the CC/CV charging behavior of the battery module used in the proposed converter system. The simulations were carried out under controlled conditions corresponding to 25 °C ambient temperature and 1000 W·m⁻² solar irradiance. During the CC stage, the controller maintains a fixed charging current while the battery voltage increases steadily. When the battery voltage reaches and slightly exceeds the reference voltage, the control shifts to CV stage. At this point, the charging current gradually decreases toward zero as the battery approaches full charge. This behavior prevents overcharging and protects the battery from thermal and electrochemical stress.

The points marked as A and B correspond to distinct time instants within the charging process: point A represents the system operating under CC mode, while point B corresponds to the CV stage near full charge. The detailed zoom-in views of these points are provided in Fig. 5.5 and Fig. 5.8, respectively.

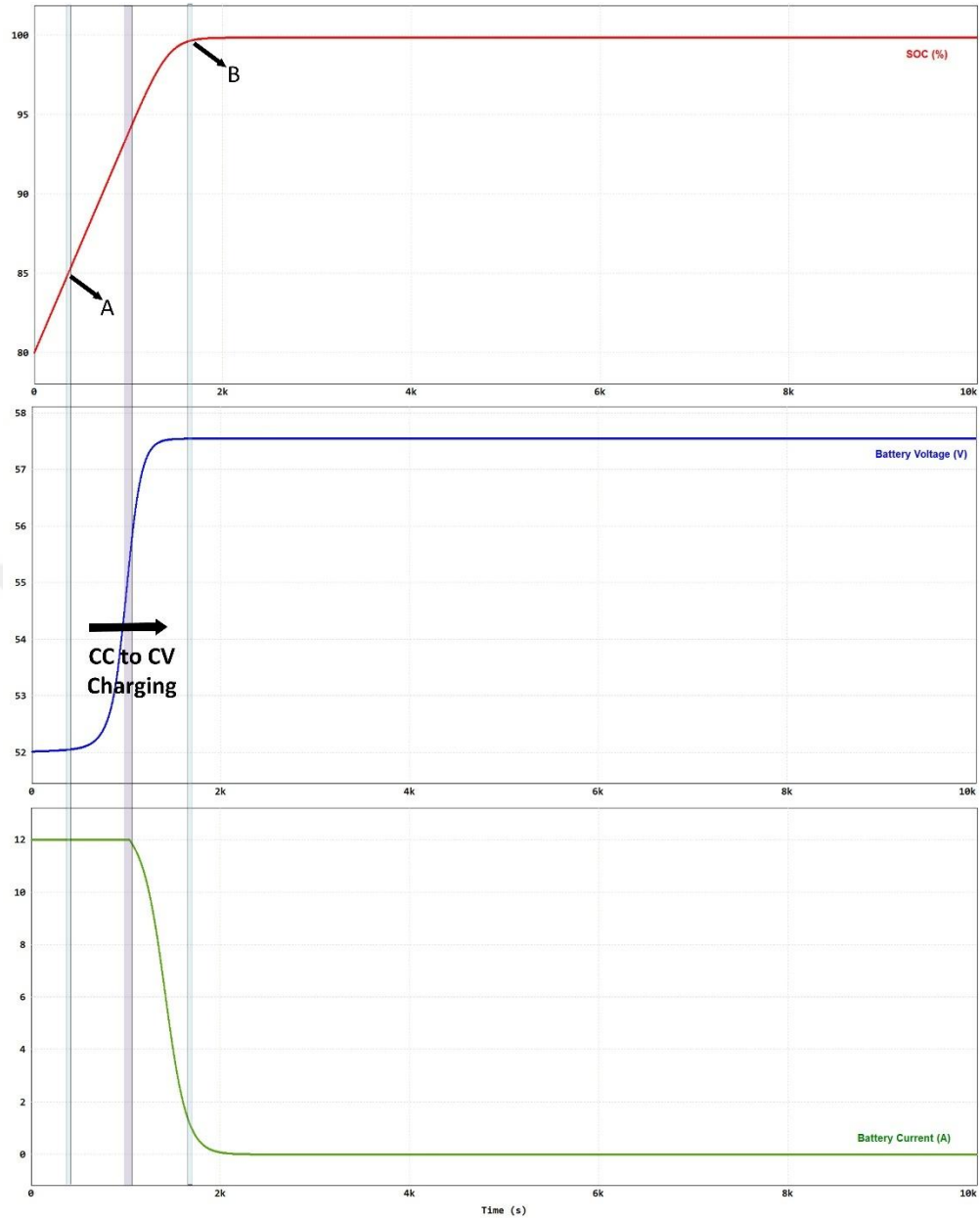


Figure 5.4. Case II simulation results: battery SOC, voltage, and current during CC–CV charging

At point A, the system operates in the constant-current (CC) charging stage, and the battery state of charge is approximately 85%, as shown in Fig. 5.5. In this phase, The controller regulates the battery charging current to its reference value of 12 A.

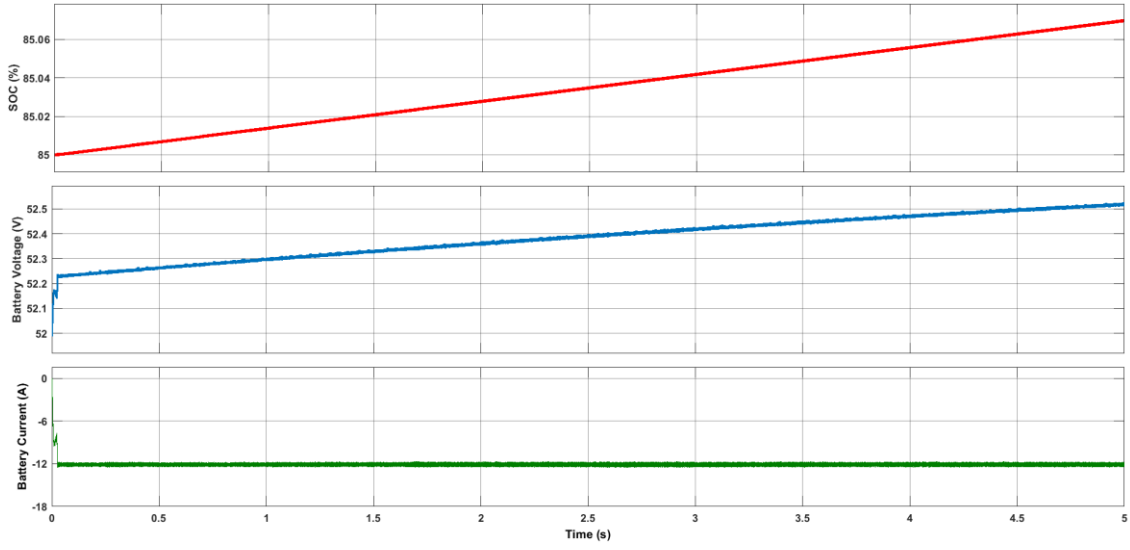


Figure 5.5. Case II CC charging simulation results: battery SOC, voltage, and current

The corresponding simulation results, presented in Fig. 5.6 and Fig 5.7. (zoomed-in view of the bipolar output voltages), illustrate the bipolar output voltages and the total DC bus voltage. These results confirm that, during CC-mode charging, the proposed converter effectively maintains balanced bipolar outputs and a stable total bus voltage.

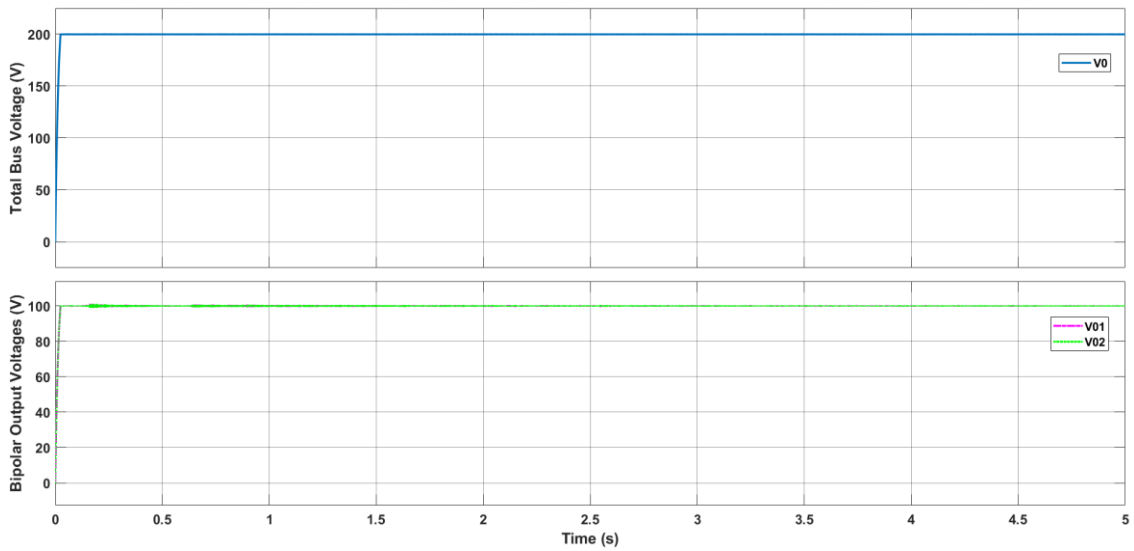


Figure 5.6. Case II CC charging simulation results: total bus and bipolar output voltages

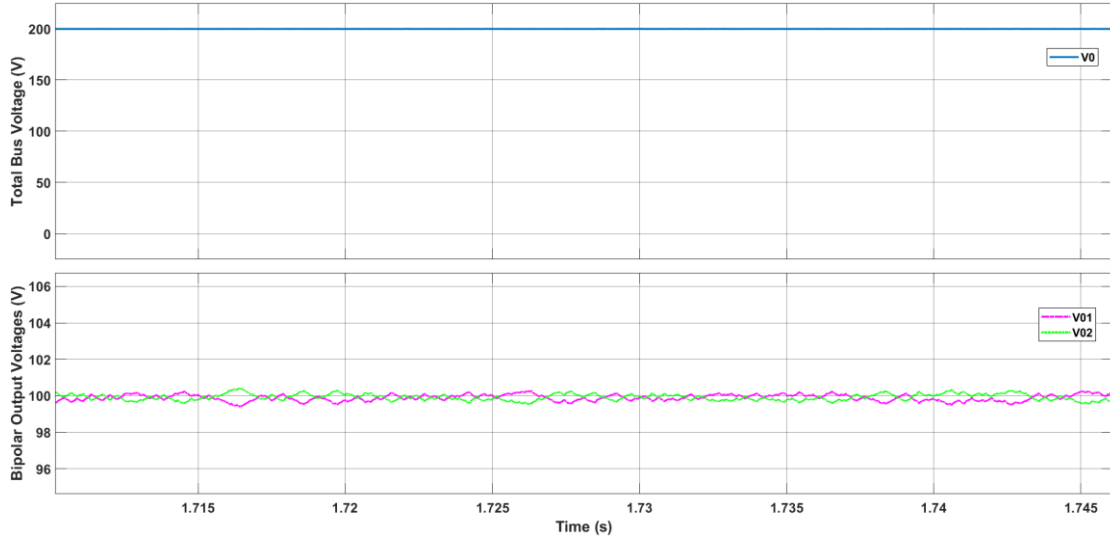


Figure 5.7. Case II CC charging simulation results: total bus and bipolar output voltages (zoom in view)

At point B of the simulation, corresponding to the CV charging stage, the battery state of charge (SOC) reaches approximately 99 % (Fig 5.8). Under this condition, the PV source operates at the same environmental parameters of 25 °C ambient temperature and 1000 W/m² irradiance. In this region, the battery voltage approaches its reference full-charge level, and the charging current decreases progressively toward zero. This controlled reduction of current ensures safe completion of the charging process without overvoltage or thermal stress on the battery cells.

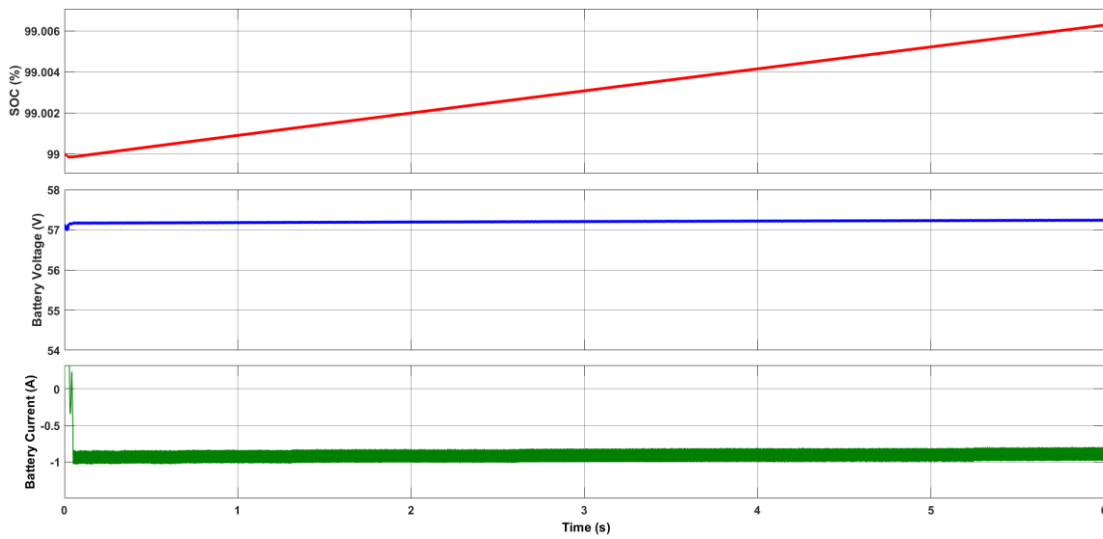


Figure 5.8. Case II CV charging simulation results: battery SOC, voltage, and current

The corresponding simulation results, presented in Fig. 5.9 and Fig. 5.10 (zoomed-in view of the bipolar output voltages), illustrate the bipolar output voltages and the total DC bus voltage during the CV mode charging stage. These results demonstrate that, even as the charging current gradually decreases toward zero and the battery approaches its full-charge voltage, the proposed converter continues to maintain balanced bipolar outputs and a stable total bus voltage, confirming reliable steady-state operation in the CV region.

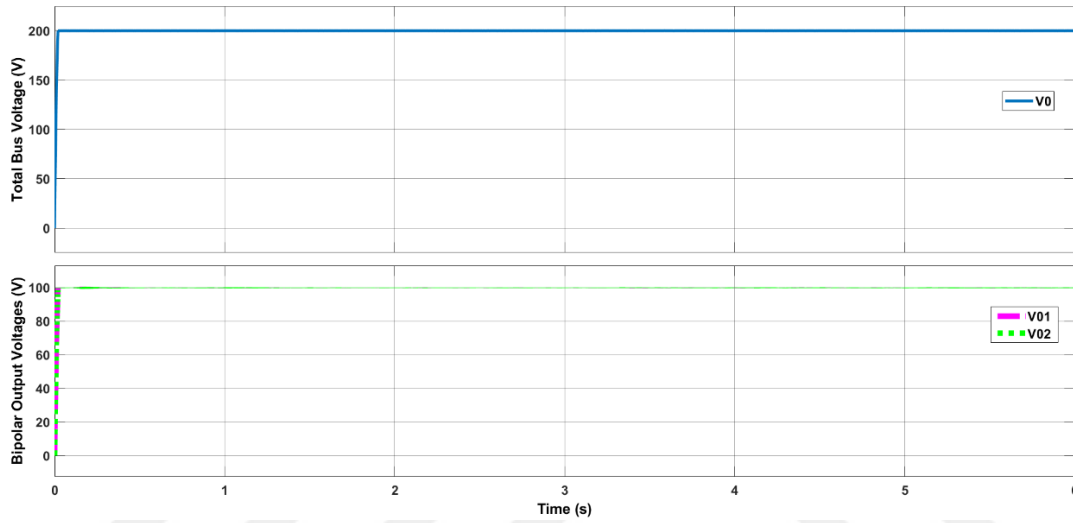


Figure 5.9. Case II CV charging simulation results: total bus and bipolar output voltages

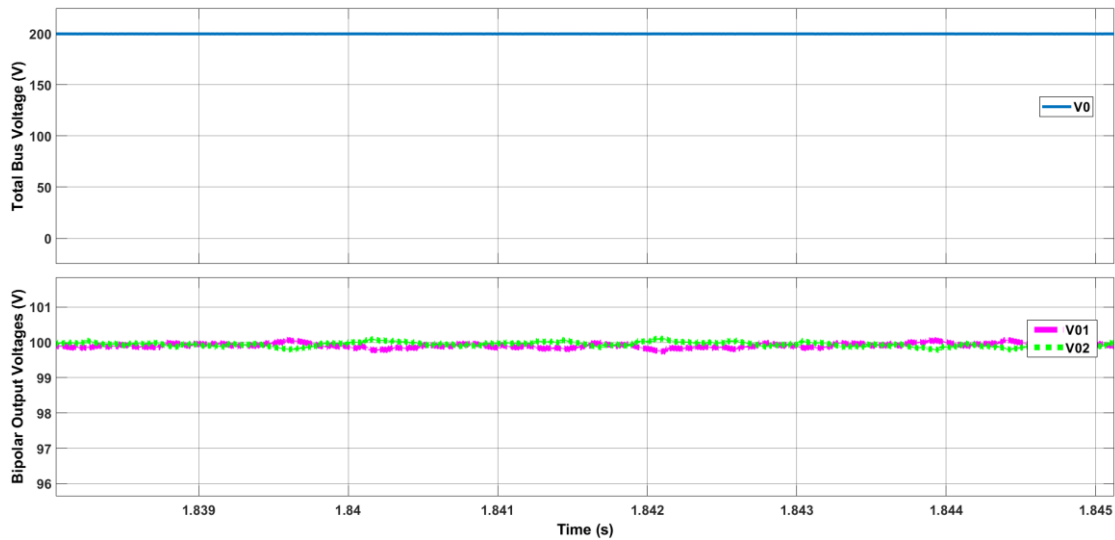


Figure 5.10. Case II CV charging simulation results: total bus and bipolar output voltages (zoom in view)

5.1.2.2. MPPT Control

In Case II, when the available irradiance is sufficient and the battery current remains below the predefined maximum limit, the system operates under MPPT control. In this condition, the PV source becomes the dominant power contributor, and the converter adjusts its duty ratio through the MPPT algorithm to extract the maximum available power corresponding to the instantaneous solar conditions.

Fig 5.11, shows the power-voltage (P-V) characteristics of the PV module used in the simulation under different solar irradiance levels. Each curve represents the PV output power variation with respect to the terminal voltage for irradiances ranging from 0.5 kW/m² to 1 kW/m² at a constant cell temperature of 25 °C.

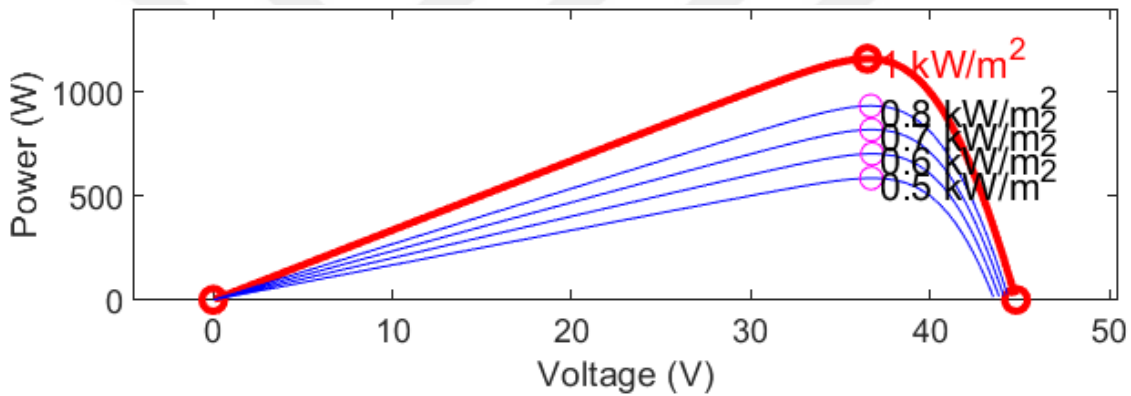


Figure 5.11. P–V Characteristics of the PV source used in the simulation study

As shown in Fig. 5.12, the P&O method was tested at a constant temperature of 25 °C under three different irradiance levels that were increased stepwise at equal time intervals. The figure presents the variation of PV power (P_{pv}) and PV voltage (V_{pv}) in response to these irradiance changes. It can be observed that although the P&O algorithm successfully tracks the maximum power point, its response becomes relatively slow during rapid irradiance transitions. Furthermore, transient oscillations appear around the steady-state operating point after each step change, indicating that the conventional hill-climbing structure of the P&O method is limited in dynamic adaptability under fast-changing environmental conditions [178], [179], [180].

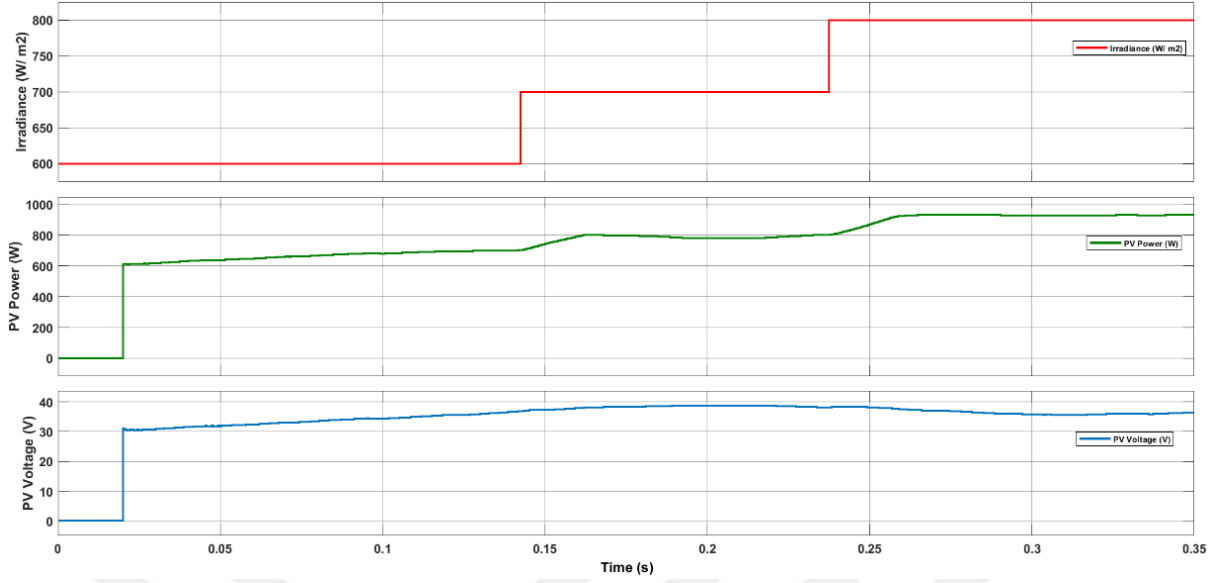


Figure 5.12. PV array voltage and power dynamics under step-change irradiance using P&O MPPT method

As an alternative to hill-climbing, ANN was developed to directly estimate the PV operating voltage at the instantaneous maximum power point. In the proposed ANN-based MPPT, four measured signals: battery current I_{bat} , battery voltage V_{bat} , PV current I_{pv} , and PV voltage V_{pv} are fed to a feed-forward network that outputs the duty ratio command d^* for switch S_0 . The ANN is trained to learn the direct mapping,

$$d^* = f_{\theta}(I_{bat}, V_{bat}, I_{pv}, V_{pv}) \quad (5.1)$$

such that d^* places the PV array at its MPP without any iterative perturbation. A dataset of 200 simulation samples was collected by sweeping irradiance and battery operating conditions (thus covering a range of V_{bat} and I_{bat}); for each condition the optimal duty yielding the maximum PV power served as the target label. Using MATLAB's ANN tool with default settings (two-layer MLP, Levenberg–Marquardt, MSE loss), the network learned the nonlinear mapping from $[I_{bat}, V_{bat}, I_{pv}, V_{pv}]$ to the MPP duty. The ANN output is applied directly to the PWM as seen in Fig. 5.13.

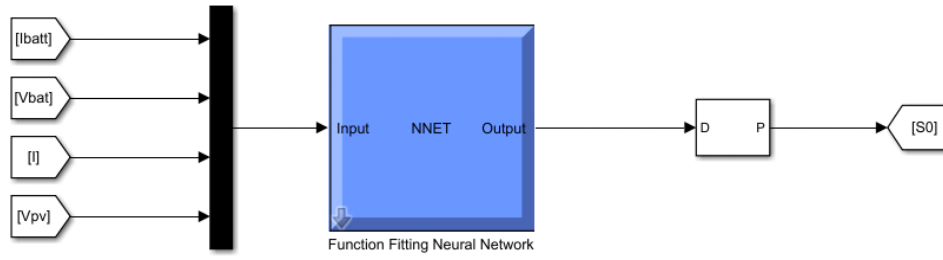


Figure 5.13. Structure of the ANN-based MPPT model for duty cycle estimation

The results of the ANN-based MPPT control, under the same battery charging condition and stepwise irradiance variations used for the P&O test, are presented in Fig. 5.14. When Fig. 5.14 (P&O-based MPPT) and Fig. 5.12 (ANN-based MPPT) are compared, it is observed that both systems respond to the stepwise irradiance changes between 600 W/m^2 , 700 W/m^2 , and 800 W/m^2 under a constant temperature of 25°C . In the P&O algorithm, the PV power follows the irradiance variation with noticeable transient oscillations and a slower convergence to the new maximum power point after each step change. In contrast, the ANN-based MPPT achieves a much faster and smoother response, maintaining the PV voltage close to the theoretical maximum power point with minimal steady-state oscillation. This simulation result has shown that the proposed ANN-based MPPT provides superior tracking speed, accuracy, and stability compared with the conventional P&O approach. However, the ANN-based MPPT method also has certain drawbacks. Its performance heavily depends on the quality and range of the training data, meaning it only works accurately within the system parameters and environmental conditions used during training. Any change in PV characteristics, converter parameters, or battery behavior may require retraining or recalibration of the model to maintain accuracy.

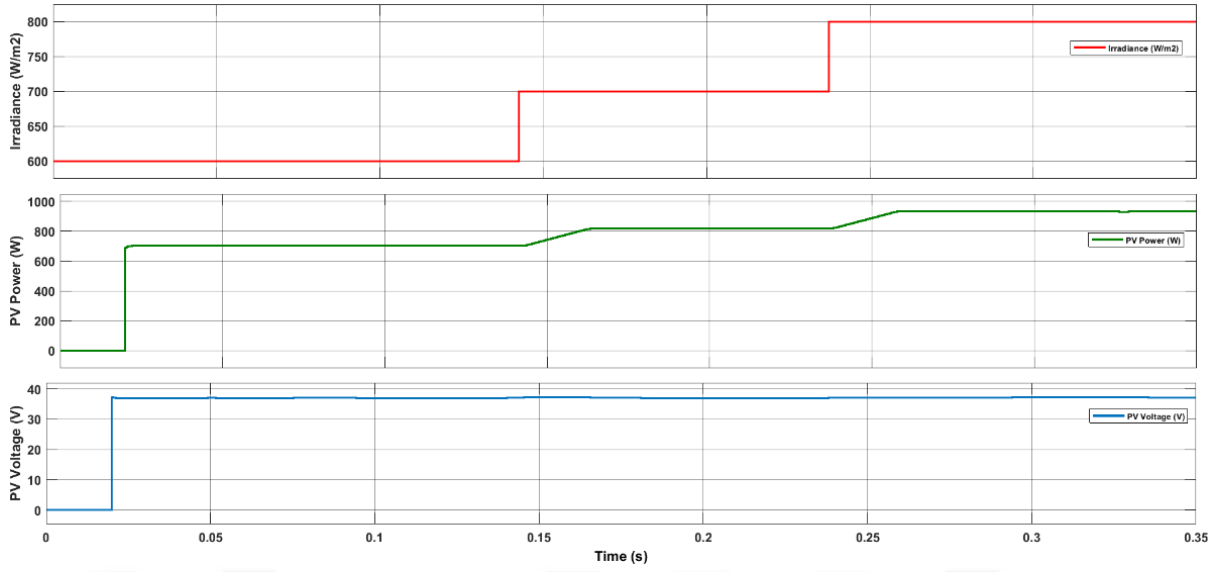


Figure 5.14. PV array voltage and power dynamics under step-change irradiance using ANN based MPPT

Figure 5.15 illustrates the dynamic behavior of the PV array under a realistic daily irradiance profile when controlled by the proposed ANN based MPPT algorithm. As the solar irradiance increases from early morning values toward its midday peak, the controller rapidly adjusts the operating voltage to the vicinity of the optimal point. Correspondingly, the PV output power exhibits a smooth and continuous rise, reaching its maximum near the irradiance apex, followed by a symmetric decline toward evening. The absence of oscillatory artifacts in both voltage and power trajectories demonstrates the strong tracking capability and stability of the ANN based MPPT under nonstationary conditions. Overall, the figure confirms that the ANN controller effectively adapts to gradual irradiance variations, sustains high power extraction efficiency over the full daylight cycle, and ensures robust voltage regulation.

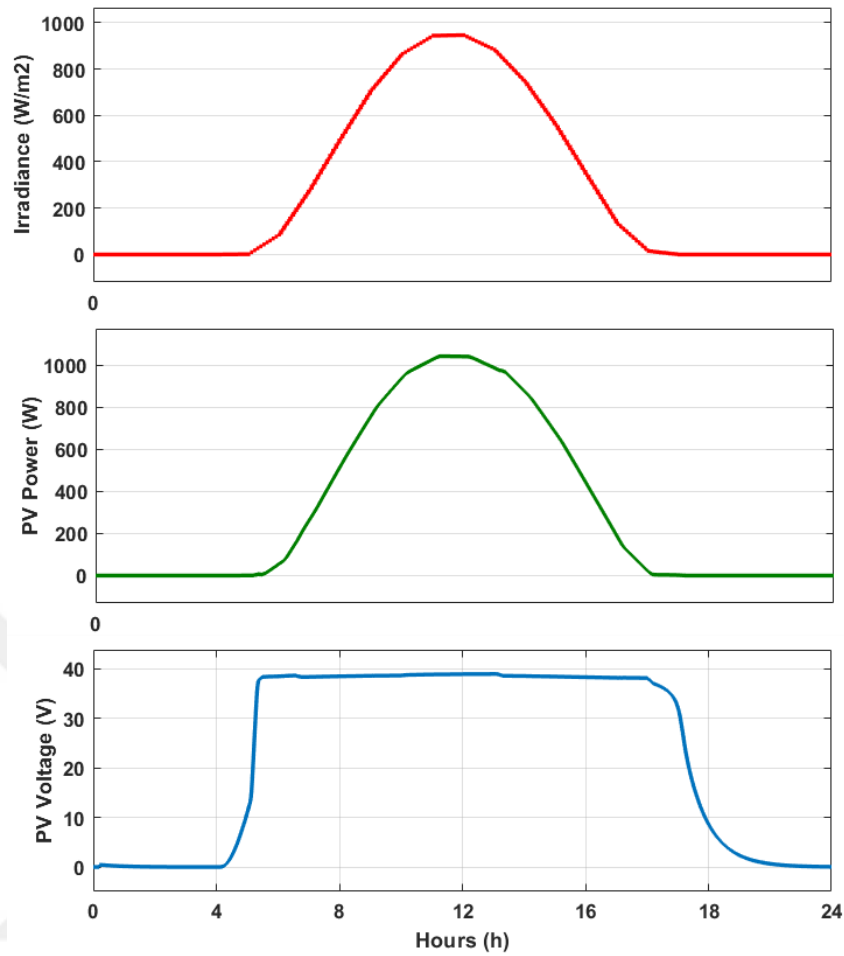


Figure 5.15. PV Array dynamic voltage and power characteristics under daily irradiance profile using ANN based MPPT

5.1.3. Case III

At the beginning of Case III, the solar irradiance is fixed at 200 W/m^2 , yielding a PV terminal voltage of about 33.3 V and a PV power of approximately 220 W . Since this level is insufficient to meet the load demand, the converter operates in battery assisted discharge mode. As shown in the Fig 5.15, under a constant irradiance of 200 W/m^2 , the battery discharges with a current of approximately 4 A , resulting in a gradual decrease in both battery voltage and state of charge. This behavior indicates that the battery compensates for the power deficit caused by insufficient PV generation.

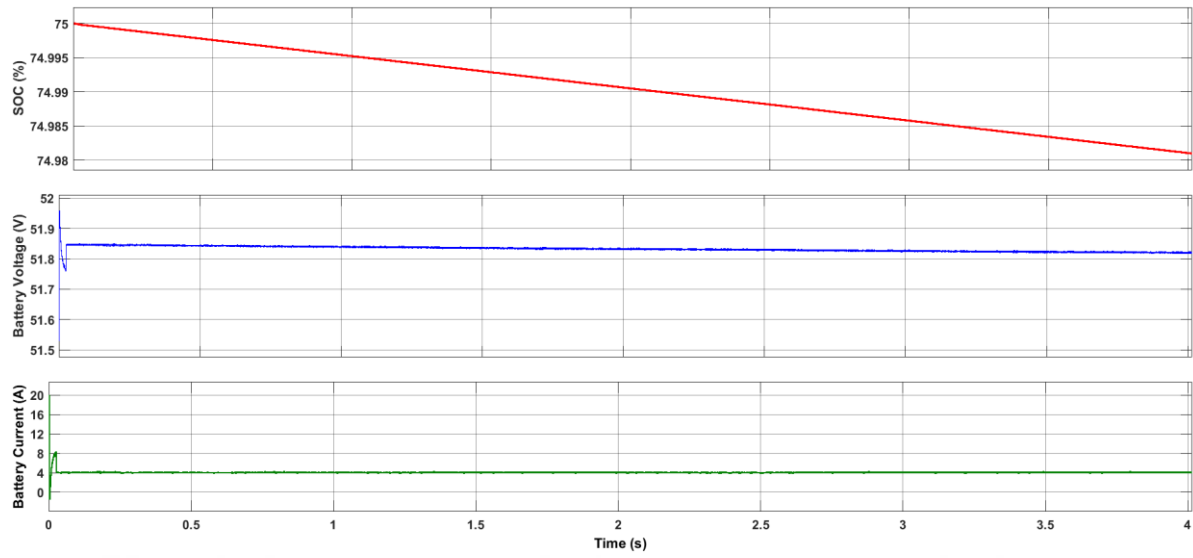


Figure 5.16. Case III Simulation Results: Battery SOC, Voltage, and Current

Consequently, as illustrated in Fig. 5.16, the bipolar output voltages remain well regulated, and the total DC bus voltage is held around its reference value of 100 V, confirming the system's ability to sustain balanced operation under low-irradiance, source-deficit conditions.

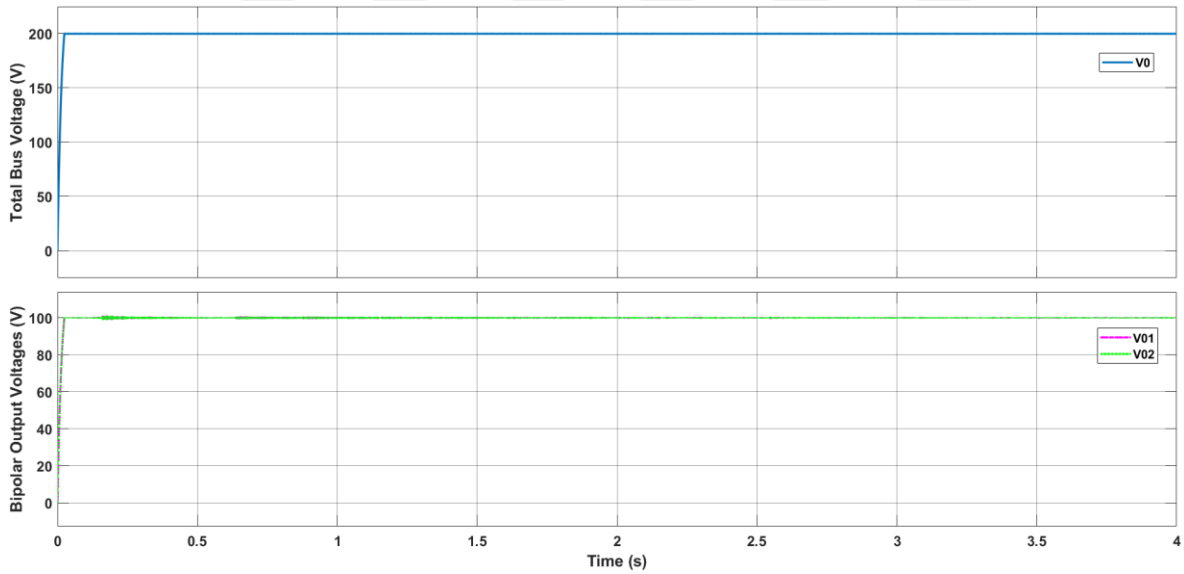


Figure 5.17. Case III Simulation Results: Total Bus and Bipolar Output Voltages

5.1.4. Battery-Coupled Bipolar DC–DC Converter Symmetrical and Asymmetrical Load Simulation Results

5.1.4.1. *Bipolar Voltage Symmetry under Dynamic Load Conditions*

During this transient-load evaluation, each bipolar output of the converter was initially terminated with a $50\ \Omega$ purely resistive load, with output setpoints $V_{o1} = +100\text{ V}$ and $V_{o2} = -100\text{ V}$ (total bus $V_{\Sigma} = 200\text{ V}$). At $t = 0.5\text{ s}$, a deterministic step disturbance was introduced by paralleling an additional $50\ \Omega$ element on each output branch via a MATLAB breaker block. This action reduced the per-polarity equivalent resistance from $50\ \Omega$ to $25\ \Omega$, thereby instantaneously doubling the output power demand while keeping load characteristics strictly resistive and balanced across the two polarities.

The measured waveforms in Fig. 5.17 (with magnified view in Fig. 5.18) show that the total bus voltage remains centered at 200 V , exhibiting only a brief, well-damped undershoot immediately following the step, after which it recovers promptly to its pre-disturbance setpoint. Concomitantly, the individual outputs V_{o1} and V_{o2} depart momentarily from their nominal magnitudes during the transient yet resettle to $\pm 100\text{ V}$ without sustained deviation, maintaining mirror symmetry with respect to the common midpoint. Importantly, this symmetric restoration occurs without any auxiliary balancing circuitry: the feedback loop directly regulates only the total bus, but the converter exhibits an intrinsic self-balancing effect that equalizes the two polarities after the disturbance.

This behavior is consistent with the proposed architecture and modulation strategy. Because both polarities share the same transformer and are energized by a single phase-shift control variable, the transient increase in power demand appears at the controller as a total-bus voltage error, which in turn modulates the transferred energy in a manner that raises (or lowers) both polarities simultaneously. Collectively, these results demonstrate that the converter preserves tight total-bus regulation and inherent bipolar voltage symmetry under sudden, two-fold load excursions, validating the robustness of the control-topology combination for dynamic operating conditions.

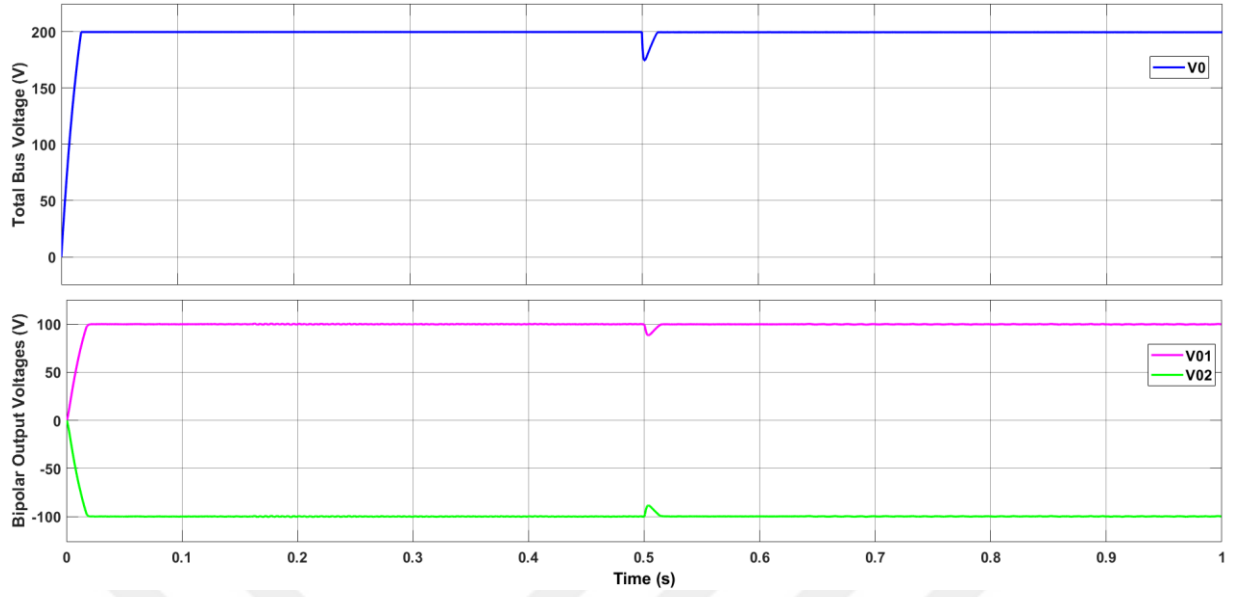


Figure 5.18. Total bus and bipolar output voltages under symmetrical load variation

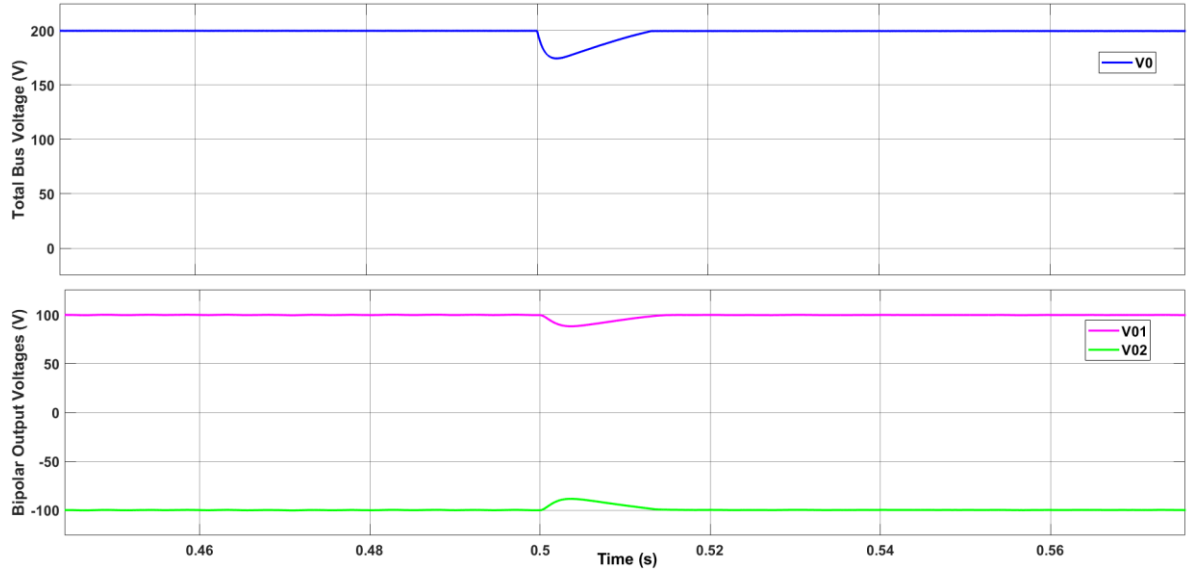


Figure 5.19. Total bus and bipolar output voltages under symmetrical load variation (zoom in view)

The corresponding gate sequences in Fig. 5.19 and Fig. 5.20 substantiate the commanded control action. Immediately after the load step, the controller increases the phase shift δ between the primary and secondary bridges (Fig. 5.20), and drives the total bus back to its reference $V_{\Sigma}^* = 200$ V. Consistent with this action, the bus voltage exhibits a brief, well-damped deviation before returning to set-point, while the bipolar outputs maintain mirror symmetry

about the midpoint during resettling. The observed phase shift adjustment and the prompt recovery collectively indicate a robust closed loop response under dynamic loading and confirm that bipolar voltage symmetry is preserved without auxiliary balancing circuitry.

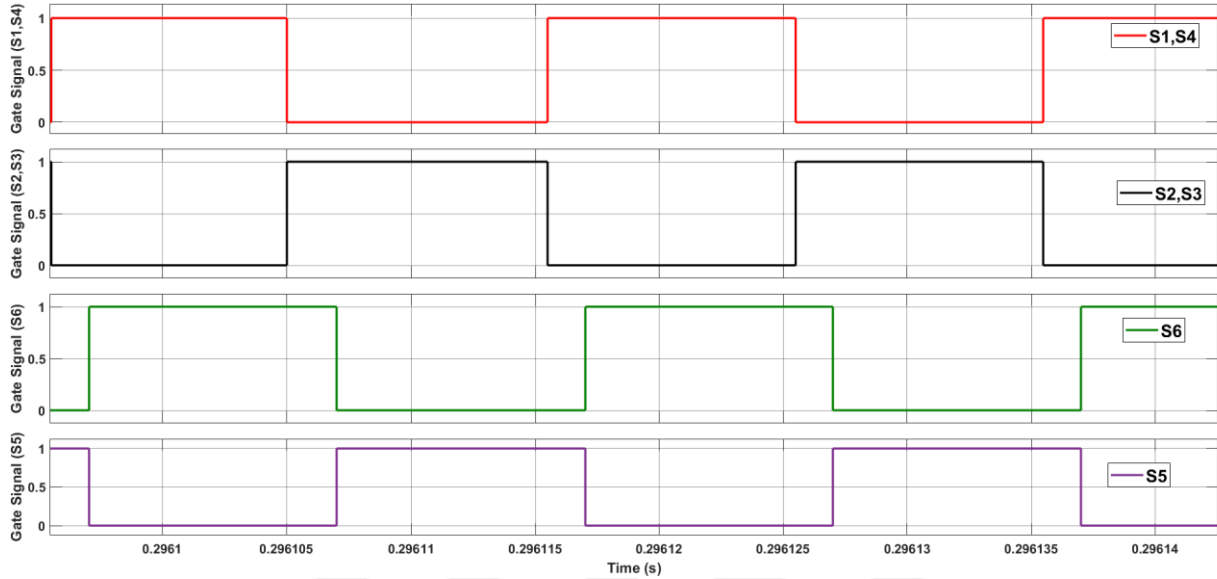


Figure 5.20. Gate drive signals of primary and secondary switches before load variation

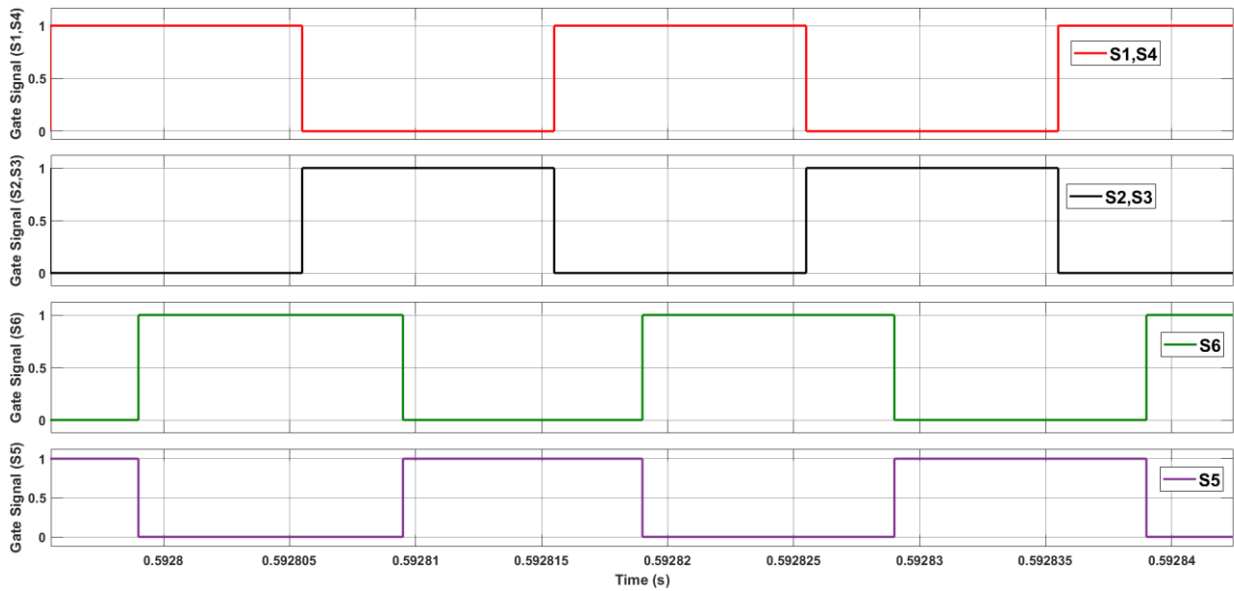


Figure 5.21. Gate drive signals of primary and secondary switches after load variation

5.1.4.2. Bipolar Voltage Regulation under Asymmetric Loads

In this simulation test, an asymmetric load condition was considered in which the bipolar outputs were terminated with $R_{01} = 50 \Omega$ and $R_{02} = 80 \Omega$. Under these conditions, the 50Ω branch has 37.5 % lower resistance and therefore draws approximately 60 % more current and power compared with the 80Ω branch. Despite this substantial asymmetry, the bipolar output voltages remain highly symmetrical as seen in Fig 5.21. The voltage difference between the two outputs is below 1 V.

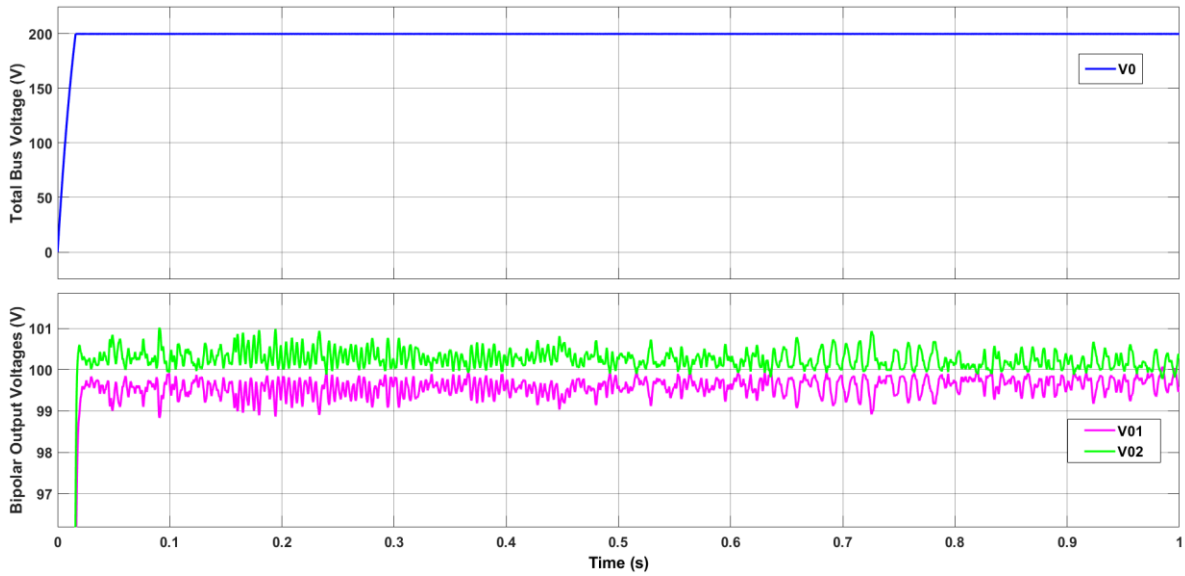


Figure 5.22. Total bus and bipolar output voltages under asymmetrical load condition

In this topology, the controller regulates only the total DC bus voltage. It does not impose separate voltage references for the positive and negative outputs. On the secondary side, a symmetric arrangement of rectifier diodes and output capacitors shares a common transformer winding. As a result, the same secondary voltage waveform is applied to both outputs in every switching period. When one output voltage becomes slightly lower, the instantaneous voltage difference between the transformer secondary and that output capacitor increases during conduction. This produces a higher average rectifier current, and therefore more charge is transferred to this capacitor in each cycle. The higher voltage output receives comparatively less charge. Consequently, the total DC bus voltage remains regulated, while the lower output is driven upward and the higher output is slightly reduced. In this way, the two bipolar outputs remain nearly balanced even under unequal loading conditions. Consequently, even under

pronounced load asymmetry, the system maintains both bipolar voltages close to ± 100 V with less than 1 V deviation and about 1 % ripple. Overall, these findings confirm the robustness of the proposed converter and highlight its inherent electromagnetic and control related self balancing capability.



6. EXPERIMENTAL STUDIES

6.1. Implementation of the Battery-Coupled Bipolar DC–DC Converter Prototype

The prototype was realized in two parts: an electronic control stage and a power stage (boost converter, high-frequency transformer, bipolar isolated converter, and Li-ion battery interface). Both stages were integrated on a custom PCB. The complete prototype was experimentally verified in the laboratory to confirm the proper functionality of the proposed converter.

6.1.1. Control System Design of the Proposed System

This section provides an overview of the design, implementation, and integration of the electronic control system that governs all measurement, protection, and switching functions of the proposed converter. It explains the overall architecture, the operation of the DSP-based control unit, the design of measurement and protection hardware, the development of embedded software for PWM and phase-shift control, and the PCB layout considerations ensuring proper isolation, grounding, and signal integrity throughout the system.

6.1.1.1. *Design Considerations of Electronic Control System*

The electronic control system developed for the proposed battery-coupled bipolar DC–DC Converter performs all measurement, control, and protection tasks necessary for real-time operation. The main components of the control system include the power supply module, the DSP-based microcontroller unit (TMS320F28335), measurement and signal conditioning circuits, digital and analog protection circuits, and gate-driver units. The overall block diagram of the developed control system is shown in Fig. 6.1.

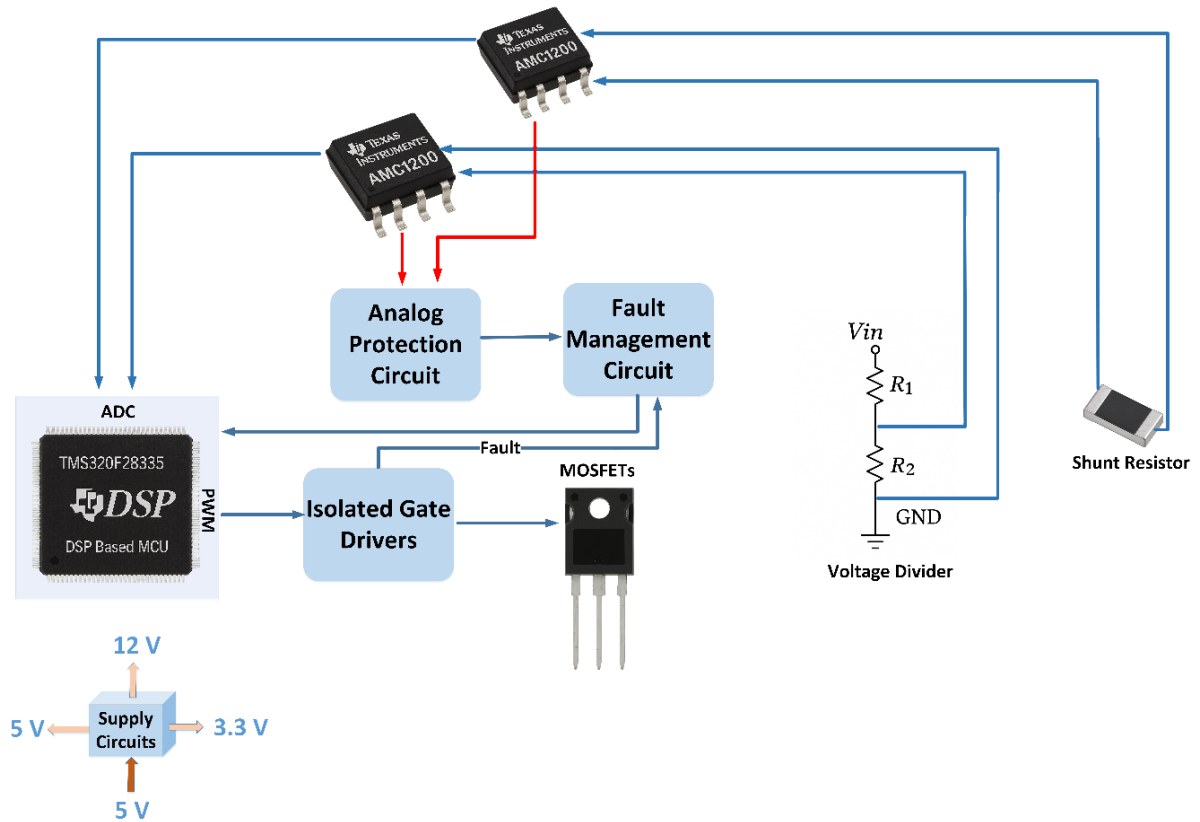


Figure 6.1. Block diagram of the converter control system

A single regulated +5 V input supply is used to energize the control board, while isolated DC–DC converters (5 V–5 V, 5 V–12 V, and 5 V–3.3 V) provide the required voltage rails for each subsystem. This structure ensures galvanic isolation between the digital control core, analog measurement circuits, and gate driver stages, thereby enhancing noise immunity and operational safety.

At the heart of the system lies the TMS320F28335 DSP, which executes the control algorithms for both the boost converter and the phase-shifted full-bridge stage. The DSP processes analog feedbacks through its on-chip 12-bit ADC module. Based on these measurements, it computes the appropriate duty ratio and phase-shift angle. The control signals are generated via the ePWM modules and transmitted to the gate driver circuits.

Accurate and isolated measurements are achieved using Texas Instruments AMC1200 isolation amplifiers. Each AMC1200 provides precise differential measurement and 4 kV RMS galvanic

isolation between the high-side sensing nodes and the low-voltage control domain. The battery current is sensed across a precision shunt resistor, and its differential voltage is fed to the AMC1200 for amplification and isolation before conversion by the DSP's ADC. Similarly, the total bipolar output voltage is divided, isolated, and conditioned through another AMC1200 channel to provide noise-free voltage feedback. The use of the AMC1200 devices allows the control board to directly interface with high-potential nodes without compromising measurement linearity or common-mode rejection.

Analog protection circuitry continuously monitors these measured variables. Over-voltage and over-current thresholds are implemented in hardware to generate fault signals that are latched by the fault-management unit and communicated to the DSP's Trip-Zone (TZ) pins for immediate PWM shutdown.

6.1.1.2. *Digital Signal Processor Based Microcontroller Unit*

The developed battery-coupled bipolar DC–DC Converter structure includes seven controllable switches: one boost switch at the input (S_0), four devices forming the primary H-bridge (S_1 – S_4), and two devices on the secondary bridge (S_5 – S_6). The microcontroller unit (MCU) must execute the real-time control algorithm at the chosen switching frequency, acquire/convert analog signals deterministically, and generate complementary gate signals with adjustable dead-time; it should also provide fast fault handling (trip/disable), sufficient on-chip PWM/ADC resources, real-time debugging via an emulator, and adequate memory for code, parameters, and logs. A Texas Instruments TMS320F28335 DSP is employed as the MCU due to its integrated ePWM generators, multi-channel 12-bit ADC, hardware protection, and real-time emulation support. Using each ePWM's A/B complementary outputs (with dead-band), the required PWM modules are minimized: primary legs share complementary pairs within a single ePWM, diagonal devices are aligned across two ePWMs, and the secondary bridge receives a programmable phase displacement. Thus, four ePWM modules drive seven switches, as summarized below.

Table 6.1. ePWM switch assignment in the prototype

Function	Switches	ePWM module / outputs
Boost stage	S ₀	EPWM1A
Primary bridge	S ₁ , S ₂	EPWM2A → S ₁ EPWM2B → S ₂
Primary bridge	S ₃ , S ₄	EPWM3A → S ₄ EPWM3B → S ₃
Secondary bridge (phase-shifted)	S ₅ , S ₆	EPWM4A → S ₆ EPWM4B → S ₅

6.1.1.3. Design of Measurement, Protection and I/O Circuits

The control algorithm employs five feedback variables: battery current, battery voltage, input current, input voltage, and the total bipolar output voltage. All of these feedback signals are measured through precision isolation amplifiers, which ensure high accuracy and galvanic isolation between high-potential sensing points and the low-voltage control domain. Since the AMC1200 input range is limited to ± 250 mV, current is sensed through a precision shunt resistor and voltage is scaled using resistor dividers to generate differential signals within this limit. The amplifier internally provides an $8\times$ output gain, amplifying the sensed signal before conditioning. This differential output is then converted into a single-ended signal by an operational amplifier stage, making it compatible with the DSP's ADC input. To preserve battery current polarity information, a mid-scale DC bias is applied at the conditioning stage. This offset centers the bipolar current waveform at 1.5 V while preserving directional sensitivity, enabling the controller to infer charging or discharging from deviations around the midpoint.

The fault logic of the electronic control system is coordinated by a dedicated fault-management stage. Its functions are threefold: (i) latching the fault signals generated by the analog comparator supervision of the measured currents and voltages, together with device/driver diagnostic faults from the power stage; (ii) resetting the latched faults when the abnormal

condition is cleared and a DSP reset command is issued; and (iii) generating a trip signal to stop the converter. All individual fault lines are first level-translated in the digital I/O interface, then applied to the fault-management network where they are logically combined and latched. The resulting consolidated protection signal is routed to the DSP Trip-Zone input to enforce an immediate, hardware-based shutdown of the PWM outputs.

In the developed control board, digital input and output interfacing between the DSP and peripheral circuits is achieved through bus transceiver units, which perform safe bidirectional logic-level translation. These circuits ensure reliable communication between the 3.3 V DSP domain and the 5 V peripheral domain, preventing logic mismatch or overvoltage stress on the DSP pins. In the implemented design, this function is realized by level-shifting transceivers that elevate the DSP's PWM outputs from 3.3 V to 5 V for the gate drivers, while simultaneously converting 5 V protection and fault feedback signals down to 3.3 V for DSP monitoring.

6.1.1.4. *Development of The Embedded Software*

In the digital control structure, the PWM generation in the ePWM modules of the TMS320F28335 DSP can be configured in three distinct counting modes: up-count, down-count, and up-down. The switching frequency of each mode is determined by the relationship between the time-base clock (f_{TBCLK}) and the period register ($TBPRD$). PWM generation can be configured in three counting modes with the following definitive frequency relations:

$$f_{sw} = \begin{cases} \frac{f_{TBCLK}}{TBPRD}, \text{ up} \\ \frac{f_{TBCLK}}{TBPRD}, \text{ down} \\ \frac{f_{TBCLK}}{2 TBPRD}, \text{ up-down} \end{cases} \Rightarrow TBPRD = \begin{cases} \frac{f_{TBCLK}}{f_{sw}}, \text{ up mode and down mode} \\ \frac{f_{TBCLK}}{2 f_{sw}}, \text{ up-down mode} \end{cases}$$

In this prototype, up (edge-aligned) mode with AQ: ZRO = SET, CAU = CLEAR is used (i.e., EPWMxA is HIGH at CTR = 0 and goes LOW at CTR = CMPA). The time-base clock is $f_{TBCLK} = 150$ MHz because the TMS320F28335 is operated at that rate in this application. For the TMS320F28335, the time base clock is $f_{TBCLK} = 150$ MHz (derived from the system clock

$\text{SYSCLKOUT} = 150 \text{ MHz}$ with $\text{CLKDIV} = \text{HSPCLKDIV} = 1$). Accordingly, for a target switching frequency f_{sw} , the ePWM period register is,

$$TBPRD = \frac{f_{TBCLK}}{f_{sw}} \quad (6.1)$$

In this configuration, CMPA is the register that defines the PWM duty ratio and, for up (edge-aligned) mode, it is given by,

$$\text{CMPA} = D (TBPRD) \quad (6.2)$$

Similarly, TBPHS is the register that applies the phase shift between synchronized ePWM modules and, in this mode, it is set as,

$$\text{TBPHS} = \frac{\varphi}{360^\circ} (TBPRD) \quad (6.3)$$

These relations are written to the corresponding ePWM registers to realize the commanded duty cycle and phase shift. During operation, the control loop periodically acquires voltage and current feedbacks through the ADC, processes these measurements, and generates corresponding error signals. Based on these, the DSP calculates the required duty ratio and phase-shift values and updates the respective PWM registers ; EPWM1 CMPA for duty control and EPWM4 TBPHS for phase-shift control.

6.1.1.5. PCB Design of Electronic Control System

The PCB was designed in Altium Designer. The process began with schematic capture using verified symbols/footprints and electrical rule checks. Schematics were then transferred to the PCB editor, the board stack-up was defined, and constraint-driven placement/routing was executed. After sign-off, manufacturing data (Gerber/ODB++, drill files, pick-and-place, and BOM) were generated. Subsequently, the PCB was fabricated in multiple copies; each bare

board underwent visual inspection and short-circuit/continuity. Only after these preliminary checks were successfully completed were the components assembled and the boards soldered. Photographs of the bare PCB and the assembled, soldered PCB are provided in Fig. 6.2 and Fig. 6.3, respectively.

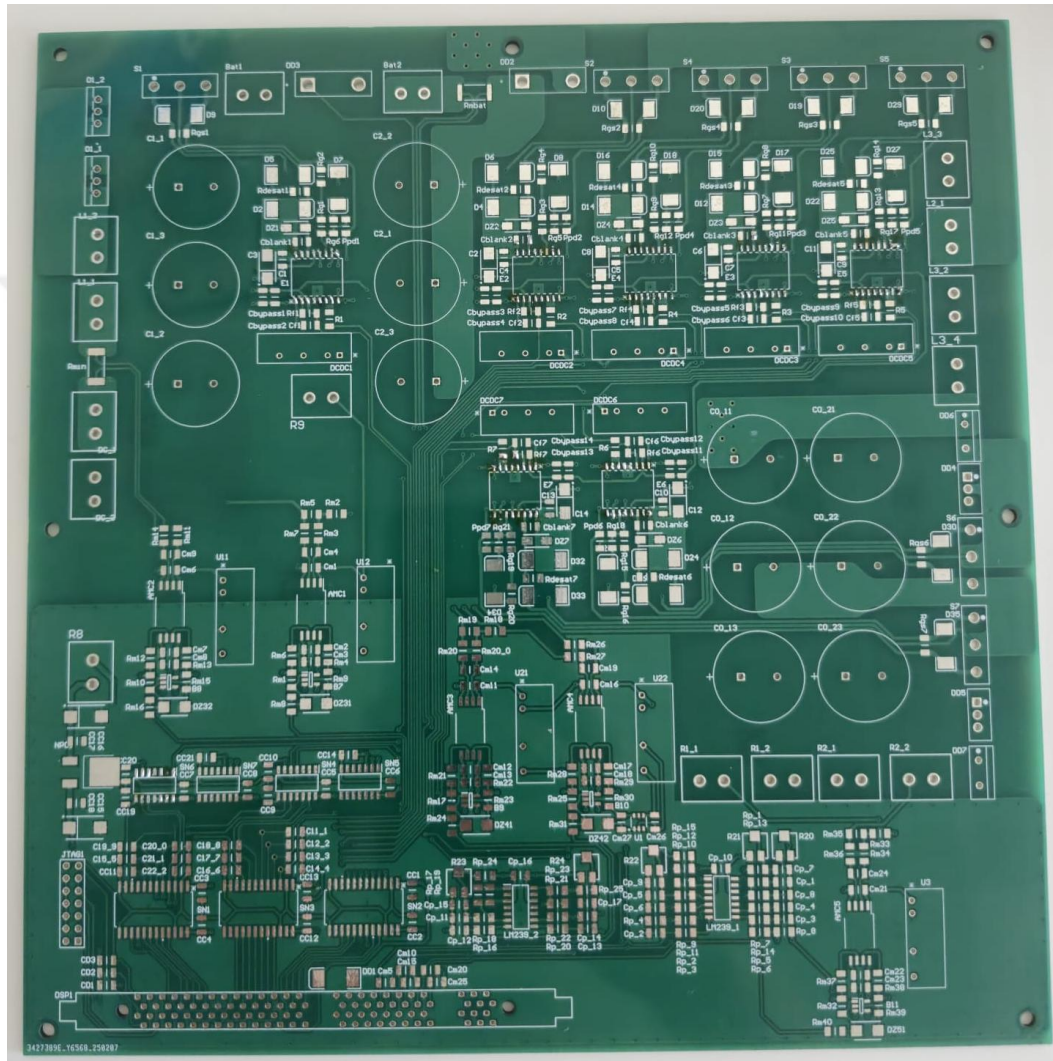
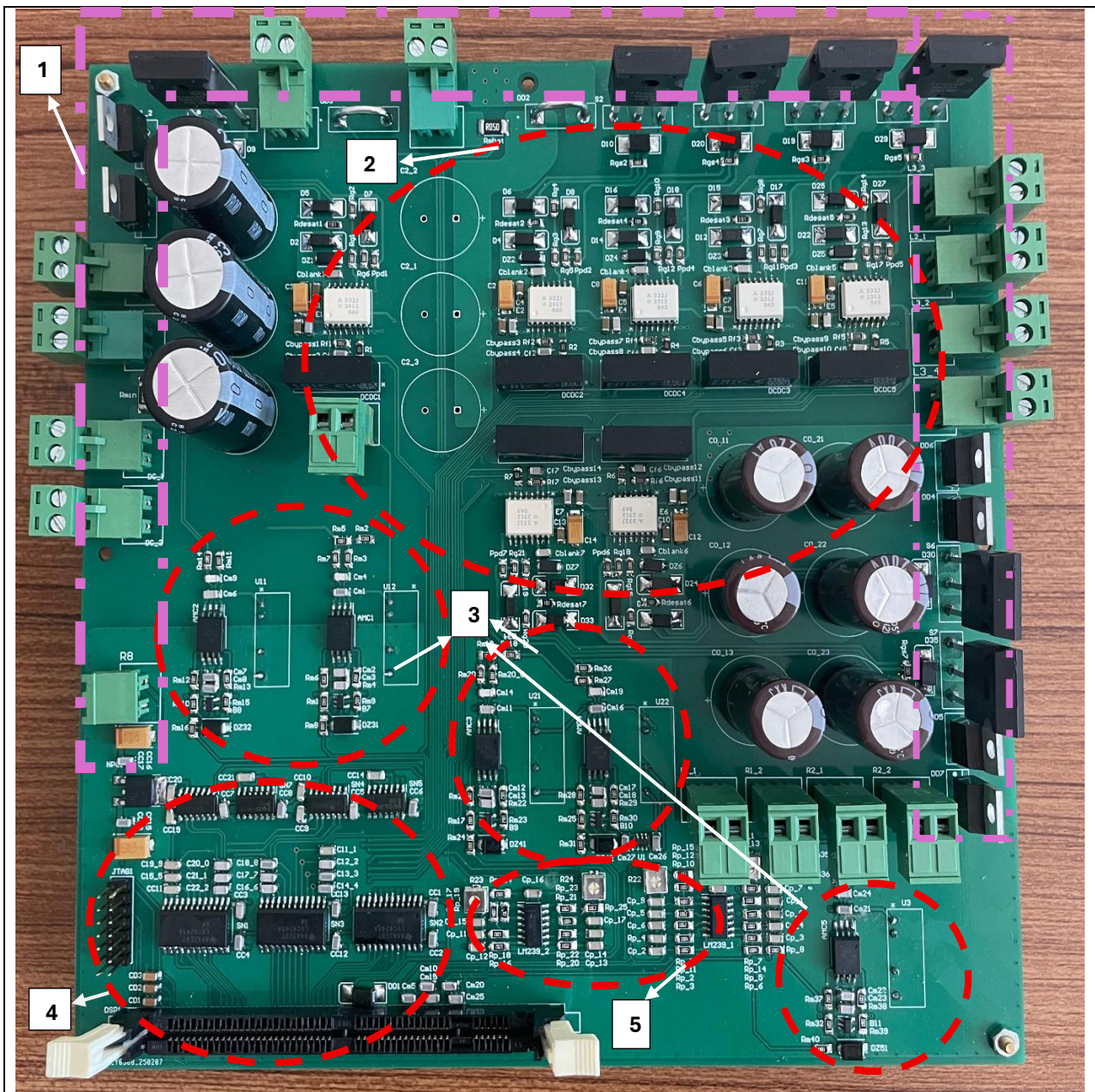


Figure 6.2. Fabricated PCB of the proposed converter before soldering process



1- Power Circuit	2- Gate Driver Circuits	3-Measurment Circuits
4-DSP and DSP voltage-level transition Circuits	5-Protecton circuits	

Figure 6.3. Assembled PCB

6.1.2. Hardware Implementation of the Proposed System

The prototype of the proposed battery-coupled bipolar DC–DC converter was developed to experimentally validate the designed control and power stages. All components, including the boost inductor, high-frequency transformer, gate-driver circuits, Li-ion battery pack, and bipolar output filters, were integrated on a custom-designed PCB.

6.1.2.1. Selection of Power Switches

In the design of the power circuit, Onsemi FCHD040N65S3 super-junction MOSFETs have been employed as the main switching elements for all seven active devices in the converter. Each MOSFET is rated at 650 V drain-source voltage and 65 A continuous drain current (at 25 °C). The MOSFETs are driven by Broadcom ACPL-332J-500E optically isolated gate-driver modules. The drivers provide desaturation detection, under-voltage lockout, active Miller clamp, fault feedback, and soft turn-off. The input/control domain is supplied from 5 V DC, while the driver's output power stage is provided with an isolated 12 V supply to energize the gate-drive path. Gate commands are generated by the DSP-based controller (TMS320F28335) according to the converter's operating mode.

6.1.2.2. Selection of High Frequency Transformer

The High Frequency Transformer (HFT) was designed following the procedure and constraints established in Chapter 3, and then custom-manufactured by a domestic supplier in accordance with the finalized electrical and mechanical parameters. The specifications of the high-frequency transformer used in this study, which is shown in Fig. 6.4, are presented in Table 6.2.

Table 6.2. Specifications of the High Frequency Transformer

Parameter	Value
Power	Power 1 kW
Primary Voltage	50 V
Secondary Voltages	100 V
Leakage inductance	2.47 μ H
Switching Frequency	50 kHz
Turns Ratio (S/P)	2
Material	PL-7
Geometry	E-core (E65/32/27)



Figure 6.4. High frequency transformer used in the prototype

6.1.2.3. Selection of Boost Inductor and DC Link Capacitor

The inductor used in the prototype was selected according to the design procedure and parameter set established in Chapter 3 for the boost stage. The required inductance, saturation current, and loss targets were derived from the worst-case operating point. The inductors were then manufactured by a domestic supplier to these specifications. Their electrical and mechanical characteristics are summarized in Table 6.3, and the manufactured parts are shown in Fig. 6.5.

Table 6.3. Specifications of the Boost Converter Inductor

Parameter	Value / Notes
Inductance	1 mH
Voltage	40 V
Current	35 A



Figure 6.5. Boost converter inductor used in the prototype

Since the boost converter operates in direct interaction with the battery, the DC-link capacitor was chosen slightly larger to ensure stable voltage and smooth power transfer during charge–discharge transitions. Accordingly, a single 2200 μF , 63 V aluminum electrolytic capacitor was connected at the DC-link to provide sufficient energy buffering and minimize voltage fluctuations during dynamic operating conditions.

Per-rail bipolar output capacitors were sized to confine the steady-state peak-to-peak ripple to 1–2% of the nominal rail voltage. Rather than a single device, three 220 μF 200 V aluminum electrolytic capacitors were connected in parallel (≈ 660 μF effective) to reduce the effective ESR/ESL and increase the allowable ripple-current rating, thereby lowering ripple voltage and improving high-frequency decoupling and transient response. To preserve voltage symmetry, matched capacitor networks (same type/value/count) were used on both bipolar outputs.

6.1.2.4. *Selection of Battery*

In the experimental prototype, the converter was evaluated at a reduced power level, with the total bipolar DC output limited to 50 V. Accordingly, the battery pack was constructed by connecting three 3.7 V lithium-ion cells in series, providing a nominal voltage of approximately 11.1 V. This configuration ensured both safe operation and compatibility with the designed converter topology while maintaining adequate energy storage for bidirectional power transfer during charge and discharge tests.

Lithium-ion chemistry was selected due to its high energy density, stable voltage profile, and low internal resistance, which enable efficient current delivery and precise voltage control under dynamic converter operation. Additionally, lithium-ion cells exhibit good cycle life, fast transient response, and low self-discharge, making them suitable for laboratory-scale dc–dc converter validation. The chosen configuration also minimized thermal risk and simplified protection requirements compared to higher-voltage arrangements.

6.1.2.5. *The Prototype of the Proposed System*

Fig. 6.6 presents the experimental implementation of the proposed battery-coupled bipolar DC-DC converter prototype. The upper photograph shows the complete laboratory setup, including

the DC power supplies, DSP (TMS320F28335) controller board, Li-ion battery pack, high-frequency transformer, and load selection unit. The lower image highlights the assembled converter hardware: the boost inductor, transformer, gate-driver/power PCB, and configurable load bank. The load selector allows precise adjustment of load symmetry and dynamic steps for repeatable testing. Under symmetric operation, each bipolar output was terminated with a $47\ \Omega$ resistor; activating a parallel branch reduced the equivalent resistance to $23.5\ \Omega$ per output ($47\ \Omega \parallel 47\ \Omega$). An asymmetric condition was created by inserting a $4.7\ \Omega$ series element in branch R_{02} , yielding $R_{01} = 23.5\ \Omega$ and $R_{02} = 28.2\ \Omega$.

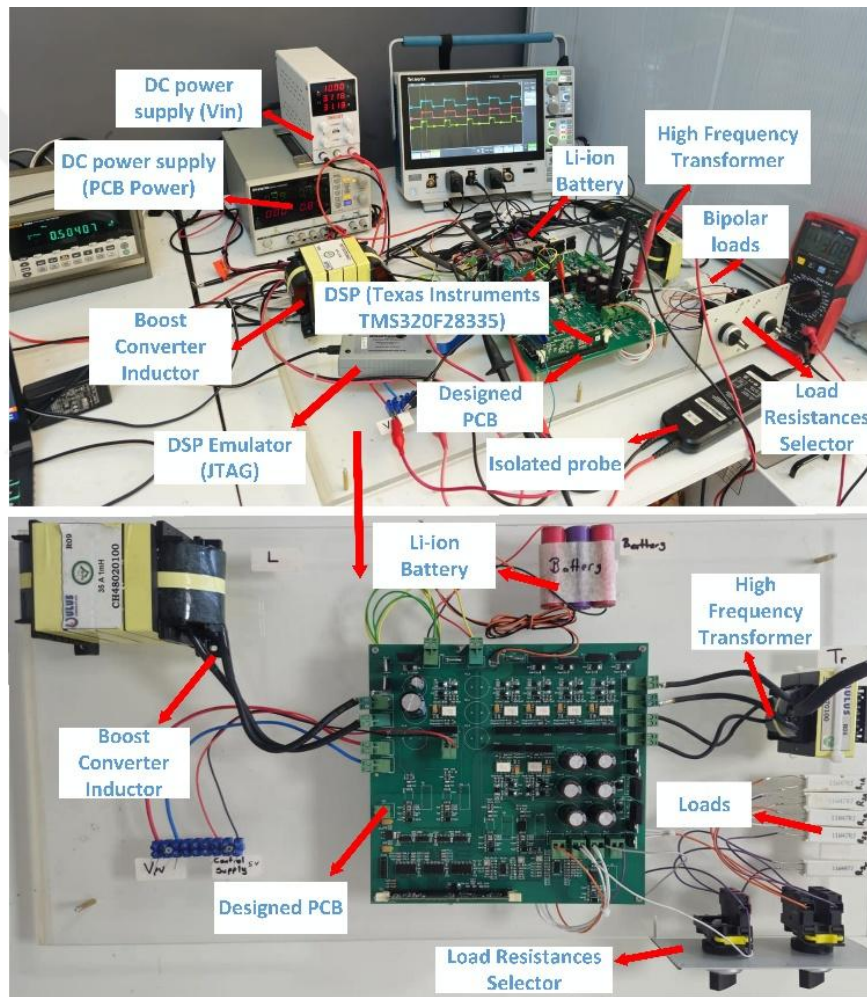


Figure 6.6. Experimental setup of the proposed converter prototype

Representative measurement connections and probe points of the experimental setup are illustrated in Fig. 6.7. In the laboratory prototype, $50\times$ isolated differential voltage probes were used and calibrated in the oscilloscope to measure floating node voltages with high accuracy.

A $0.05\ \Omega$ precision shunt resistor integrated on the PCB provides current feedback to the DSP for control purposes. In the absence of a dedicated current probe, the same shunt signal was mirrored to the oscilloscope, and the instantaneous current was reconstructed as,

$$I(t) = \frac{V_{\text{shunt}}(t)}{0.05\ \Omega}.$$

Each bipolar output voltage was measured independently during symmetric and asymmetric loading conditions.

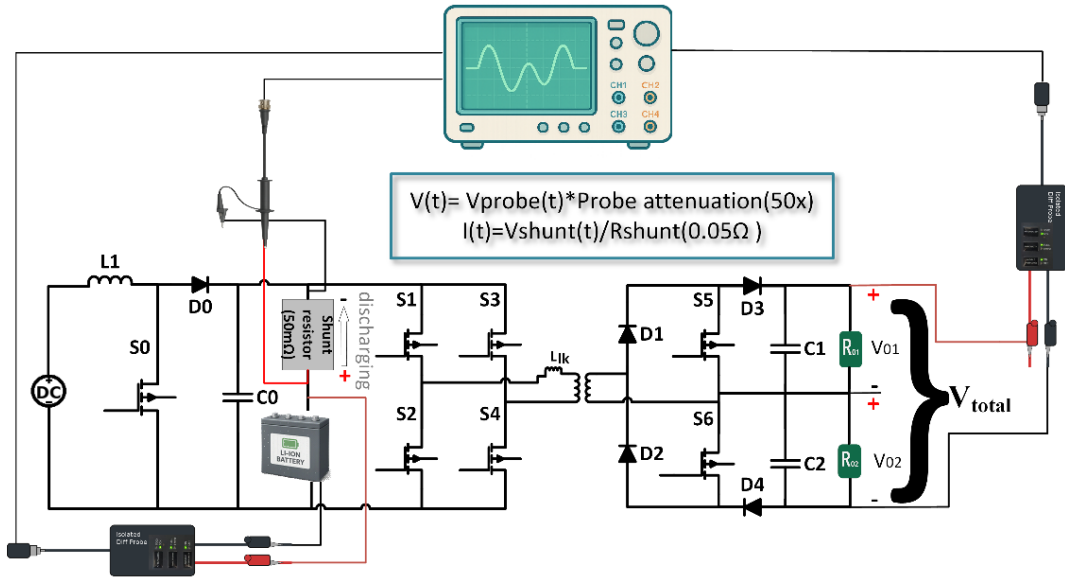


Figure 6.7. Experimental measurement setup and probe connection diagram of the proposed converter

Overall, the prototype successfully integrates sensing, control, and measurement interfaces in a compact setup that enables both static and dynamic evaluation of the proposed converter topology.

6.2. Experimental Results

To evaluate the real-world performance of the proposed battery-coupled bipolar DC–DC converter, a series of experiments were conducted under practical operating conditions. These tests were designed not only to examine the power flow interactions among the input DC source, the battery, and the bipolar outputs, but also to demonstrate the converter’s inherent voltage-balancing capability through symmetrical and asymmetrical load conditions. Although the

converter was designed and prototyped for 1 kW operation, the experimental procedure was executed at lower power levels due to laboratory constraints and safety considerations. In addition, a PV source was not available in the hardware setup; therefore, the MPPT functionality was not evaluated experimentally.

6.2.1. Battery-Coupled Bipolar DC–DC Converter Operation Case Results

To comprehensively assess the functional performance of the proposed converter, three distinct operating cases were defined, each representing a realistic power-flow condition between the input DC source, the battery, and the bipolar outputs.

- **Case I** corresponds to the condition in which only the battery operates as the active source, supplying power to the bipolar bus when the primary DC source, for instance a photovoltaic (PV) array, is not producing energy as during nighttime hours.
- **Case II** represents the situation in which the input DC source simultaneously powers the bipolar outputs and charges the battery.
- **Case III** describes a source-deficit scenario, in which the input source alone cannot sustain the total load demand, and the battery provides supplementary power to maintain the regulated bipolar outputs.

6.2.1.1. Case I

Fig. 6.8 presents the steady-state waveforms obtained when the converter operates solely from the battery, representing conditions in which the primary DC source is inactive. In this mode, the battery provides the required energy to sustain the regulated bipolar outputs.

At the recorded operating point, the battery delivered 2.3 A at a terminal voltage of 10.18 V, maintaining a total bus voltage of 43 V under symmetric operation. These instantaneous values confirm that the converter operated exclusively from the battery while maintaining a regulated bipolar bus voltage of 43 V (± 21.5 V per-output). As the battery energy decreases, the control algorithm dynamically increases the phase shift angle to enhance power transfer and preserve voltage regulation across the bipolar outputs.

The inset in Fig. 6.8 illustrates the phase-shift modulation signals for switches S1–S2 on the primary side and S5–S6 on the secondary side. The converter operates with a fixed 50% duty ratio for each leg, and the phase-shift angle δ between the primary and secondary bridge waveforms determines the instantaneous power transfer. The observed switching transitions exhibit proper dead-time intervals and clean commutation, validating the effectiveness of the gate-driving design and timing synchronization.

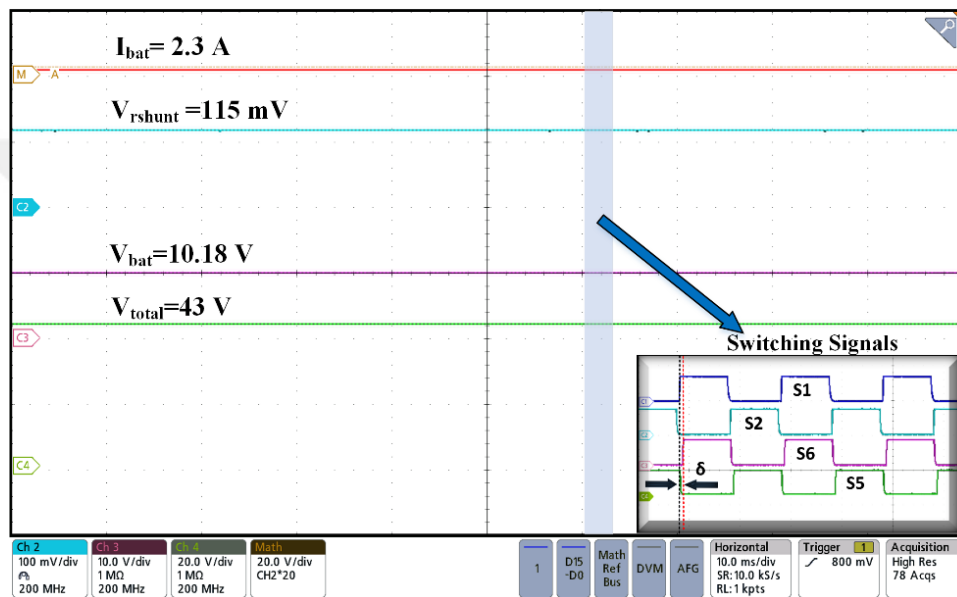


Figure 6.8. Case I operation: battery voltage, battery current, and total bus voltage waveforms

6.2.1.2. Case II

In this case, the input DC source simultaneously supplies the bipolar outputs and charges the battery. Two operating points were evaluated to verify the converter's behavior under different charging-current levels while the total bus remained regulated at 43 V ($\pm 21.5 \text{ V}$ per polarity). All Case II measurements were obtained with the input DC source fixed at 10 V.

As seen in Fig. 6.9 in the first test, the battery current reference was set to 0.5 A, resulting in a measured battery voltage of approximately 11 V. Under these conditions, the converter maintained a stable total bus voltage of 43 V ($\pm 21.5 \text{ V}$ per polarity) without noticeable fluctuation.

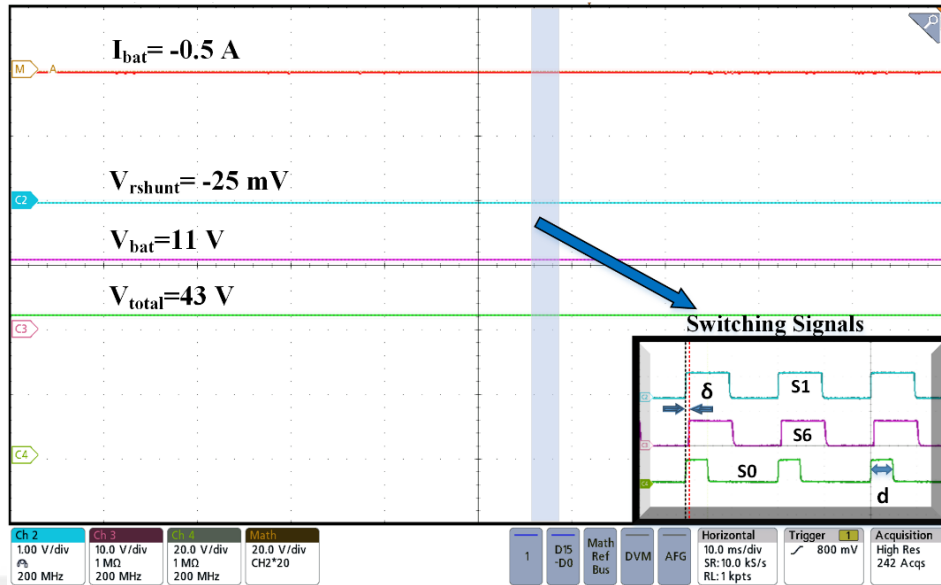


Figure 6.9. Case II operation: battery voltage, battery current, total bus voltage waveforms

Subsequently, the charging current reference was increased to 1 A (Fig. 6.10), and battery voltage was approximately 11.5 V. The total bus voltage was held at 43 V, indicating that voltage regulation was maintained. These measurements confirm that the converter maintains stable bus regulation and precise current control throughout the charging process.

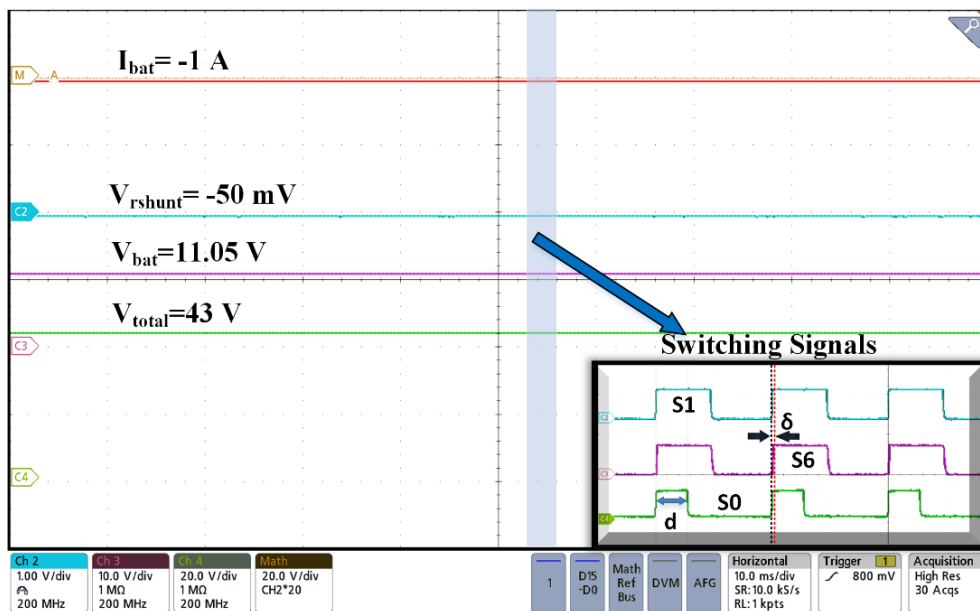


Figure 6.10. Case II operation: battery voltage, battery current, and total bus voltage waveforms

The switching-signal inset shows coordinated operation of the boost stage and the isolated bridge. As the battery voltage increases, the controller gradually reduces the phase-shift angle (δ) because a smaller transfer power is sufficient to sustain the regulated bus at higher battery potential. At the same time, when the charging-current reference is increased, the boost stage duty ratio grows to raise the inductor current and meet the new current command. This coordinated adjustment ensures stable charging behavior and continuous voltage regulation.

6.2.1.3. Case III

In this case, the input DC source was kept constant at 8 V, and the converter operated such that both the source and the battery supplied power to the bipolar outputs. Two experiments were conducted to evaluate performance at different bus-voltage and battery-current references.

At the first operating point (Fig. 6.11), the total-bus reference was set to 43 V and the battery-current reference to 1 A. The measured waveforms show $I_{\text{bat}} = 1.0$ and $V_{\text{bat}} \approx 10.5$ V. Under these conditions the converter maintained stable bus regulation at 43 V while drawing the remaining power from the battery to complement the limited input power.

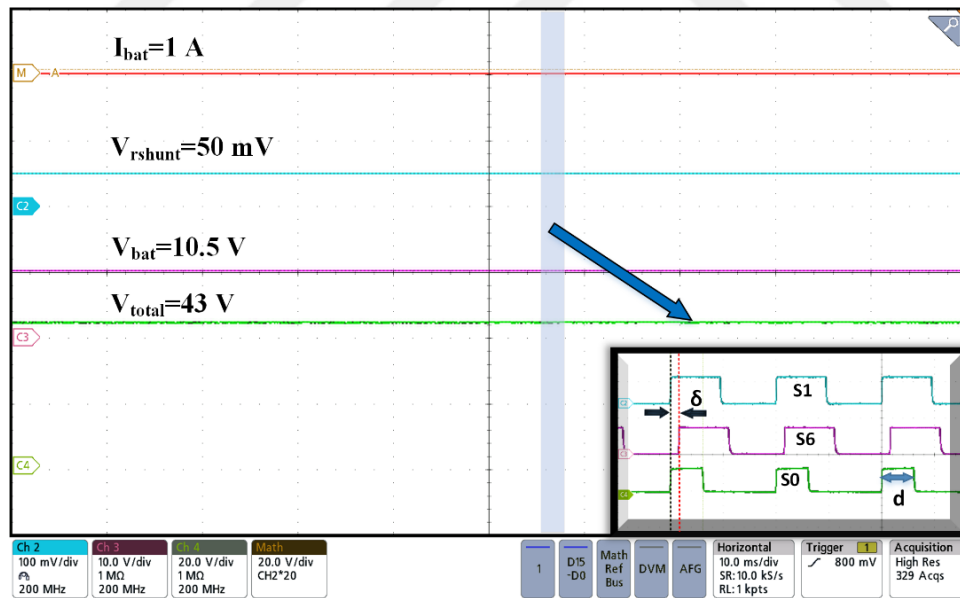


Figure 6.11. Case III operation: battery voltage, battery current, and total bus voltage waveforms

At the second operating point (Fig. 6.12), the total-bus reference was increased to 50 V, and the battery-current reference was set to 1.6 A. The captured signals indicate $I_{\text{bat}} = 1.0$ and $V_{\text{bat}} \approx 10.35$ V. To satisfy the higher bus demand, the control system enlarged the phase-shift angle (δ), thereby increasing transferred power through the isolated stage, while the boost-stage duty ratio was adjusted to sustain the commanded discharge current. In both tests the converter preserved voltage regulation and smooth synchronization between the input, battery, and bipolar outputs.

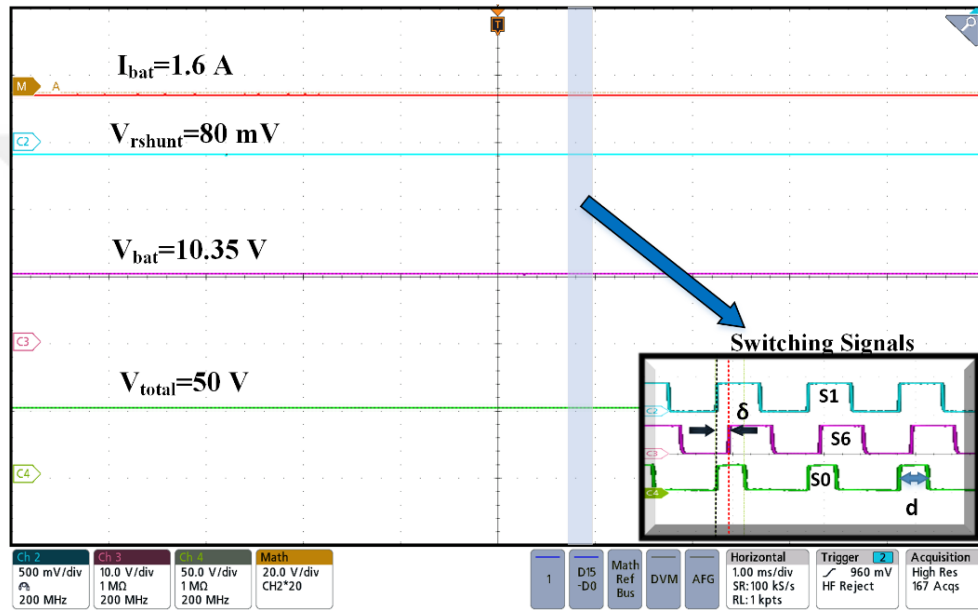


Figure 6.12. Case III operation: battery voltage, battery current, and total bus voltage waveforms

6.2.2. Experimental Verification of the Battery-Coupled Bipolar DC–DC Converter Under Symmetrical and Asymmetrical Load Conditions

To further assess the converter's voltage regulation and inherent balancing characteristics, a series of load variation tests were performed under both symmetrical and asymmetrical loading conditions. These experiments aimed to examine how the converter maintains the total-bus reference and the voltage symmetry between the positive and negative outputs during static and dynamic load transitions. In the symmetrical case, identical resistive loads were connected to both output polarities to verify balanced operation. In contrast, the asymmetrical tests

intentionally introduced load mismatch between the two polarities to evaluate the converter's natural voltage-balancing capability without any auxiliary circuitry.

6.2.2.1. Bipolar Voltage Symmetry under Dynamic Load Conditions

Dynamic load experiments were carried out to evaluate the converter's ability to preserve bipolar voltage symmetry during rapid load transitions. Each output polarity was initially terminated with a $47\ \Omega$ resistive load, and the load condition was abruptly changed to $23.5\ \Omega$ by engaging a parallel $47\ \Omega$ branch through a manual load selector.

Throughout the tests, as seen in Fig. 6.13 the total bus voltage remained regulated at $43\ \text{V}$, corresponding to $\pm 21.5\ \text{V}$ per polarity, with negligible transient deviation. The recorded waveforms show that both polarities remained nearly equal in magnitude before and after the transient load change. These observations indicate an inherent voltage balancing property of the converter.

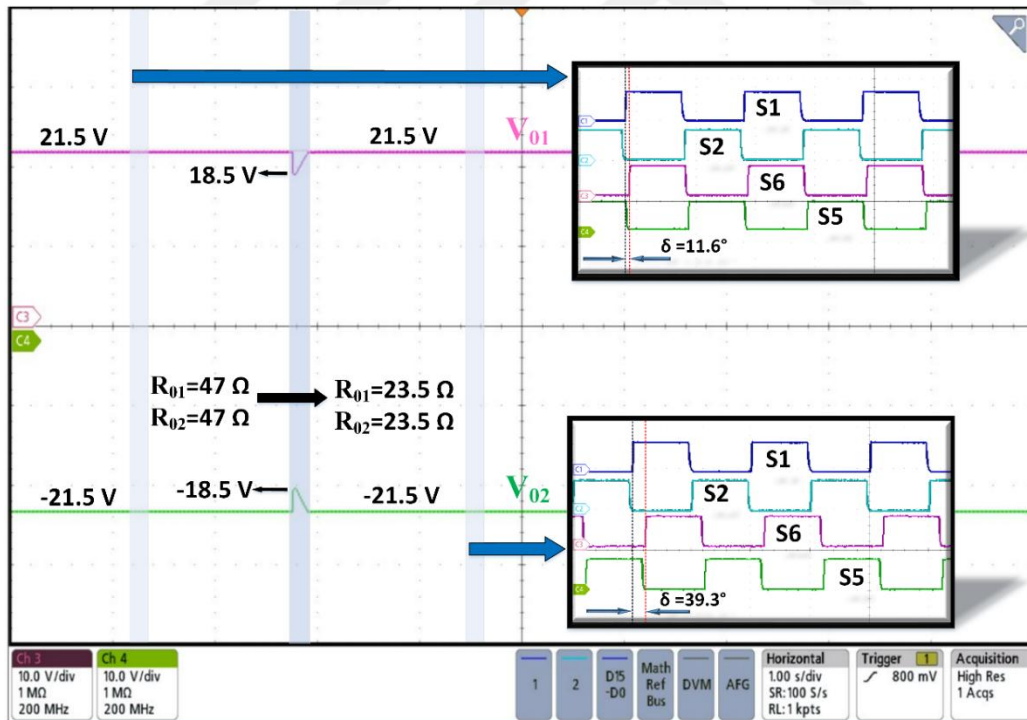


Figure 6.13. Bipolar voltage symmetry under dynamic load condition

6.2.2.2. Performance Under Asymmetric Load Conditions

In this test, the converter was operated under intentional load asymmetry to investigate its natural voltage-balancing capability without controller intervention. The positive output was connected to an equivalent load of $23.5\ \Omega$, while the negative output was changed to $28.2\ \Omega$ by adding a $4.7\ \Omega$ series resistor through a manual load selector, creating approximately a 20% load difference between the two polarities.

When the load selector was switched to introduce the asymmetry, a slight voltage deviation of about 2% was observed between the two outputs (Fig 6.14). These results demonstrate that the proposed topology possesses an inherent self-balancing characteristic, maintaining near-symmetric bipolar voltages even under sudden and unbalanced loading conditions.

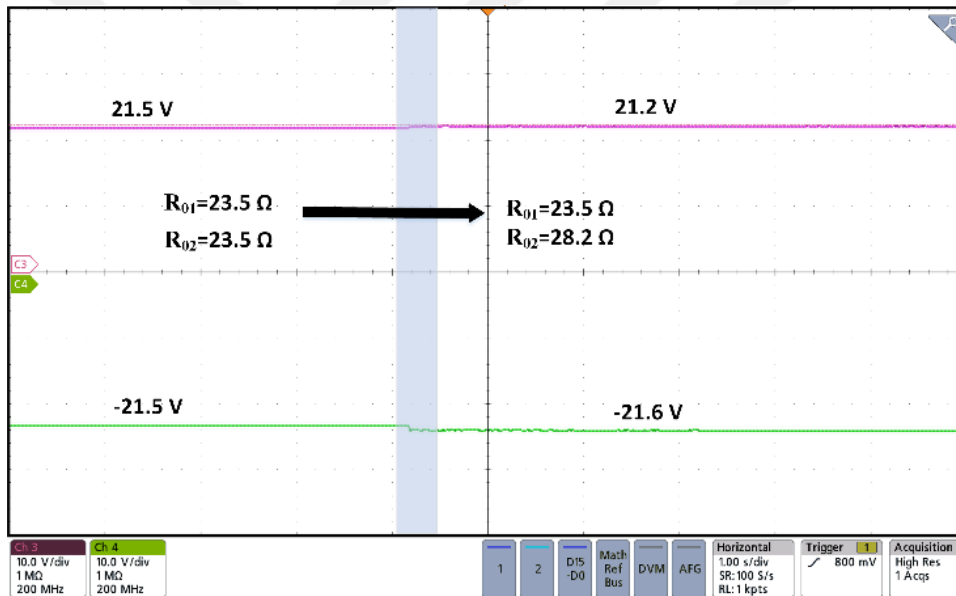


Figure 6.14. Bipolar outputs under asymmetrical load condition

6.2.3. Comparison With Other Studies

In this section, the proposed converter is comparatively evaluated with the related studies by considering the following aspects: Battery and renewable energy source integration, number of switches, balancing method, control method, and the presence or absence of electrical isolation.

Table 6.4. Comparison of the proposed topology with similar studies

Study	Battery / RES Integration	Num. of Switch	Balancing Method	Control Method	Electrical isolation
[181]	DC input source and ESS (through separate converters)	6	Extra voltage balancing circuit. Bipolar voltage balancing control requires two voltage signals and one current signal	One PID loop for the buck converter, one PI loop for ESS control, and one PI loop for bipolar voltage balancing.	No
[182]	Single DC input source	12	Structural balancing via coupled inductor.	SPS control; one output-voltage PI loop	Yes
[183]	DC input source and ESS (through separate converters)	16	Extra voltage balancing circuit. Separate sensing of bipolar outputs.	SPS control with two PI loops for input–output power transfer and bipolar voltage balancing.	Yes
[133]	Single DC input source	12	Structural balancing via T-type bridge. Separate sensing of bipolar outputs.	SPS control; two PI loops for balancing	Yes
[144]	Single DC input source	8	Structural balancing via the coupled inductor	SPS control; one output-voltage PI loop	Yes
[Proposed converter]	DC input source and ESS integrated in one stage	7	Structural balancing via transformer symmetry and natural conduction paths between bipolar outputs	SPS control; one PI loop regulates the bipolar bus voltage. A second PI loop also governs battery power flow by adjusting the duty cycle.	Yes

As shown in the table, when the number of switches is compared, the proposed topology uses the lowest switch count among the galvanically isolated converters. Moreover, studies [182], [133], and [144] provide only a single DC source port. Therefore, with a lower component count, the proposed topology simultaneously interfaces both the DC input source and the energy storage system, while at the same time maintaining the bipolar output voltages.

Among the compared converters, only studies [181] and [183] include ports compatible with energy storage systems; however, in both cases these ports are interfaced through separate

converter stages. Furthermore, [183] does not provide any experimental validation of the voltage balancing capability under operating conditions with a battery. Furthermore, [181] and [183] incorporate auxiliary voltage-balancing branches, whereas [133] attains structural balancing through a T type bridge. In all three topologies, bipolar voltage regulation depends on multiple feedback measurements. This requirement leads to more involved sensing arrangements and a markedly higher control complexity compared with the proposed converter. Studies [182] and [144] employ an additional coupled inductor to achieve active bipolar bus regulation and are fundamentally designed as single-input bus-balancing interfaces whose primary objective is to maintain voltage symmetry under a wide range of load asymmetries. In contrast, the proposed converter integrates the dc source and the energy storage unit within a single stage without any auxiliary balancing circuitry and is intended to coordinate source–battery power flow while simultaneously ensuring balanced bipolar outputs.



7. CONCLUSION AND FUTURE WORK

In recent years, the growing global demand for energy and the pursuit of carbon neutrality have driven significant interest in renewable and decentralized generation. Producing and consuming energy locally has become a key strategy to minimize transmission losses and enhance system resilience. In this context, DC-DC converters play an increasingly crucial role, particularly in localized PV applications where efficient power management and voltage regulation are essential. Moreover, the integration of these converters with energy storage systems has become indispensable to ensure stable and flexible operation under variable solar conditions. Accordingly, this thesis aims to address these contemporary needs by developing a converter topology that enables coordinated operation between renewable generation and storage subsystems. The proposed design has been thoroughly validated through both simulation and experimental studies, demonstrating its capability and practical applicability in modern renewable energy systems.

Operation of the proposed converter was examined in three power flow modes that represent typical microgrid use cases. These modes correspond to battery to load, source to battery and load, and source plus battery to load operation conditions. Each represents a distinct energy flow scenario observed in hybrid microgrid environments where generation, storage, and consumption interact dynamically. The control architecture was designed to be simple, focusing on a practical and easily implementable scheme. In parallel, an ANN based MPPT strategy was developed to account for photovoltaic and battery interaction within the same control framework.

The principal contributions and conclusions of the thesis are summarized below.

- The proposed converter preserved the total bus reference while maintaining balanced bipolar outputs in all investigated cases both in simulation and experimental studies.
- A simple and easily implementable control structure was employed in this study. Regulation of the bipolar DC bus voltage was achieved through a single PI loop, without the need for midpoint sensing or auxiliary balancing circuits. In addition, a separate PI controller applied to the single switch of the boost converter simultaneously managed the power contribution from both the input source and the battery.

- Under symmetric loading, abrupt load steps were tracked without overshoot, and the bipolar voltages returned to nominal values.
- In the experimental tests, with a 20% difference between the two output loads, the bipolar outputs remained balanced within approximately 0.4 V deviation, which corresponds to about 2%. In simulation, even under a higher 60% load difference (50 Ω vs. 80 Ω), the deviation remained around 1 V, further validating the robustness of the proposed topology.
- Experimental verification on a DSP controlled prototype with a real battery confirmed reliable performance under practical operating conditions, and the battery port did not require any additional interfacing stage.
- An ANN based MPPT strategy was developed, treating the photovoltaic array and the battery storage unit as a coupled system. This approach enables coordinated energy management between the source and storage while ensuring accurate tracking of the MPPT under dynamically changing operating conditions.

As a future work, the following studies are suggested;

- Artificial intelligence control algorithms, which have only been studied in a simulation environment, will also be examined experimentally. This step will verify their practical applicability and dynamic performance.
- The objective is to achieve higher voltage gain, lower device stress, and improved efficiency in low-voltage PV applications. High-gain boost converter structures are to be investigated as alternatives to conventional single-stage boost configurations.
- Bidirectional converter configurations enabling controlled power transfer from the load side back to the battery are to be developed and analyzed. These converters aim to

increase operational flexibility and enable regenerative operating modes. They are also expected to support advanced energy management in bipolar DC microgrid architectures.





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APPENDIX A : TECHNICAL SPECIFICATIONS OF DSP BASED MCU

Table A.1. Technical Specifications of TMS320F28335 (Texas Instruments, 2022)

Parameter	Specification
Clock rate	150 MHz
Instruction cycle	6.67 ns
Supply Voltage	Core 1.9 V / 1.8 V, I/O 3.3 V
On-chip flash memory (16-bit word)	256 K
Single-access RAM (SARAM) (16-bit word)	34 K
Code security module	Yes
External interface (XINTF)	16/32-bit
Direct memory access (DMA)	6 channels
PWM outputs	ePWM1–ePWM6
High-resolution PWM (HRPWM)	ePWM1A–ePWM6A
Capture inputs	eCAP1–eCAP6 (32-bit)
Quadrature encoder pulse channels	eQEP1–eQEP2 (32-bit)
Watchdog timer	Yes
ADC resolution	12-bit
ADC channels	16
ADC sampling rate	12.5 MSPS
ADC conversion time	80 ns
CPU timers	3 (32-bit)
Multichannel buffered serial port (McBSP)	2 (A/B)
Serial communications interface (SCI)	3 (A/B/C)
Enhanced controller area network (eCAN)	2 (A/B)
General-purpose I/O pins (shared)	88
External interrupts	8

APPENDIX B: TECHNICAL SPECIFICATIONS OF ISOLATED DIFFERENTIAL AMPLIFIER

Table A.2. Key Specifications of AMC1200-Q1 (Texas Instruments, 2016)

Parameter	Specification
Differential input range	± 250 mV nominal (clipping near ± 320 mV)
Fixed gain	8 (accuracy ± 0.5 %)
Nonlinearity	≤ 0.075 % with 5 V supply
Input offset voltage	typ. ± 0.2 mV, max ± 1.5 mV
Input bandwidth	typ. 100 kHz, min 60 kHz
Common-mode rejection ratio	108 dB (0 Hz), 95 dB (at 50 kHz)
Output noise	typ. 3.1 mV RMS
Output common-mode voltage	≈ 1.29 V (at $V_{DD2} = 3.3$ V), ≈ 2.55 V (at $V_{DD2} = 5$ V)
High-side supply V_{DD1}	4.5 V – 5.5 V (typ. current ≈ 5.4 mA)
Low-side supply V_{DD2}	2.7 V – 5.5 V (typ. current ≈ 3.8 – 4.4 mA)
Isolation rating	UL1577 and VDE V 0884-10 certified, 4250 V peak
Working isolation voltage	1200 V peak
CMTI	≥ 10 kV / μ s (typ. 15 kV / μ s)
Operating temperature range	-40 °C to $+105$ °C
Package	SOIC-8 (DWV)

APPENDIX C: TECHNICAL SPECIFICATIONS OF MOSFET

Table A.3. Key Specifications of FCHD040N65S3 (ON Semiconductor, 2024)

Parameter	Specification
Device type	N-channel SuperFET III Easy Drive MOSFET
Drain–source voltage	650 V
Continuous drain current (25 °C)	65 A
On-resistance	typ. 35.4 mΩ, max 40 mΩ
Gate threshold voltage	2.5 – 4.5 V
Total gate charge	typ. 136 nC
Input capacitance	typ. 4740 pF
Turn-on delay	typ. 35 ns
Turn-off delay	typ. 95 ns
Reverse-recovery time	typ. 534 ns
Avalanche energy	358 mJ
Thermal resistance	0.3 °C/W
Operating junction temperature	–55 °C to +150 °C
Package	TO-247AD