



**REPUBLIC OF TÜRKİYE
ADANA ALPARSLAN TÜRKES SCIENCE AND TECHNOLOGY
UNIVERSITY**

**GRADUATE SCHOOL
ELECTRICAL AND ELECTRONIC ENGINEERING DEPARTMENT**

**DESIGN AND REALIZATION OF A SOLID-STATE TRANSFORMER
FOR SMART GRIDS**

BURAK ESENBOĞA

Ph.D.

ADANA 2023



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Ph.D.

**THESIS ADVISOR
ASSOC. PROF. DR. TUĞÇE DEMİRDELEN**

ADANA, 2023

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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.

01/08/2023

[Signature]

Burak ESENBOĞA

ÖZET

AKILLI ŞEBEKELER İÇİN KATI HAL TRANSFORMATÖRÜNÜN TASARIMI VE GERÇEKLEŞTİRİLMESİ

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Akıllı güç dağıtım sistemleri, gelişmiş ülkelerde artan enerji talebini karşılamak için umut verici teknolojiler olarak ortaya çıkmıştır. Akıllı şebekelerin giderek yaygın olarak benimsenmesi, akım ve gerilim harmoniklerindeki bozulmalar, ani gerilim dalgalanmaları, arızalar ve izolasyonda bozulmalar gibi çeşitli güç kalitesi sorunlarını ortaya çıkarmıştır. Yaygın olarak akıllı transformatörler olarak adlandırılan katı hal transformatörleri, bu sorunların üstesinden gelmek için akıllı şebekelerde güç kalitesini artırmaya yönelik ileriye dönük çözümler sunmaktadır. Bu transformatörler, gelişmiş enerji verimliliği, gelişmiş şebeke güvenilirliği, güç akışı kontrolü, voltaj regülasyonu, çift yönlü güç akışı, arıza akımı sınırlaması, harmonik engelleme ve galvanik izolasyon dahil olmak üzere çok sayıda avantaja sahiptir.

Bu tezde, üç seviyeli (AC-DC-DC-AC) bir katı hal transformatör (KHT) modeli ve akıllı şebekelerle bağlantılı güneş fotovoltaik (PV) enerji santralleri için uyarlanmış bir güç yönetim şeması sunulmaktadır. Önerilen topoloji, doğrultucu, yalıtılmış çift yönlü dönüştürücü ve eviricideki yumuşak anahtarlama özelliklerini içerir. Bu, düşük iletim kayıpları, minimum elektromanyetik girişim (EMI), yüksek verimlilik ve dönüştürücüler için sıfır voltajlı anahtarlama ve elektrikli yardımcı sistemler için sıfır akım anahtarlama ile verimli çalışmayı mümkün kılar. Tez ayrıca, darbe genişliği modülasyonu (PWM) kontrolü ve dönüştürücüler arasındaki voltaj ve akım kontrolü dahil olmak üzere KHT'nin operasyonel stratejilerini ana hatlarıyla belirtir. 10 kVA, gücünde bir katı hal transformatörü elde etmek için orta gerilim (OG) yüksek frekanslı bir transformatör kullanılır.

Deneysel sonuçlar KHT'nin güneş enerjisi santralinin dengesiz güç üretim koşullarında bile optimize edilmiş kontrol stratejisi sayesinde %96,7'lik etkileyici bir dönüştürme verimliliği elde ettiğini göstermektedir. 344 Vac ve 10,4 A yük akımı altında elde edilen deneysel sonuçlar, transformatörün 6,8 kV/us'lik bir dv/dt oranını işleme kapasitesini doğrulamaktadır. Önerilen çift yönlü akıllı transformatörün dinamik ve deneysel sonuçları, fotovoltaik entegrasyonu içeren akıllı güç dağıtım sistemlerindeki güç kalitesi sorunlarını azaltmadaki etkinliğini göstermektedir.

Anahtar Kelimeler: akıllı elektrik şebekesi; akıllı transformatör; güç kalitesi; güneş enerjisi; katı hal transformatörü.



ABSTRACT

DESIGN AND REALIZATION OF A SOLID-STATE TRANSFORMER FOR SMART GRIDS

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Smart power distribution systems have emerged as potential solutions to fulfill the rising demand for energy in developed nations. Smart grids have become increasingly common, but this has led to a number of power quality problems, including distorted current and voltage harmonics, instantaneous voltage changes, failures, and isolation breakdowns. Solid-state transformers, commonly referred to as smart transformers, provide prospective solutions to enhance power quality in smart grids in order to deal with these issues. Numerous advantages are provided by these transformers, including increased grid dependability, greater energy economy, control over power flow, voltage management, bidirectional power flow, fault current limiting, harmonic blocking, and galvanic isolation.

In this thesis, a solid-state transformer (SST) model with three level (AC-DC-DC-AC) and a power management scheme tailored for solar photovoltaic (PV) power plants connected with smart grids are presented. The isolated bidirectional converter, inverter, and rectifier in the suggested topology all have soft-switching components. This offers great efficiency, low electromagnetic interference (EMI), reduced conduction losses, zero-voltage switching for converters, and zero-current switching for electrical auxiliary systems, among other advantages. The thesis also describes the SST's operating tactics, such as voltage and current management between converters and pulse-width modulation (PWM) control. To create a solid-state transformer with a power of 10 kVA, a medium-voltage (MV) high-frequency transformer is used.

The enhanced control technique enables SST to achieve an outstanding conversion efficiency of 96.7% despite the solar power plant's variable power generating conditions, according to experimental results. The transformer's ability to withstand a dv/dt ratio of 6.8 kV/us is confirmed by experimental findings at a load current of 344 Vac and 10.4 A. The suggested bidirectional smart transformer's dynamic and experimental findings show how well they work to mitigate power quality issues in photovoltaic integration-based smart power distribution systems.

Keywords: power quality; smart power grid; smart transformer; solar PV; solid-state transformer.



Dedication

In the preparation of this thesis, I would like to thank my advisor Assoc. Prof. Dr. Tuğçe DEMİRDELEN who has always been exemplary of her knowledge, business approach, skills and a broad perspective.

I would like to thank my dear wife Mrs. Tuba ESENBOĞA, who has always inspired me with her love of education, desire to learn and hard work, who has encouraged me in every project and job in my professional life, and who has not spared her endless belief and support, especially during this thesis process. I would like to thank my son Mr. Erol Mahir ESENBOĞA, who participated in my thesis process, refreshed my life energy and gave me hope for my future work.

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LIST OF ABBREVIATIONS

SST	: Solid State Transformer
PV	: Photovoltaic
PWM	: Pulse Width Modulation
EMI	: Electromagnetic Interference
MV	: Medium Voltage
LV	: Low Voltage
FACTS	: Flexible AC Transmission Systems
STATCOM	: Static Synchronous Compensators
UPFC	: Unified Power Flow Controllers
AC	: Alternating Current
DC	: Direct Current
DAB	: Dual Active Bridge
ZVS	: Zero Voltage Switching
ZCS	: Zero Current Switching
OLC	: Open Loop Control
MMC	: Modular Multilevel Converter
ISOP	: Input-Series and Output-Parallel
Q2L	: Quasi-2-Level
HF	: High-Frequency
SRC	: Series Resonant Converter
QAB	: Quadruple-Active-Bridge
DES	: Distributed Energy Storage
DER	: Distributed Energy Resource
LF	: Low Frequency
AWG	: American Wire Gauge
MPSST	: Multi-Port Solid-State Transformer
CHB	: Cascaded H-Bridge
NPC	: Neutral Point Clamped
HC-DCM SRC	: Half-Cycle Discontinuous-Conduction-Mode Series Resonant Converter
DSP	: Digital Signal Processor

PR	: Proportional Resonant
SPWM	: Single Pulse Width Modulation
RMS	: Root-Mean-Square
MPPT	: Maximum Power Point Tracking
WK	: Watts Per Kilogram
FEA	: Finite Element Analysis
NTC	: Negative Temperature Coefficient
HFT	: High-Frequency Transformer



LIST OF SYMBOLS

δ	: Skin Depth
D_{AWG}	: Wire Diameter
A_w	: Base Wire Area
T	: Time Period
t_{on}	: Maximum Transistor on Time
t	: time
P_{o1}	: Power out in secondary No.1
P_{o2}	: Power out in secondary No.2
P_T or P_{omax}	: Total Secondary Power
$I_{pri,max}$	: Max Primary Current
$R_{in,eq}$	: Equivalent Input Resistance
L	: Primary Inductance
$W-S$	: Energy Handling Capability
K_e	: Electrical Condition
K_g	: Core Geometry
J	: Current Density
$A_{pw}(B)$	: Primary Wire Area
S_{np}	: Required Number of Primary Strands
N_p	: Number of Primary Turns
l_g	: Required Gap
F	: Fringing Flux Factor
B_{pk}	: Peak Flux Density
R_p	: Primary Winding Resistance
N_{s01}	: Secondary Turns
$I_{s01(pk)}$	: Secondary Peak Current
$A_{sw01}(B)$	: Secondary Wire Area
S_{ns01}	: Required Number of Secondary Strands
R_{s01}	: Winding Resistance

N_{s02}	: Secondary Turns
$I_{s02(pk)}$	: Secondary Peak Current
$A_{sw02(B)}$	: Secondary Wire Area
S_{ns02}	: Required Number of Secondary Strands
R_{S02}	: Winding Resistance
K_u	: Window Utilization
P_{cu}	: Total Copper Loss
α	: Regulation
WK	: Watts per Kilogram
B_{ac}	: AC Flux Density
P_{fe}	: Core Loss
P_{Σ}	: P_{fe} and Copper P_{cu} , in watts
Ψ	: Watt Density
T_r	: Rising Temperature
V_{OL}	: Maximum Low-Level Output Voltage
I_{CC}	: Maximum Supply Current
R_g	: Gate Resistor
P_T	: Total Power Consumption Of The Gate Driver
P_E	: Emitter Power
P_O	: Output Power
I_{sc}	: Short Circuit Current
V_{oc}	: Open Circuit Voltage
L_p	: Primary Leakage Inductance
L_m	: Magnetizing Inductance
V_{P1}	: H-Bridge Input Voltage
f_s	: Switching Frequency
φ	: Phase Difference Between The Two Bridge Circuits
L_{sst}	: MV side bridge circuit's leakage inductance
T_{hs}	: One-Half Of The Switching Time
I_{sst1}	: Primary Leakage Inductance Current
I_{sst2}	: Secondary Leakage Inductance Current
V_d	: Voltage Drop

1. INTRODUCTION

Power generation, transmission, and distribution are the fundamental components of the modern power system. Among them, power transformers play a vital role by efficiently increasing or decreasing voltage levels. They facilitate the long-distance transmission of electricity by stepping up the voltage during generation and stepping it down for distribution to industrial, commercial, and residential consumers. Traditionally, the focus of power transformer development has been on improving magnetic and insulation materials, manufacturing processes, and cost-effectiveness. However, with recent technological advancements, power electronics have emerged as a promising technology for empowering future smart grids across all levels of electrical power systems.

Power electronics, along with other innovations, have found extensive applications in distribution and transmission power systems. It is utilized in high-voltage DC (HVDC) transmission systems, flexible AC transmission systems (FACTS) devices such as static Var compensators, static synchronous compensators (STATCOM), unified power flow controllers (UPFC), and renewable energy systems, particularly in scenarios with significant integration of solar and wind resources. In the past decade, a noteworthy high-power converter called the solid-state transformer (SST), also known as the power electronic transformer or intelligent universal transformer, has gained considerable attention and has undergone extensive research for distribution systems. The initial concept of the SST, known as the electronic transformer, occurred in marking the beginning of its exploration.

In recent years, significant progress has been made in the development of low-voltage and power prototypes for SSTs. However, their adoption in distribution systems remains limited due to the constraints imposed by the voltage and power ratings of available power devices and circuit topologies. The underlying concept of SSTs revolves around achieving voltage transformation through medium- to high-frequency isolation. This approach holds the potential to reduce the size and weight of the transformer compared to traditional power transformers.

In an SST, the standard 50/60-Hz AC voltage is converted into a high-frequency voltage ranging from several kilohertz to tens of kilohertz. This high-frequency voltage is then stepped up or down using a high-frequency transformer, resulting in a significantly reduced volume and weight. Finally, the voltage is converted back to the desired 50/60-Hz frequency to supply power to the load. The configuration of the SST offers additional functionalities that are not present in traditional transformers. By employing solid-state semiconductor devices and circuits, similar to FACTS devices, voltage, and current regulation become feasible. This enables advantageous features such as power flow control, voltage sag compensation, fault current limitation, and more, which are not possible with traditional transformers. Furthermore, voltage source converters connected to the secondary terminal of the SST can support a regulated DC bus. This opens up the possibility of connecting the SST to DC microgrids, allowing for innovative microgrid architectures.

Although the concept of the SST is straightforward, its design and implementation are challenging. The SST essentially represents a high-voltage and high-power electronic circuit, presenting difficulties in achieving reliable operation. Additionally, the SST comprises various components beyond the high-frequency transformer, including power devices, gate drivers, heatsinks, control circuits, cooling systems, auxiliary power, and ancillary circuitry. Therefore, achieving the desired size and weight reduction in the SST compared to traditional transformers requires meticulous design. Effective reduction in size and weight can only be realized when the high-frequency transformer constitutes a reasonably larger proportion of the overall structure. Consequently, substantial efforts are still required to accelerate the development of SSTs.

This thesis study introduces a novel approach by utilizing a three-level solid-state transformer model in a real smart power grid for the first time. The implemented SST model, based on the dual active bridge (DAB) converter, is incorporated into an on-grid solar power system to address power imbalances caused by varying output power under partial shading conditions. The soft-switching capabilities of the proposed SST offer advantages such as minimized conduction loss, reduced electromagnetic interference (EMI), greater efficacy, and regulated power distribution to consumers.

The management approach and driving techniques in the converters are focused on achieving conversion efficiencies of over 98% for the DAB converter and 96% for the SST to achieve effective power conversion. In an actual solar PV power plant, the offered SST topology and its user-friendly control mechanism are experimentally evaluated. By assessing ZVS gain, dv/dt ratio, and efficiency of conversion, the SST's capacity to control power during sudden fluctuations in solar PV plants is demonstrated. Additionally, the capacity for bidirectional power flow makes it possible to link suppliers or loads.

The SST's recommended control technique is flexible and simple to implement, and it successfully addresses power quality problems such as sudden power fluctuations. In particular in power grids that experience unplanned electrical flow from linked sources like grid-connected renewable power plants, this soft-switching SST gives power control capabilities and supports efficient utilization of electrical energy.

This thesis is structured into the following sections. A summary of several DC-DC converter structures utilized in various SST applications is given in the literature review section. The literature review evaluates converter topologies based on switching characteristics, application areas, frequency, isolation level, zero voltage switching (ZVS), and zero current switchings (ZCS) operations. The material and methods section presents the proposed SST topology, materials, and operation principle of the converters, and control schemes. The main focus of the results and discussion sections is the development of a dynamic model in MATLAB/Simulink to simulate the effectiveness of the offered pv combined power grid based on a PV solar system and the SST. The proposed SST model's design and control mechanism are analyzed. Experimental and simulation studies for the proposed SST Application are presented in detail. Finally, The conclusion section highlights the conclusions drawn from simulation and experimental findings and emphasizes how well the suggested SST accomplished its goals.

2. LITERATURE REVIEW

Concerns over energy quality have emerged in smart power grids as a result of the growing deployment of renewable energies and distributed generation. SSTs have developed as crucial and clever technologies to address these issues, making it possible to integrate both renewable and non-renewable power sources into a unified electrical grid. SSTs are prospective replacements for traditional power transformers because they offer a number of advantages. These benefits include harmonic blocking, galvanic isolation, fault current limiting, power flow management, voltage regulation, reactive power compensation, and bidirectional power flow.

SSTs stand out due to their notable reduction in weight and size compared to traditional power transformers. This advantage is accomplished by incorporating a dc-dc isolated converter structure within the SST. At the heart of the SST lies the dc-dc conversion stage, enabled by a high-frequency transformer. This crucial component meets important criteria, including higher conversion rate, excellent power density, controllable electric flow, enhanced isolation, and compact dimensions.

In complex electrical power grids, the dc-dc conversion stage plays a critical role in power control and ensuring grid reliability. To meet these essential requirements across various application areas, researchers have conducted extensive studies on SSTs integrated into smart grids. A literature review has been conducted to summarize the existing studies on dc-dc converter-based SSTs, providing valuable insights into this field of research. For further details, please refer to Table 2.1 in the literature review section.

The solid-state transformers are introduced, which utilize a three-stage modular multilevel converter [1-7]. The ISOP converter employed in the smart transformer utilizes an open-loop control (OLC) method and ensures satisfactory isolation [1]. The solution utilizes a modular multilevel converter to improve the power quality, minimize voltage harmonics and dv/dt stress, and remove reverse electric current flow. The implementation of the dual shift control method ensures efficient voltage regulation and balanced current distribution [2]. Good conversion ratio, ZCS operation, adjustable conversion, and phase shift management are all

aspects of the modular multilevel converter (MMC). Its modular design makes it simple to get around faults, and its resonant operation mode makes ZCS easier. The converter's inductorless design also delivers excellent efficiency and cost-effectiveness [3]. The recommended converter design is perfect for boosting the converter power rating and attaining great efficiency thanks to the power switches' zero voltage switching function. Voltage balancing within the converter is made possible by the control strategy, which combines a PI controller with phase shift modulation [4]. In order to limit freewheeling currents of the DAB, which would otherwise result in higher losses and a lower effective duty cycle, a 100 kVA SST-based power station is designed using an isolated unidirectional power converter with an ISOP modular topology. Simulation results in the Matlab-Simulink environment demonstrate the effectiveness of this approach. A proportional resonant (PR) controller performs better while monitoring a sinusoidal reference signal than a PI controller. The standard PR controller, however, faces stability issues because of its constrained bandwidth. In order to achieve a stable system with low steady-state error and reliable output, an updated PR controller is constructed [5]. A quasi-2-Level (Q2L) modulation approach, renowned for its potency and improved current transfer capability, is suggested for control applications. The high-frequency (HF) transformer's dv/dt stress is reduced thanks to this control strategy. The outcomes show that electromagnetic interference (EMI) is decreased and capacitor voltage balance is successfully attained by the Q2L modulation technique [6].

A series resonant converter's proposed architecture is presented [8–18] in order to satisfy the phase modular SSTs' constant voltage-transfer requirement. The chosen operating principle for the isolation stage is a half-cycle discontinuous conduction mode, which ensures efficient zero current switching. However, this operating strategy lacks power flow control and allows the circulation of power fluctuations from the medium voltage (MV) side through the series resonant converter (SRC) to the common low voltage (LV) side [10]. To simplify system control, a new single-stage control method is implemented. By using a quasi-resonant frequency mode of operation, this converter can maintain a constant voltage for the common DC-side voltage [8]. By reconfiguring the conventional series resonant converter with a voltage doubler rectifier, voltage stability is achieved. This modified converter structure facilitates post-fault operation and reduces the need for additional power devices [12]. For next-generation DC

data centers, a CLLC resonant converter is utilized, offering zero voltage switching (ZVS) operation for primary power switches. Synchronous rectifier control is employed on the output side power switches to minimize power flow losses. The effectiveness of the conversion is increased when the switching frequency and resonant frequency are used simultaneously. The planned converter architecture exhibits good power density and great peak efficiency of 98% [14]. To enhance system efficiency, the proposed dc-dc converter employs resonant soft-switching modulation as a control technique, effectively reducing losses associated with switching transitions [18].

The medium voltage DC-DC isolation stage for the SSTs is suggested to be a DAB converter [19-47]. Numerous benefits, including ZVS, easy control, high electrical isolation, and bidirectional power flow, are provided by this converter. However, switching losses and a reduction in the rate of voltage change (dv/dt) are brought on by the presence of high parasitic capacitances at high frequency and voltage levels [19]. An extended phase shift control approach is used to reduce power losses since it enables adjusting the direction and amount of power flow. This control method successfully lowers backflow power [30]. For voltage control in the DAB, the phase angle control method is used due to its simplicity and ease of implementation. Additionally, a star-delta connection is employed in the HF transformer to reduce the turn ratio. The HF transformer's small and precisely wound turns allow ZVS operation [40]. For the isolation stage of the ST, a soft-shift modulation control method is adopted to reduce the impacts of harmonics and inrush current. The DC-link capacitor voltages of the power converter are controlled by this control mechanism, which also maintains voltage balance [44]. The utility grid is planned to use a power electronic-based adaptive universal transformer with an LLC-based converter acting as the isolation stage for soft-switching operation. A control device based on digital signal processors manages the voltage and power flow. Due to the leakage inductances in the proposed converter, it has a low magnetizing current and can operate at a maximum efficiency of 98% under light load conditions [46]. An commercial model SST with a 500 kVA output is given a medium-frequency isolated transformer structure. Medium-frequency transformers and cascaded voltage source converters make up the isolation step. A digital signal processor (DSP) controls the imbalanced power

output by means of PWM control and a current/voltage control structure. The DSP also makes fault protection and monitoring activities possible [47].

The framework of a modular ST structure is used to examine a Quadruple-Active-Bridge (QAB) DC-DC converter [48–50]. This converter topology is chosen because it enables soft switching and calls for fewer high-frequency transformers. While the converter is controlled by triangle current modulation, the power switches are regulated using the duty cycle method. This technique effectively reduces the circulating reactive energy in the SST's isolation stage [48]. This converter type reduces the need for high-frequency transformers by offering additional bridges for connecting them. To accomplish the soft-switching operation of the power switches, which reduces power switching losses and enables zero-current switching operation for large loads, the triangle current-mode modulation control is used [49].

The use of smart transformers, or SSTs, in intelligent power networks is explored in the literature study. It demonstrates the widespread use of resonant converters, MMC, DAB converters, and QAB converters in SST applications. The research identifies various primary DC-DC conversion structures, each with a unique control scheme to accommodate various applications and power needs. In the context of this study, SST models specifically designed for solar PV power systems are presented in References [51-56]. To enhance the effectiveness of the dual active bridge converter in various voltage conversion scenarios and reduce dual-side backflow power, on the basis of the DAB converter a single-stage SST is created. To demonstrate their efficacy, the recommended technique uses improved single-phase shift control and triple-phase shift control techniques in the DAB converter. The proposed control technique enables the SST model based on the DAB converter to attain an amazing efficiency of 96% [51]. Power converters for clean energy systems and electrically powered trains are two examples of MV grid-tied applications that use single level loosely connected resonant DAB converters. Instead of employing high-frequency transformers found in traditional SSTs, loosely coupled coils are employed. The DAB converter and SST model proposed in this study demonstrate efficiencies of 97.4% and 95.2%, respectively, at a power level of 2.4 kW [52].

A novel forward DAB topology is introduced for smart transformers. This topology offers several advantages over unidirectional topologies, including reduced volume by minimizing the number of active switches. The proposed ST topology demonstrates superiority in three key aspects: higher power density utilizing commercially available power modules, reduced component count, and simplified control [53]. For cascaded multilayer converter-based SST topologies, a reactive power distribution strategy is recommended for solar PV systems. With this method, the issue of declining solar cell efficiency brought on by increasing reactive power and imbalance is resolved. By enabling particular power cells to withstand more AC for a given active power in unbalanced conditions, the method effectively lowers the unbalanced power on the solar cells [54]. An integrated three-level SST and solar PV power plant management approach and simulation model are described. The device includes a DAB converter, a three-phase voltage source control inverter, and a PWM rectifier. A review of the SST's performance under varying solar radiation, load requirement, and anomalies in the grid supply reveals that it operates satisfactorily and is resistant to field disturbances [55]. Using a MATLAB/Simpower lab prototype, the offered DAB integrated SST model for solar PV systems is verified. Simulation and experiment findings validate the efficient grid and load integration of solar PV modules in the proposed ST. According to the phase shift between power switches, the DAB architecture enhances performance and power transmission, and the LCL filter utilized effectively suppresses undesired harmonics [56].

Table 2.1. A survey of the literature on SSTs based on dc-dc converters

Categories of DC-DC Converters-Based SSTs		References
Converter Topologies	Modular Multilevel Converter (MMC)	1-7
	Series Resonant Converter (SRC)	8-18
	Dual Active Bridge (DAB)	19-45
	Quadruple Active Bridge (QAB)	48-50
	LLC Converter	8, 9, 46
Power Rating	High Voltage	1,4, 5, 8, 10, 11, 20, 40, 41, 43, 47, 48
	Medium Voltage	19, 21, 9, 22, 23, 46, 25, 28, 3, 12, 29, 13, 31, 32, 14, 34, 15, 35, 36, 50, 16, 17, 42, 7, 18, 45
	Low Voltage	2, 6, 24, 26, 27, 30, 37, 38, 39, 44, 49
Switching Frequency	≤ 1 kHz	6, 22, 27, 31
	>1 kHz, 100 kHz $<$	1-5, 8-13, 15-21, 23, 24, 27-30, 32, 33, 36-40, 42-44, 46-50
	≥ 100 kHz	7, 14, 17, 34, 35
Control Method	Pulse Width Modulation	3, 15, 24, 35, 34, 42
	Voltage or Current Control	27, 29, 31, 39, 42, 43, 46
	Frequency Control	17
	Phase Control	2, 4, 5, 7, 21, 23, 25, 26, 28, 30, 32, 34, 37, 40, 41, 43, 45, 47
	Quasi-2-Level (Q2L) Modulation	6
	Trapezoidal Modulation	36
	Triangular-Current Mode	11, 48, 49
	Discontinuous Conduction Mode	13
	Duty Cycle Control	9, 23, 35
	D-Q Control	19, 20
	Open Loop Control (OLC)	1, 12, 16, 23, 36

3. MATERIALS AND METHODS

3.1. Solid State Transformer

SST development is currently in the research and development stage at the moment. These transformers are significantly smaller and less expensive than conventional transformers because they employ sensible semiconductor methods to produce great power density at high frequencies. In comparison to their similar counterparts, SSTs can be up to three times smaller and lighter. Additionally, because they do not require fluid or liquid dielectrics for cooling, they are more environmentally friendly.

The research and development phase of SST development is underway. These transformers use sensible semiconductor techniques to provide a high power density at high frequencies, making them substantially smaller and less expensive than conventional transformers. SSTs can be up to three times smaller and lighter than their comparable counterparts. They are also more environmentally friendly because they don't need fluid or liquid dielectrics for cooling.

SSTs provide several benefits, including reduced weight and small size, the use of active controlled devices, the ability to regulate voltage, the ability to isolate faults, rectify power factors, minimize harmonics, be cost-effective, and be environmentally friendly.

Several studies and research institutions are currently pursuing research and testing prototypes for particular applications, even though SSTs have not yet reached the commercialization level. Within the next year or so, the SST market is predicted to be commercialized, and it has substantial growth potential in the years to come. McMurray initially proposed the idea of SSTs in 1968, emphasizing solid-state switches and HF isolation. SSTs were acknowledged by the Massachusetts Institute of Technology as one of the top ten most promising technologies and by the National Science Foundation's Generation-III Engineering Research Center [57].

The main challenge in implementing SSTs lies in selecting the most suitable topology. Extensive analysis of various existing topologies is being conducted to identify the most suitable one for achieving unidirectional power flow. Some topologies and configurations do

not support bi-directional power flow, and efforts are being made to categorize SST topologies and determine the most appropriate options based on specific requirements and needs.

SST topologies can be broadly categorized into single-stage, two-stage, and three-stage configurations. These topologies enable the conversion of input DC or AC to output DC or AC by manipulating the switches to achieve bipolar (AC mode) or unipolar (DC mode) voltage and current settings [58]. Figure 3.1 presents the classification of SSTs based on their configurations.

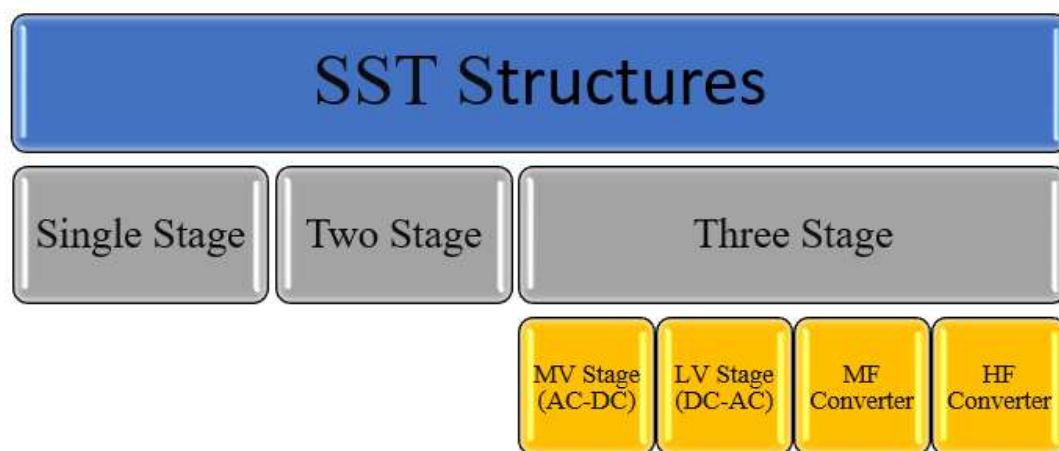


Figure 3.1. Different SST structures

All conversion stage of the SST is also given in Figure 3.2 [59].

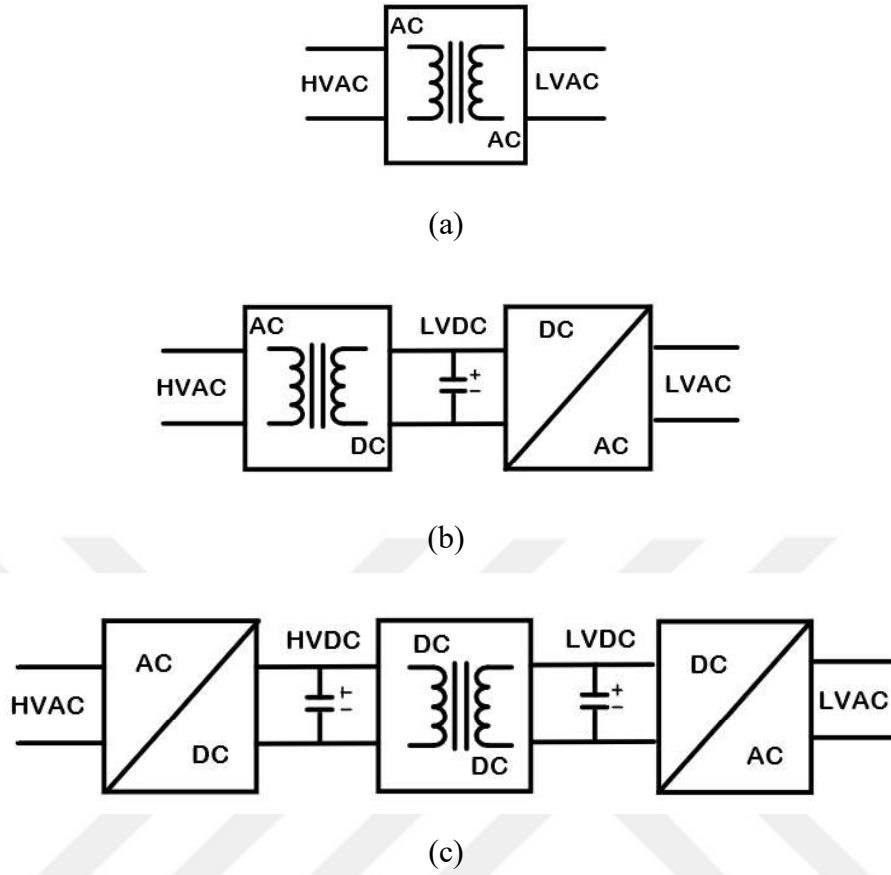


Figure 3.2. SST power conversion level a) single-stage, b) two-stage with LVDC connection, c) three-stages [59]

3.1.1. Single-Stage SST Structure

The single-stage SST topology is a straightforward and efficient design that focuses on direct power conversion with transformer isolation to step down high voltage to low voltage. It operates from direct current to direct current or from alternating current to alternating current power conversion stages, offering advantages such as enhanced efficiency and increased reliability. Compared to the more complex three-stage SST topology, the single-stage configuration eliminates conduction and switching losses, resulting in improved performance. This topology operates at a high switching frequency. There is no need to require electrolyte capacitors, further enhancing its reliability. The control of power transfer in the single-stage topology is achieved by manipulating the phase shift angle between the secondary and primary bridges. With the use of bi-directional power flow and four-quadrant switches, the control strategy is improved, enabling effective regulation of harmonic content. A single-stage SST

cannot solely rely on the HF-Link for development because the power transistor used in the HF-Link cannot block the negative input voltage during the negative half cycle of the AC input voltage. To overcome this limitation, an Low Frequency (LF) Bridge is introduced to convert the bipolar input voltage into a unipolar input voltage. This enables efficient transfer from the primary side to the secondary side using the HF-Link. The unipolar voltage is then converted back to the bipolar output voltage.

There are two approaches to realizing the low-frequency commutation bridge: the AC-switch-based approach and the folding-unfolding-based approach. In the folding-unfolding structure, two full bridges are placed between the input voltage and the HF-Link, as well as between the HF-Link and the output load. On the other hand, the AC-switch structure modifies the switches within the HF-Link to perform the low-frequency commutation. The folding-unfolding approach offers the advantage of lower total conduction losses compared to the AC-switch approach, particularly when used with the same HF-Link design. This is because the current of the HF-Link is higher than the input or output low-frequency current. When the AC-switch approach is used, all MOSFET switches are utilized, resulting in higher losses. However, for low-power SST applications, the AC-switch approach may be preferred due to its lower switch count when combined with a half-bridge HF-Link converter.

Figure 3.3 exhibits a resemblance to a standard DC/DC dual active bridge (DAB), where the power transfer direction and magnitude are regulated by the phase shift between the secondary and primary bridges.

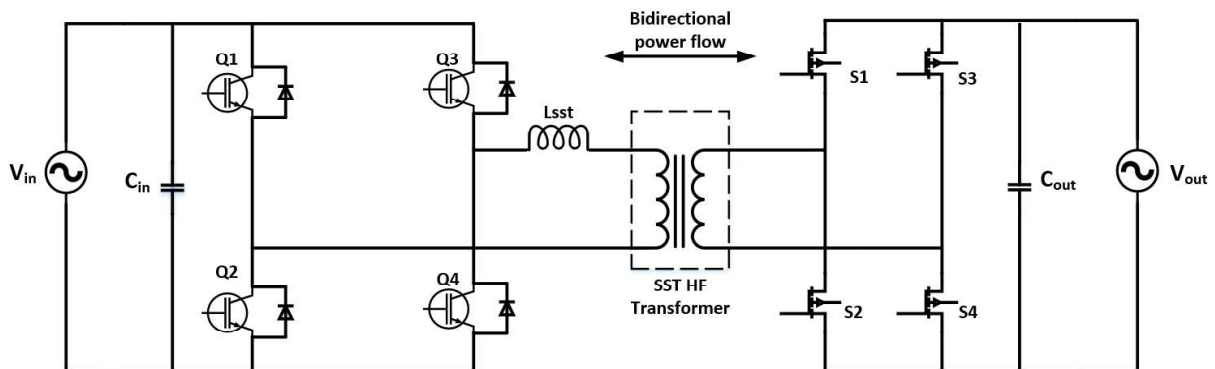


Figure 3.3. Single-stage SST structure

Figure 3.3 depicts an AC-AC full-bridge converter that utilizes a straightforward control method to convert the power into a high-frequency square wave with a 50% duty cycle. This square wave is then rectified to produce a sinusoidal waveform on the low-voltage (LV) side.

3.1.2. Two-Stage SST Structure

The introduction of a DC-link, either on the HV or LV side, is the primary difference between single-stage and two-stage SST. Both topologies allow for the passage of power in both directions, although the two-stage SST can support only one way by adding an unregulated rectifier to the LV side of the AC-DC converter. The complexity of the system is raised by the need for a complicated switching method and more switching devices as a result of this modification. However, the two-stage SST has benefits like compensating for reactive power. The two-stage SST with LVDC link is not suited for HV operations due to the difficulty in sustaining ZVS across a wide input voltage range. Although the lack of an LVDC link restricts the use of renewable energy at the LV level, this problem can be solved by switching from LVAC to HVDC. As an alternative, the three-stage SST architecture offers an answer to these problems.

A recent proposal introduces a novel AC-AC three-phase input and three-phase-four-wire output SST based on high-frequency matrix technology. This topology utilizes PWM to convert the utility source voltage into 50% duty square wave pulses, which are then transmitted through a single-phase high-frequency transformer. This design offers several advantages, including bidirectional power flow, flexible voltage transfer ratio, controllable input power factor, high power density, and electrical galvanometer isolation. Additionally, the topology employs modified three-dimensional spaced vector PWM (3D SVPWM) to reduce harmonic distortion in both input and output currents, particularly under unbalanced load conditions.

Due to a large amount of energy voltage and inadequate grid isolation facilities., the inclusion of a direct current (DC) link in this design may not be ideal for the integration of distributed energy storage (DES) and distributed energy resources (DER). This configuration might not be implementable in real-world SST applications. Additionally, this configuration uses a two-stage

conversion process, with the DC-DC converter stage addressing both galvanic isolation and voltage step-down. It's important to note that one of these configurations lacks a low-voltage DC link.

3.1.3. Three-Stage SST Structure

After conducting a critical analysis of different topologies for implementing a SST, it was determined that the three-stage configuration is the most suitable and practical choice. This configuration offers additional functionalities beyond those of a regular transformer, such as on-demand reactive power support, voltage regulation, power quality improvement, and current limiting.

A single-stage SST topology, which lacks a DC link, was found to be unsuitable for integrating renewable energy sources and energy storage devices. It also lacks provisions for voltage regulation and reactive power compensation, making it inadequate for future grid requirements.

On the other hand, a two-level SST connected a HV DC link was deemed the least suitable option due to the absence of grid isolation, making it incompatible with smart grid requirements. Additionally, this topology does not cater to distributed energy storage and integration of distributed energy resources. The need for large filters to mitigate ripple currents presents a major challenge for achieving voltage regulation in this configuration.

In contrast, the three-stage topology, incorporating both HV and LV DC links, emerged as the most feasible and practical option. This topology enhances the ride-through capability of SSTs and improves power quality at both input and output ends. It provides all the desired functionalities of an SST and simplifies control design. The LVDC link enables the integration of distributed energy resources and energy storage systems. As shown in Figure 3.4, the structural design includes a unique rectifier, an HF connected DC-DC converter, and an inverter. This structure enables notable advantages, including flexibility, control performance, and practicality, making it the most viable solution. The most practical and effective solution for an SST is a three-stage configuration with two DC links, offering exceptional flexibility and control performance. This design surpasses the functionalities of traditional transformers.

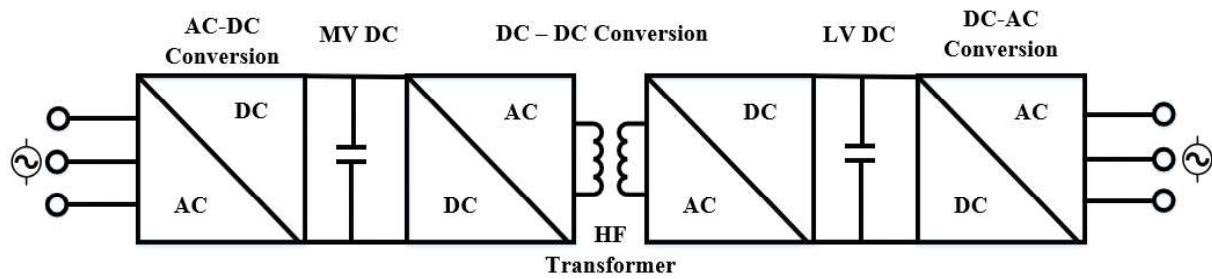


Figure 3.4. Single-stage SST structure

This arrangement creates a medium voltage direct current (MVDC) link by first converting the AC voltage at the input into a similar DC voltage. The next step is to convert this MVDC into high-frequency AC voltage, which is then transferred via a medium-frequency transformer. In this phase, the voltage level is decreased, rectified once again to a low-voltage DC level, and a low-voltage DC link is created. The next step involves further transforming the resulting low-voltage direct current (LVDC) to create a 50 Hz AC voltage. The SST system's many functions can be integrated through the use of this sequential voltage conversion and transformation process, which also ensures the desired power flow.

3.1.3.1. HF Transformer

A high-frequency transformer is a type of power transformer designed to operate at frequencies higher than the medium frequency range (typically above 10kHz). It is primarily utilized in high-frequency switching power supplies, high-frequency inverters, and high-frequency welding machines.

These transformers are classified based on their working frequencies into several categories, such as 10kHz-50kHz, 50kHz-100kHz, 100kHz-500kHz, 500kHz-1MHz, and 10MHz or higher. High-frequency switching power supplies use high-frequency transformers to convert AC voltage, and they consist of windings, cores, and bushings that enable efficient voltage transformation at these elevated frequencies. The high-frequency transformer plays a vital role in switching power supplies, which utilize various circuit topologies. One example is the semi-bridge power conversion circuit, where two switching transistors alternate to generate a high-frequency pulse wave around 100kHz. This pulse wave is then transformed into AC output

through the high-frequency transformer. The output voltage is determined by the turn ratio of each winding in the transformer. The circuit of high-frequency transformer is shown in Figure 3.5.

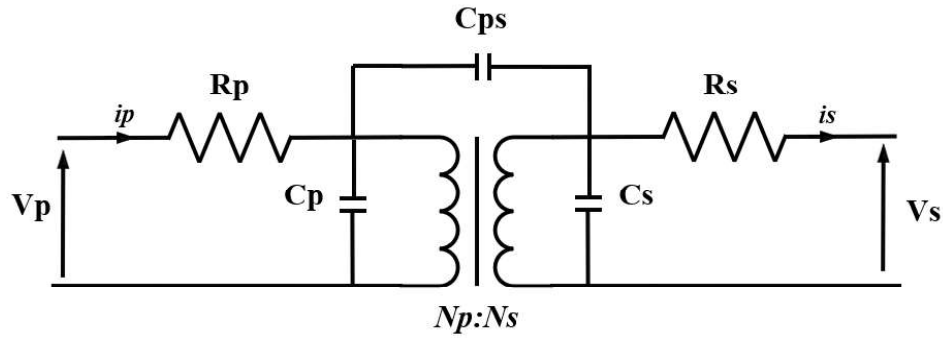


Figure 3.5. The circuit of the high-frequency transformer

In a typical half-bridge transformer circuit, three high-frequency transformers are prominent: the main transformer, transformer, and auxiliary transformer (also known as a standby transformer). Each transformer has specific requirements based on its function. For instance, the main transformer used in power supplies above 200W should have a core diameter (or level) of at least 35mm. On the other hand, the auxiliary transformer for power supplies up to 300W can suffice with a core diameter of 16mm.

A transformer is a device that converts AC voltage, current, and impedance. When an AC flows through the primary coil, it generates an AC magnetic flux in the iron core or magnetic core, inducing voltage or current in the secondary coil. A transformer comprises an iron core or magnetic core and multiple coils, known as windings. In the design of high-frequency transformers for switching power supplies, minimizing leakage inductance and distributed capacitance is crucial. These factors can cause undesirable effects during the transient transmission of high-frequency pulse square wave signals, such as surge currents, peak voltages, and oscillations, leading to increased losses. To address these issues, it is important to control the leakage inductance of the transformer, aiming for it to be around 1% to 3% of the primary inductance. Leakage inductance occurs due to incomplete magnetic flux coupling between the primary and secondary coils, between layers, and between turns of the transformer.

Distributed capacitance, on the other hand, refers to the capacitance between turns of the transformer winding, between upper and lower layers of the same winding, between different windings, and between the windings and the shielding layer. Minimizing distributed capacitance is essential to prevent unwanted coupling and disturbances. In transformer winding placement, several considerations come into play. The primary winding should be positioned in the innermost layer to ensure the shortest length of each turn, reducing the overall wire length and effectively reducing distributed capacitance. For the secondary winding, it is recommended to add (3-5) layers of insulation backing after winding the primary winding. This approach helps reduce capacitance between the primary and secondary windings while improving insulation strength.

The placement of the bias winding, whether between the primary and secondary windings or in the outermost layer, depends on whether the switching supply is adjusted based on the secondary or primary voltage. Overall, minimizing leakage inductance and distributed capacitance in high-frequency transformers is essential for the efficient and reliable operation of switching power supplies, mitigating unwanted effects, and optimizing performance. The high-frequency (HF) transformer plays a pivotal role in the structure of the SST, offering significant advantages over conventional low-frequency transformers. Its compact size and reduced weight are a direct result of the high-frequency magnetic field, which requires smaller winding dimensions. This miniaturized transformer design eliminates any physical contact between the primary and secondary sides of the converter, leading to improved isolation strength.

The primary winding produces a magnetic wave while it is in operation, cutting across the secondary winding to produce the necessary high-frequency voltage level. This voltage is then transformed into LF using a different converter. Although the HF transformer operates at higher frequencies, its fundamental operation is identical to that of a normal transformer.

The core material, wire choice, and winding arrangement are three crucial design factors that must be taken into account to maximize the HF transformer's performance. To reduce core losses and prevent core saturation under typical operating conditions, careful core material

selection is essential. Size, power density, and power losses of the transformer are all directly impacted by the choice of core material.

Additionally, wire selection plays a vital role in achieving efficient operation. The chosen wire must have suitable electrical and thermal properties to handle high-frequency currents and minimize losses. Proper winding placement is also essential to ensure optimal magnetic coupling and efficient energy transfer. By addressing these design considerations, the HF transformer can be optimized for superior performance, enabling the SST to deliver its intended functionality with improved size, power density, and efficiency.

Flux density and power loss trade-offs offered by various core materials differ. High flux density is provided by silicon steel, however it also loses a lot of power. Although soft ferrite requires a bigger core size, it has a moderate power loss. Despite having a good flux density, amorphous material loses a lot of power when used in high-power applications. Despite being more expensive, nanocrystalline material is the best core material for power density and efficiency because to its balance of average flux density and the lowest power loss. Amorphous material is the second option, however it is not recommended for high-power applications.

Frequency management is taken into consideration to reduce power loss in the core material. The skin effect must be taken into account at high frequencies because the skin depth gets smaller than the wire diameter. Litz wire, which divides the conductor into smaller strands with enough shielding, is selected as a remedy to lessen the effect of the skin effect. Better high-frequency performance is produced by thinner wire strands. Currently, 50 American wire gauge (AWG) is the thinnest Litz wire strand that is offered on the market.

The geometry of the windings, such as solenoidal and coaxial winding, affects the design of high-frequency transformers as well. Coaxial winding has controllability advantages, although solenoidal constructions are advised since they are more affordable and simpler to produce. There are numerous solenoidal transformer structures, including matrix-, shell-, and core-types. Greater insulation between the main and secondary sides is provided by core-type, better

mechanical protection is provided by shell-type, and inter-wiring characteristics that lower leakage inductance and winding AC resistance are possible with matrix-type with many cores.

Important design factors affecting power loss and system effectiveness include the distribution of wire turns and where the windings are placed on the core. They affect the windings' coupling coefficients and mutual inductances. It is crucial to properly take these considerations into account.

Isolation between primary and secondary voltage zones is a key challenge in high-frequency transformer design. Additionally, using multiple power electronics converters for grid integration can increase cost, size, failure rate, and control complexity. To address these issues, a multi-port solid-state transformer (MPSST) can be employed. The MPSST utilizes a multi-winding high-frequency transformer to create an adjustable, modular, isolated, and compact converter that integrates multiple elements in the power grid. By combining the functionality of several converters in one, the number of components is reduced, and the power density in the system is increased. The MPSST acts as a centralized controller, managing power flow and regulating voltage [60].

3.1.3.2. Rectifier (AC-DC)

Rectifiers are versatile components used in power management for a range of applications, including voltage regulation, power conditioning, and switching. They are vital for providing power to electronic devices like computers and televisions, as well as for motors and welding. The main function of a rectifier is to control voltage in a circuit, maintain a constant output power, and enable switching between different load sources. Additionally, rectifiers can combine low-ripple pulse waveforms into higher voltage waveforms. They offer a simple and efficient solution for managing energy requirements in diverse scenarios.

A rectifier utilizes a diode, which allows electric current to pass in one direction (forward bias) but blocks it in the opposite direction (reverse bias). This unique property of diodes enables them to function as rectifiers. Rectifiers consist of various components, including diodes, capacitors, and resistors, working together to convert AC power into DC power effectively.

Depending on the specific requirements, different types of rectifiers are available, such as half-wave, full-wave, bridge rectifiers, and high-frequency switching rectifiers. The design of a rectifier is crucial, as its electrical parameters impact the performance of the power management system.

Essentially, a rectifier is an electrical device that converts AC power to DC power. It achieves this by using a diode bridge to convert the AC into pulsating DC, which is then smoothed out by a filter. Rectifiers play a fundamental role in power management systems, such as computer hardware and power supplies. They are classified as single-phase or three-phase based on their circuit design and the number of output terminals. Understanding rectifiers, their design, and classification is crucial for selecting the appropriate components to ensure effective power management.

Rectifiers are designed to meet specific electrical requirements, including current and voltage levels, to ensure optimal performance. They are essential for converting AC power into DC power, which is often necessary for powering various applications. AC power refers to alternating current, where the current flow changes direction periodically, while DC power is direct current, with electricity flowing in one fixed direction only. Rectifiers bridge the gap between AC and DC power, making it usable for specific applications. They are widely used in devices like computers, televisions, medical equipment, and charging systems. Consulting a comprehensive guide on rectifiers can provide further insights into their uses and how they contribute to efficient power management.

Rectifiers enable more efficient energy usage by regulating the current flow. They are crucial power management tools that convert AC power to DC power, with different types catering to specific applications. Single-phase and three-phase rectifiers are commonly employed in the power industry due to their ability to control output current and provide smoother current flow. Other types, like four-quadrant rectifiers, offer more complex configurations, allowing for both positive and negative outputs. Understanding the distinctions between rectifier types and their applications is essential for achieving maximum efficiency and reliable performance in power management.

In a multi-level SST control system, the rectifier stage, which converts AC to DC power, is regulated by monitoring the input voltages, currents, and HV DC capacitor voltages. This ensures that the desired power factor is maintained at the input terminal of the second-stage converter. The second-stage conversion is controlled by comparing the sensed output voltage with the desired voltage, generating control signals to achieve the desired output at the input terminal of the three-stage converter. In the three-stage AC-DC conversion, the controller only senses the voltage and current at the output side to generate the necessary control signals for the converter.

Different topologies have been discussed for the HV rectifier side, including Neutral Point Clamped (NPC), Cascaded H-Bridge (CHB), and MMC. The NPC topology uses waveform sensing from the input and output to generate a Pulse Width Modulation (PWM) control signal. CHB is controlled using a single-phase vector control method. On the other hand, MMC requires additional measurements compared to CHB and NPC, such as cell capacitor voltages, DC, arm current, AC currents, and AC voltages. Consequently, the MMC topology is more expensive and has higher control complexity.

3.1.3.3. Converter (DC-DC)

The dc-dc conversion stage is a crucial component of three-level SSTs since the effectiveness of the entire power system is typically dependent on the efficiency of the direct current conversion part. HF power converters are crucial parts for prospective smart-grid applications because they can function as a DC transformer. The SST's dc-dc conversion stage poses the most difficult and noticeable challenges since it must adhere to a number of stringent specifications in order to achieve effective conversion and durable isolation. Quick switching rates, zero voltage and current switching, zero loads current, smallest loss at forceful turn-on, low di/dt characteristic, HF isolation, power density, sensitive switching, and substantial power are a few of these needs. Researchers use numerous variable converter topologies in applications while supplying specific requirements. These converters include modular multilevel converters, resonant converters, dual active bridge converters, and converters with four active bridges.

The modular multilevel converter structure employed in the SSTs is shown in Figure 3.6. MMCs provide modular structure design, making it simple to raise the power ratings of SSTs. There is redundancy offered by this converter type. This indicates that a substitute for electrical failures is an additional converter cell. Consequently, the power electronic-based system's dependability is secured. Utilizing modular converters also results in a reduction in the system's filtering process due to their different output voltages. The SST system's electrical losses and product costs are reduced because of the modular structure's low dv/dt and ability to use LV power switches. The MMC's voltages are challenging to manage, though. The output capacitor voltages must be balanced, which calls for additional control techniques. Voltages must be continuously checked to control capacitor voltages. Due to the circulating power and back-to-back power that are caused by the double-frequency components, the power switch losses have increased.

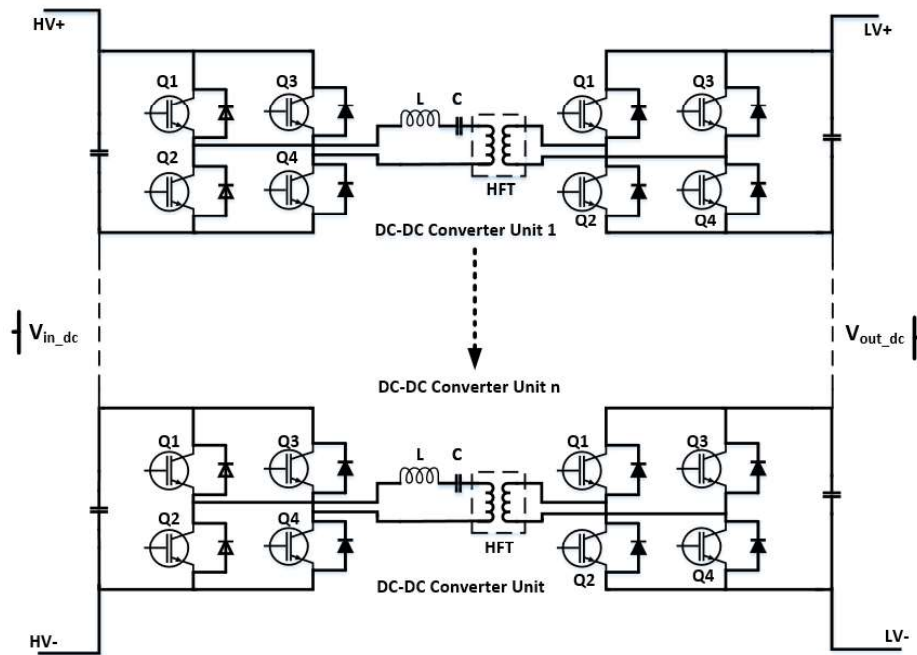


Figure 3.6. MMC converter structure used in SSTs

The half-cycle discontinuous-conduction-mode series resonant converter (HC-DCM SRC) is a type of SRC that can sustain gentle switching over a large load range and regulate output voltage via frequency control. Figure 3.7 shows the suggested SRC converter topology. The switches

can operate using soft switching thanks to the HC-DCM SRC topology. By doing so, switching losses are reduced and total system efficiency is raised. For applications requiring high output current and low output dc voltage, the SR technique is suggested for LLC converters. The technique revolves around the resonant frequency to get the best efficiency and voltage regulation. When soft switching is present, the SRC in sub-resonant mode performs well. To manage the output power at a light load, active control techniques must be added. On the power switches, there is also a lot of energy flowing about. The main issues with SRC converters are the low load regulation, increased turn-off current, and high circulating energy.

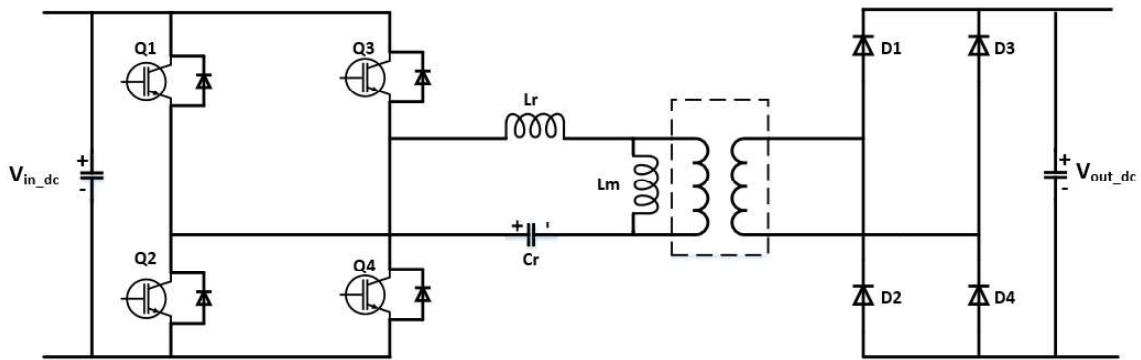


Figure 3.7. The full-bridge LLC-based SRC converter topology for SSTs

Since they can serve as a DC transformer, MF direct current power converters are essential components for upcoming smart-grid implementations. To achieve high conversion efficiency and galvanic isolation, the SST's dc-dc conversion stage must adhere to some stringent specifications. Some of these requirements include high-frequency isolation, quick switching rates, zero voltage and current switching, zero load current, minimal loss of powerful turn-on, low di/dt characteristics, sensitive switching, and power density. In applications, researchers employ a variety of variable converter topologies while fulfilling certain requirements. Among these converters are resonant converters, dual active bridge converters, converters with four active bridges, and modular multilevel converters.

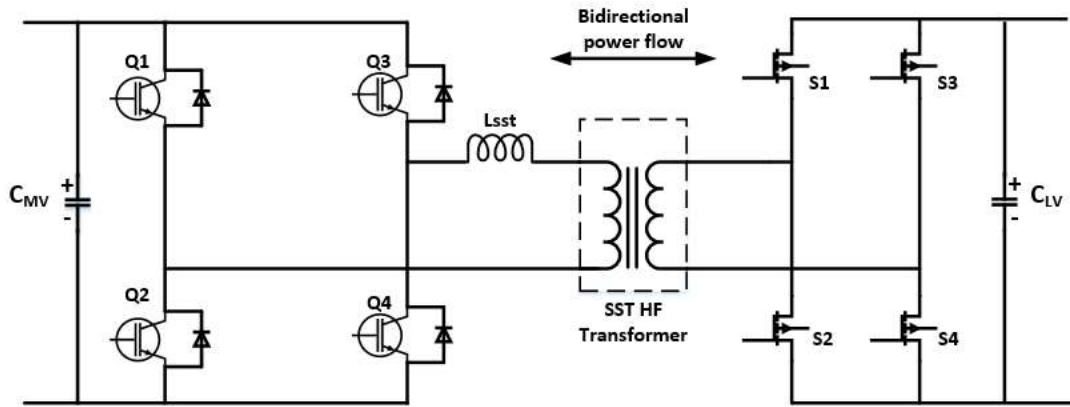


Figure 3.8. Topology for an isolated DAB converter for SST

In terms of price and efficiency, a QAB dc-dc converter is more common. owing to the converter's modular design, which results in a lower number of cells and high-frequency transformers. Additionally, this converter's cross-connection arrangement allows for direct power transfer from the HF transformer to the inverter. The fundamental drawback of QAB converters is that the power flow is entangled among the nodes, and changes in one node's power have an impact on all the other nodes. To control the power flow, a better control technique is required. Figure 3.9 shows the QAB modular active bridge dc-dc converter topology. Also,an overview of the direct current power converters utilized in SSTs is shown in Table 3.1.

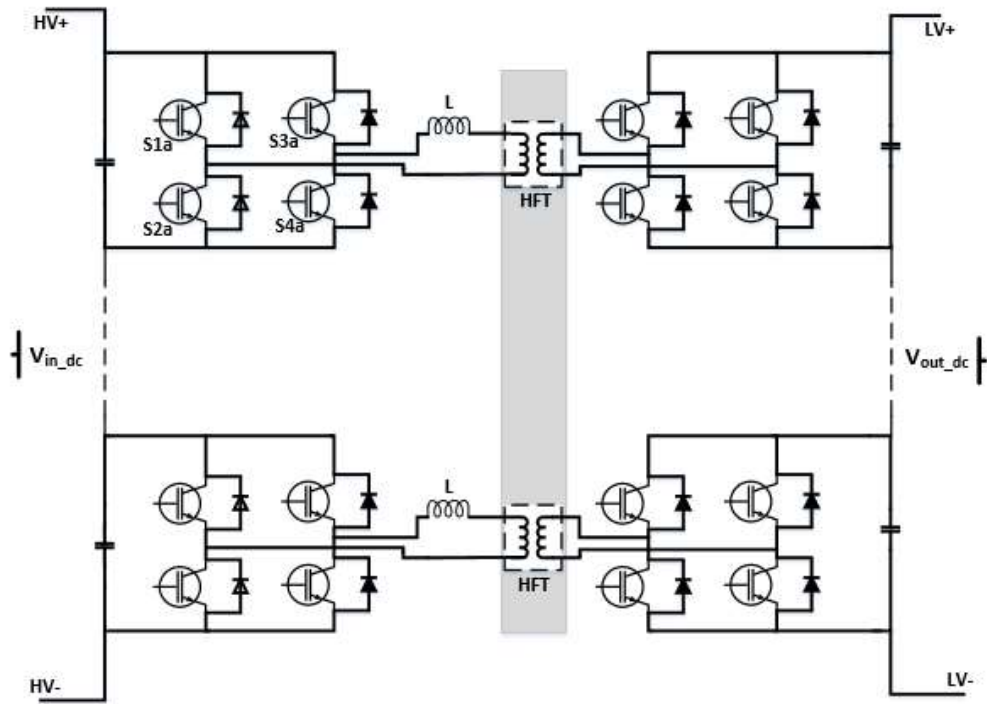


Figure 3.9. QAB modular active bridge dc-dc converter topology for SST

Table 3.1. Dc-dc converter used in SSTs

Converter	dv/dt at turn-on of the main switch	Resonant Circuit Connection	Bidirectional Power Flow Speed	Circulating Current	Soft-Switching Range	Multiple DC Cable Connections
MC	<10kv/us	Resonant Tank	>1us	High	>100ns, <600 ns	✓
SRC	>10kv/us, <100kv/us	Series Resonant Tank	<1us	Medium	>100ns, <100 us	x
DAB	<10kv/us	LLC Resonant Tank	<1us	Low	>100ns, <100 us	✓
QAB	<10kv/us	No	>1us	High	>100ns, <600 ns	✓

3.1.3.4. Inverter (DC-AC)

An inverter is a compact electrical device that converts DC voltage into AC voltage, which is commonly used in various household appliances. It finds application in equipment such as solar power systems and power batteries that rely on DC. The design of an inverter involves different power circuit configurations and voltage control methods. One crucial aspect of inverter technology is the output waveform, which can be a square wave, quasi-sine wave, or sine wave. Capacitors and inductors are employed to filter the waveform, while low-pass filters are utilized to minimize harmonic components. In the case of fixed output frequency, resonant filters can be used, while for inverters with adjustable output frequency, the filter needs to be tuned above the maximum fundamental frequency. Feedback rectifiers are employed to manage the peak inductive load current when the switch turns off.

According to Fourier analysis, a square wave exhibits odd harmonics (e.g., third, fifth, seventh) only if it is symmetrical around the 180-degree point. However, if the square wave has steps of specific widths and heights, these additional harmonics can be canceled out. By introducing a zero voltage step between the positive and negative parts of the square wave, the harmonics divisible by three can be eliminated. To achieve this, the pulse width of each positive and negative step should be one-third of the period, while the pulse width of each zero voltage step should be one-sixth of the period. This arrangement eliminates the fifth, seventh, eleventh, and thirteenth harmonics, and so on. PWM technology allows for altering the characteristics of the square wave by modulating and regulating the switching pulses before supplying them to the load. When the inverter does not require voltage control, a fixed pulse width can be utilized.

The Single Pulse Width Modulation (SPWM) technique utilizes a square wave reference signal and a triangular wave carrier signal. The gate pulse, which controls the technique, is generated by comparing these two signals for each half cycle. However, a major drawback of SPWM is the presence of higher harmonics in the output. To overcome this drawback, Multiple Pulse Width Modulation (MPWM) technique is employed. Instead of a single pulse, multiple pulses are used for each half-cycle of the output voltage. The output frequency is regulated by adjusting the carrier frequency.

In Sinusoidal Pulse Width Modulation, a sine wave is utilized as the reference signal, while a triangular wave serves as the carrier. The output waveform becomes a sine wave, and the RMS voltage value of the sine wave is controlled by the modulation index.

In Modified Sinusoidal Pulse Width Modulation, the carrier wave is applied only during the first and last sixty-degree intervals of each half cycle. This modification is implemented to improve the harmonic characteristics by reducing switching losses and enhancing the fundamental component of the output waveform.

3.2. Electrical and Control Components used in Proposed SST

3.2.1. IGBT Selection for Converters

In the experimental studies for connecting renewable energy systems to the proposed dc-dc converter, a 600 V, 50 A IGBT (Insulated Gate Bipolar Transistor) was chosen. The current and voltage values of the DC-DC converter are crucial for activating the inverter of the SST (presumably a related system). IGBTs are commonly used in amplifier circuits due to their fast switching capability, enabling them to turn on and off rapidly. They are especially useful in synthesizing complex waveforms using the PWM technique and low-pass filters. The selected IGBT, with its high switching speed capability, voltage control, and low inductance module structure, was deemed suitable for the desired application. Detailed characteristics of the chosen IGBT can be found in Table 3.2.

Table 3.2. Features of the proposed IGBT used in the SST

Definition	Features	Definition	Features
Mounting type	Gövde	Temperature (IC)	100°C
Package / Case	34x94Case	Voltage	600V
Technology	V SERIES	Operating Temperature	-40°C ~ +125°C
Diode	Internal	Tj °C	175°C
Ic 25°C	50A	VCE _{sat}	1.9V
Flow	45A	Power Dissipation	555W
Rohs	5.6 Ω	Configuration	Odd

3.2.2. Control Card

The DSP control card experiment kit, specifically the C2000 DIMM100 Experimenter's Kit, was utilized for various tasks such as voltage and current control, soft switching, zero switching, and gate control of power switches and DC-DC systems. This kit provides a convenient and cost-effective way to evaluate the performance of TMS320F28x devices. It comprises a docking station and an F28x control board. The docking station acts as a base board, accepting any DIMM100 add-on control card and allowing access to important GPIO and analog-to-digital converter (ADC) signals of the F28x device. It also includes two prototyping areas for wire wrapping and soldering. The DSP TMS320f28335 control board enables PWM control of the switching elements. The PWM control is facilitated through the TMDSDOCK28335 program, which is defined using Matlab/Simulink. The TMDSDOCK28335 is an evaluation and development tool for the C2000 Delfino F2833x series microcontroller products, and the docking station powers the control board while providing a breadboard area for prototyping. The controller board integrates with Matlab/Simulink, allowing the designed controller blocks

to be loaded onto the board, and the code is automatically generated. The controller board is illustrated in Figure 3.10

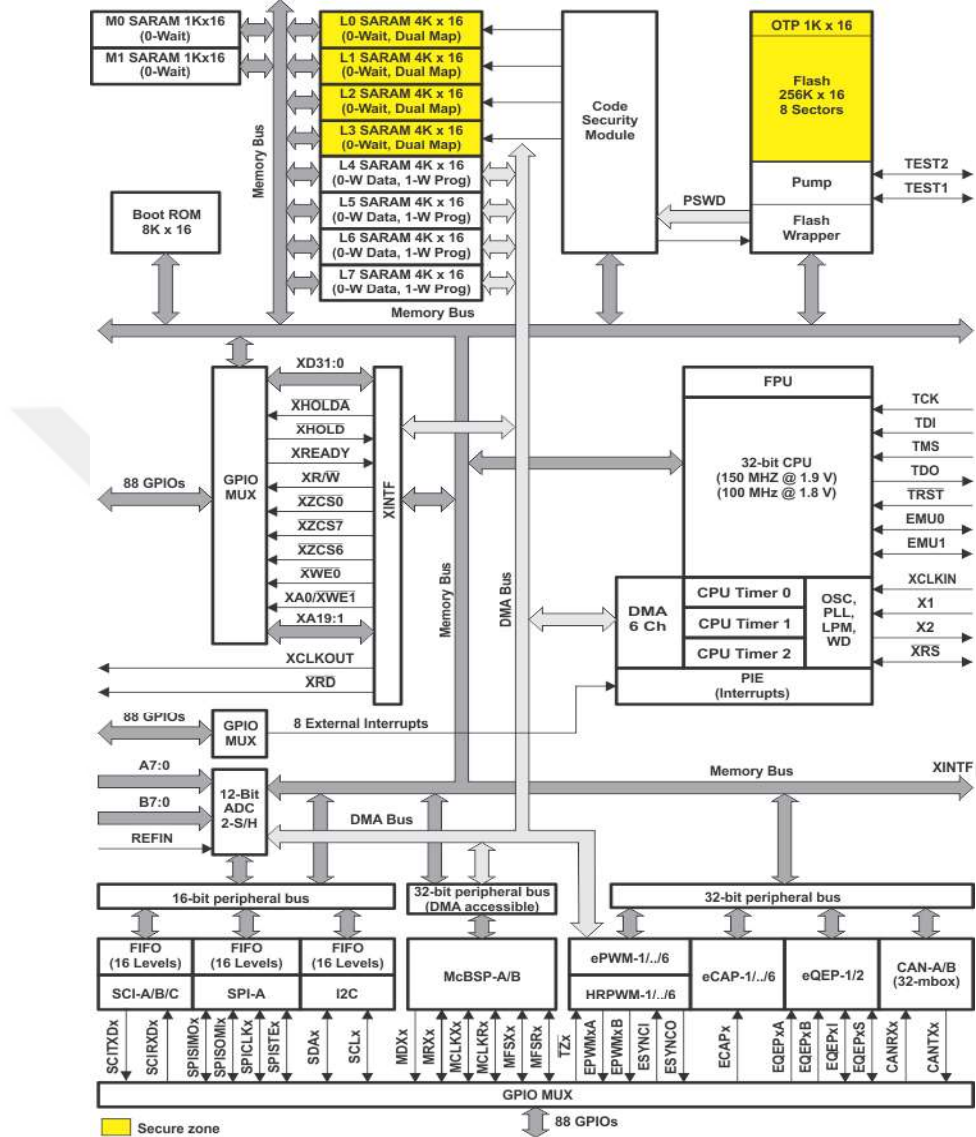


Figure 3.10. TMS320F28335 function block diagram [61]

3.2.3. IGBT Gate Driver

Gate drivers with a high-frequency operating range have been selected for the IGBT models used in the system. A gate driver serves as a power amplifier that takes low-power input from a controller IC and generates a high-current gate drive signal for the power devices. Gate drivers play a crucial role in power electronics systems as they drive power semiconductor devices

such as transistors, FETs, BJTs, MOSFETs, and IGBTs, which are widely used in applications like switched-mode power supplies, universal power supplies, and motor drives.

The switching behavior of power devices, such as MOSFETs and IGBTs, can be influenced by parasitic capacitances between their non-linear terminals (gate-to-source, gate-to-drain, and drain-to-source). To turn these devices on, a positive voltage must be applied to the gate relative to the source/emitter. The gate voltage needs to exceed the gate threshold voltage (V_{th}) to establish conduction between the source and drain regions. In high-power applications, the gate of a power switch cannot be driven directly by a logic IC (PWM controller) output due to the low current capabilities of these outputs. Special gate drivers are required to provide the necessary voltage and current to charge the gate capacitance of the power devices efficiently.

Gate drivers are essential for power electronics systems, acting as an interface between high-power electronics and control circuitry. They ensure proper operation and protection of power semiconductor devices. The design of gate driver circuits is critical for power electronics converters, as their performance directly affects the output of DC-DC converters or SMPS. Additionally, gate drivers often require electrical isolation for functional and safety purposes, as mandated by regulatory and safety certification bodies. This isolation prevents shock hazards, protects low-voltage electronics from high-power side faults, and allows for different ground potentials between circuits. Inductive, capacitive, or optical methods are used for signal and power transmission between isolated circuits while maintaining electrical separation.

In certain applications that demand high power density and efficiency, such as converters, an isolated gate drive circuit is necessary. This is particularly evident in power converter topologies like half-bridge, full-bridge, buck, two-switch forward, and active clamp-forward. In these topologies, both high and low switches are present because the low-side drivers cannot directly drive the upper-power device. This is because the upper power devices require an isolated gate driver since the source and emitter of these devices do not sit at ground potential.

Isolated gate drivers can be implemented using two popular techniques: magnetic isolation using gate driver transformers and optical isolation using optocouplers. In our upcoming

studies, we will focus on the magnetic isolation technique, specifically exploring the features and design requirements of gate-driven transformers.

In this thesis, the HCPL-3120 IC is employed to activate the IGBT power switches. The HCPL-3120 gate driver optocoupler consists of a GaAsP LED that is optically coupled to an integrated circuit with a power output stage. This configuration is ideal for turn on/off power IGBTs and MOSFETs utilized in motor control and inverter applications. The output stage's high operating voltage range provides the necessary drive voltages for gate-controlled devices. The voltage and current capabilities make the HCPL-3120 suitable for turn on/off power IGBTs increased to 600 V/50 A. For higher-rated IGBTs, the HCPL-3120 series can be employed to drive a separate power stage responsible for driving the IGBT gate. The HCPL-3120 has a peak insulation voltage of 630 V. The application circuit depicting the usage of HCPL-3120 to drive the IGBT is illustrated in Figure 3.11

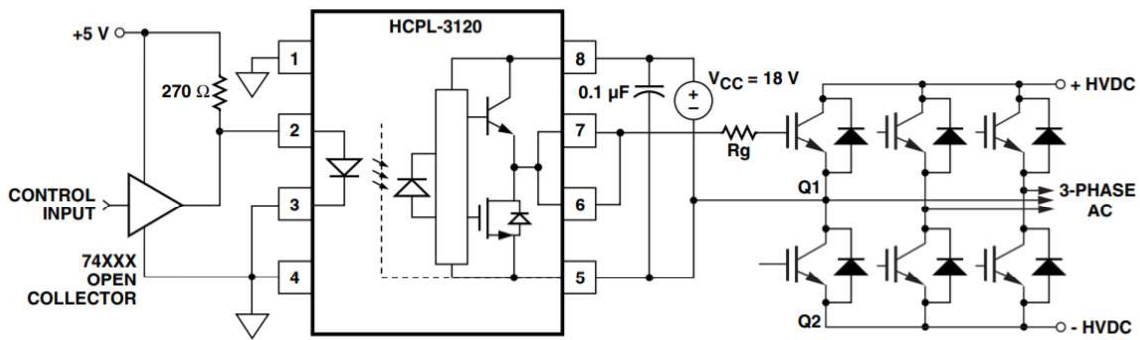


Figure 3.11. Application circuit of the HCPL-3120 to drive the IGBT [62]

The driver integrated with the additional design offers several advantages, including:

- The maximum peak output current of 2.5 A and minimum peak output current of 2.0 A, provides sufficient current capability for driving power devices.
- Minimum Common Mode Rejection of 25 kV/μs at $V_{CM} = 1500$ V, ensuring the effective rejection of common-mode noise.
- Maximum low-level output voltage (V_{OL}) of 0.5 V, eliminating the need for a negative gate driver by tightly keeping the IGBT closed.
- Maximum supply current (I_{CC}) of 5 mA, ensuring low power consumption.

- Hysteresis with Low Voltage Lockout protection offers additional protection and ensures reliable operation.
- Wide operating VCC range of 15 to 30 Volts, providing flexibility in power supply options.
- Maximum switching speeds of 500 ns, enabling fast and efficient switching of power devices.
- Industrial temperature range of -40°C to 100°C, ensuring reliable operation in various environments.

The design of the gate driver eliminates the need for a negative IGBT gate driver by achieving a very low VOL specification of 0.5V. This is accomplished through the use of a DMOS transistor with 1Ω resistance in the pull-down circuit. When the gate driver logic is in the low state, the IGBT gate is effectively shorted to the emitter through Rg + 1Ω. By minimizing the resistance and inductance in the connection from the HCPL-3120 to the IGBT gate and emitter, a negative IGBT gate driver is not required in many applications.

It is important to consider the PCB design to avoid routing the IGBT collector or emitter traces close to the gate driver's input, as this can lead to undesired coupling of transient signals and degrade performance. If the IGBT drain needs to be routed near the gate driver's input, it is recommended to keep the LED in the off state to prevent transient signals from the IGBT drain from affecting the gate driver's operation.

3.2.4. Gate Resistance Value Determination

To minimize IGBT switching losses, the following steps can be followed for choosing the gate resistor (R_g):

Step 1: Calculate the minimum R_g value based on the I_{OLPEAK} specification:

$$R_g \geq \frac{(V_{CC} - V_{EE} - V_{OL})}{I_{OLPEAK}} \quad (3.1)$$

Using the V_{OL} specification (2V) at 2.5A peak current, the minimum R_g value can be determined. It's important to note that the actual voltage supplied by the gate driver may not be

an ideal voltage step, resulting in lower peak currents than predicted. When a negative gate driver is not used, V_{EE} in the equation can be considered as zero volts.

Step 2: Check gate driver power consumption and increase R_g if necessary:

The total power consumption of the gate driver (P_T) is the sum of the emitter power (P_E) and output power (P_O). This calculation helps ensure that the power consumption is within acceptable limits. If the calculated P_O exceeds the maximum allowed P_O ($P_{O(MAX)}$), it indicates that the gate drive power consumption is too high, and R_g should be increased to reduce it.

$$P_T = P_E + P_O \quad (3.2)$$

$$P_E = I_F \cdot V_F \cdot DutyCycle \quad (3.3)$$

$$P_o = P_{O(BIAS)} + P_{O(SWITCHING)} \quad (3.4)$$

3.2.5. Mosfet Driver

A driver circuit has been designed using the IR2104 IC along with a 4N25 isolated optocoupler to drive high-power MOSFETs. The IR2104 is a power control IC that comes in a DIP-8 package. It provides voltage in the range of 10V to 20V and can deliver output currents of up to 210mA. The IR2104 IC is specifically designed for Half-Bridge applications.

The IR2104 IC has a fast input delay of 680ns and an output delay of 150ns, ensuring efficient and timely operation. It offers a range of integrated specifications, which are provided in Table 3.3. These specifications include various electrical parameters and performance characteristics of the IR2104 IC that are important for its proper functioning in the driver circuit.

Table 3.3. IR2104 integrated specifications used in the SST

Definition	Features
<i>Product Category</i>	<i>Power Control Integrations</i>
<i>Product Brand</i>	<i>IR</i>
<i>Mounting Type</i>	<i>THT</i>
<i>Package/Box</i>	<i>DIP-8</i>
<i>Input (Motor) (Gate) Type</i>	<i>Not Reversible</i>
<i>Time</i>	<i>680ns</i>
<i>Output Current</i>	<i>210mA</i>
<i>Supply voltage</i>	<i>10V - 20V</i>
<i>Operating temperature</i>	<i>-40°C ~ +125°C</i>

In a typical MOSFET driver circuit, the IR2104 IC and the 4N25 optocoupler are commonly used components. The circuit involves connecting the control signal source to the input side of the optocoupler. This input signal is then transmitted through the optocoupler, which provides electrical isolation, to the IR2104 IC. The IR2104 IC amplifies and shapes the input signal, generating the appropriate drive signals for the MOSFET. These drive signals are then connected to the gate of the MOSFET, enabling precise control of its switching behavior. The specific connection details and component placements can be referred to in Figure 3.12.

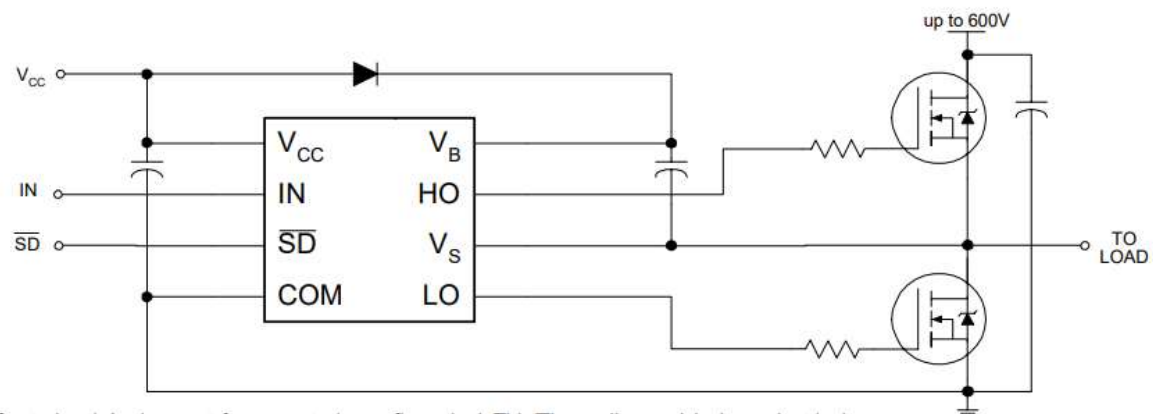


Figure 3.12. Application circuit of the IR2104 to drive the MOSFET [63]

3.3. Control Strategies for Converters

The control mechanisms used by SSTs in dc-dc converters have a significant impact on how well they perform. These converters frequently utilize variable control systems, with power control techniques being used to guarantee voltage balance and output current sharing in the SST. These control methods are essential for protecting the power switches from wear and tear. IGBT device premature failure is reduced by maintaining voltage balance, and power switch heat stress is reduced via current sharing management. Overcurrent and overvoltage dangers in power electronic systems can be reduced by using voltage and current management measures.

To enable ZVS and ZCS, frequency management is frequently applied in series resonant converter-integrated SSTs. This is accomplished by maintaining the switching frequency near to the resonant frequency. This minimizes the switching losses of the power switches. Reduced turn-off losses are achieved through trapezoidal modulation, which permits switch-off operation under ZVS conditions. When series resonant converter-based SSTs operate in discontinuous conduction mode, ZVS operation is achieved on the MV side and ZCS operation is achieved on the LV side. Due to the low di/dt capabilities of this control approach, EMI on the HF or MF transformers is reduced. The capacitor voltage balance-based Q2L management method improves active power transfer efficiency while shielding the HF transformer from EMI and dv/dt stress on the MF transformer.

For SSTs to convert power as efficiently as possible, duty cycle regulation is crucial. A large ZVS range, the lowest possible EMI, and a decrease in switching losses can be obtained by modifying the duty cycles while taking into account the harmonic circumstances on the MV and LV sides. Triangular current modulation is a useful method for achieving the wide range of soft-switching, especially when the output capacitor voltage is unstable since it helps stop reactive power from circulating. The high root-mean-square (RMS) value of the currents on the power switches and HF transformer as a result of its triangular design, however, may cause increased conduction losses. It is possible to independently manage the LV side voltage and D-axis current with D-Q control, which modifies the duty ratio to produce ZVS operation. This permits quick reaction from the power switches and overcurrent protection.

To control the flow of electricity and address current overload issues, phase shift modulation is widely used. Modifying the phase shift across bridges enables ZVS operation. However, to reduce peak currents and prevent abrupt phase shifts in the dc-dc converter, a second control strategy is needed. One such technique is the OLC, which controls the LV side voltage of the converter by raising the phase shift angle proportionately to the time step. For all dc-dc converters, PWM is a common control method. Power switches are controlled by PWM signals with a 50% duty ratio, providing control of power flow through on/off actions. The charging and discharging durations can be regulated, lowering circulating currents and limiting losses in SSTs, by regulating the MV-side and LV-side capacitor voltages using PWM.

In conclusion, phase shift modulation strategy are frequently used because of how simple they are to use and how well they can be controlled. However, handling overcurrent, overvoltage, and soft-switching activities frequently necessitates the use of additional control algorithms. These could be output voltage/current control techniques, single-phase shift control techniques using PI controllers, or extended-phase shift control techniques. For example, grid-connected solar PV systems use these additional algorithms to perform power balance control in dc-dc power converters.

4. RESULTS AND DISCUSSIONS

4.1. Simulation Studies for Proposed SST Application

This simulation study aims to analyze performance efficiency properties of a HF connected DAB converter-based SST. In order to improve grid dependability and power stability in a solar PV power plant connected grid, the study recommends utilizing a DAB-based SST. Figure 4.1 depicts an SST based on an on-grid PV solar system.

In the SST system, the DAB converter performs several functions. As a power electronics bridge for loads, additional power sources, and battery modules, it enables the construction of a DC microgrid. This is made possible by the device's capacity for bidirectional power flow, ZVS, and galvanic isolation. Since it allows for active control of the transferred power, the DAB converter provides advantages over conventional converters in situations where precise output voltage monitoring or current flow management are required.

Additionally, the DAB converter operates at resonance while under normal load conditions. This process ensures both ZVS turn-on at the primary side and ZCS turn-off at the secondary side. By increasing the converter's efficiency and reducing switching losses, this resonant action.

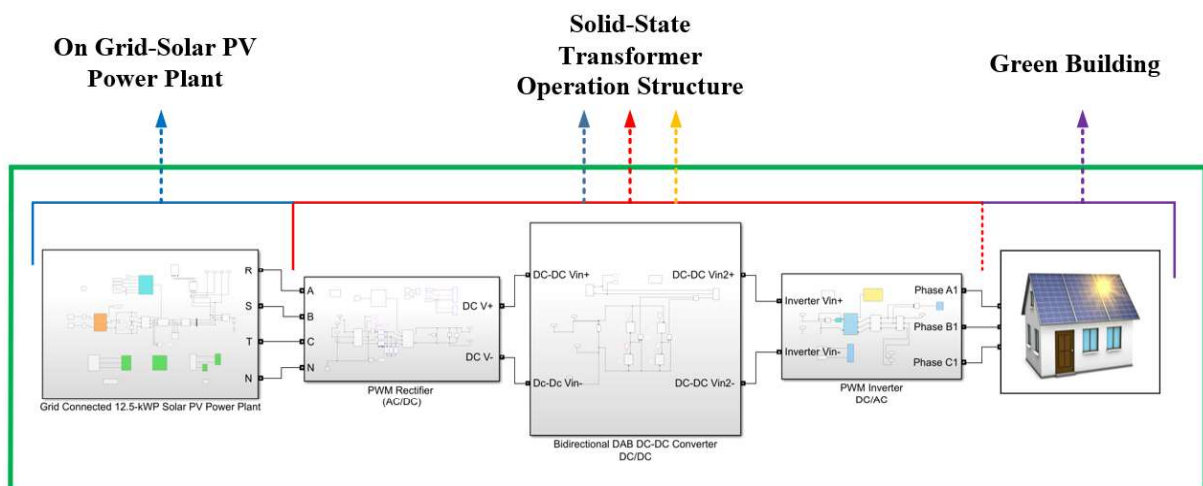


Figure 4.1. DAB converter-based three-level SST topology for smart grid

The solar PV power plant shown in Figure 4.2 represents a 12.5 kW grid-connected system designed to supply sustainable energy to a smart home. The power plant's capacity is specifically created to correspond to the typical home's hourly electricity use, which is approximately 10 kW.

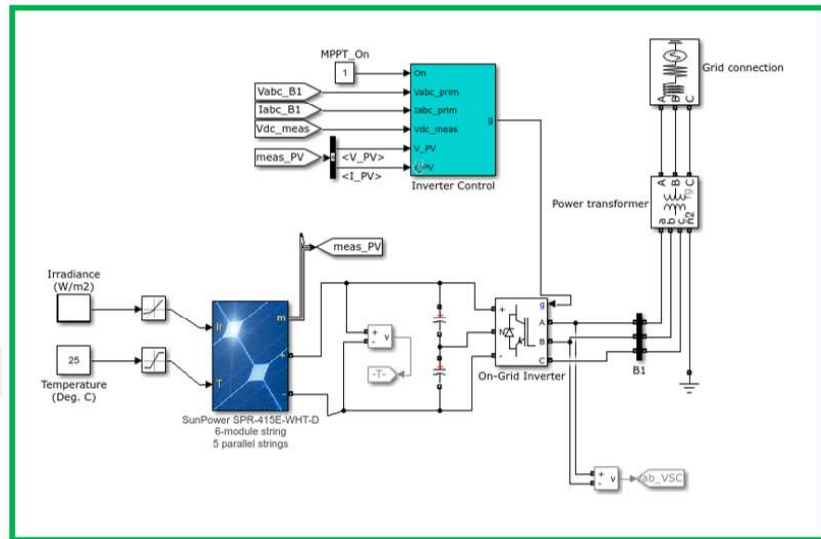


Figure 4.2. Real-time solar PV power system application

To optimize the performance of the solar PV system, a Maximum Power Point Tracking (MPPT) charge controller is employed. The MPPT controller utilizes the widely used "Perturb and Observe" technique to track and maintain the operating point at the maximum power point of the solar PV panels.

The MPPT controller continuously measures the voltage and current of the solar PV panels and calculates the instantaneous power output based on these measurements. By perturbing the operating voltage and observing the corresponding change in output power, the controller determines the correct direction of voltage change that leads to an increase in power output. This iterative process allows the controller to dynamically adjust the operating voltage to achieve the maximum power point, even in rapidly changing environmental conditions.

The proposed MPPT method aims to enhance the accuracy of power tracking and improve the system's dynamic performance. By accurately tracking the maximum power point, the solar PV

power plant can efficiently harness the available solar energy and provide a reliable and sustainable power supply to the connected smart home.

Figure 4.3 presents the proposed power tracking algorithm, which involves the calculation and comparison of output powers from the solar PV panels and the load side. The algorithm begins by comparing the voltages and then adjusting the reference voltage by a specified amount. This process continues until the maximum power point is reached, at which point the algorithm operates in a perturb and observe mode to maintain operation at that optimal point.

Renewable power plants often face challenges due to fluctuating weather conditions, which can lead to significant imbalances between the power produced and the power supplied to the grid. The proposed system addresses these issues, resulting in extended device lifetimes and increased profitability for smart homeowners. The simulation incorporates varying solar radiation levels and temperatures, allowing the assessment of power quality changes and the efficiency of the proposed solid-state transformer.

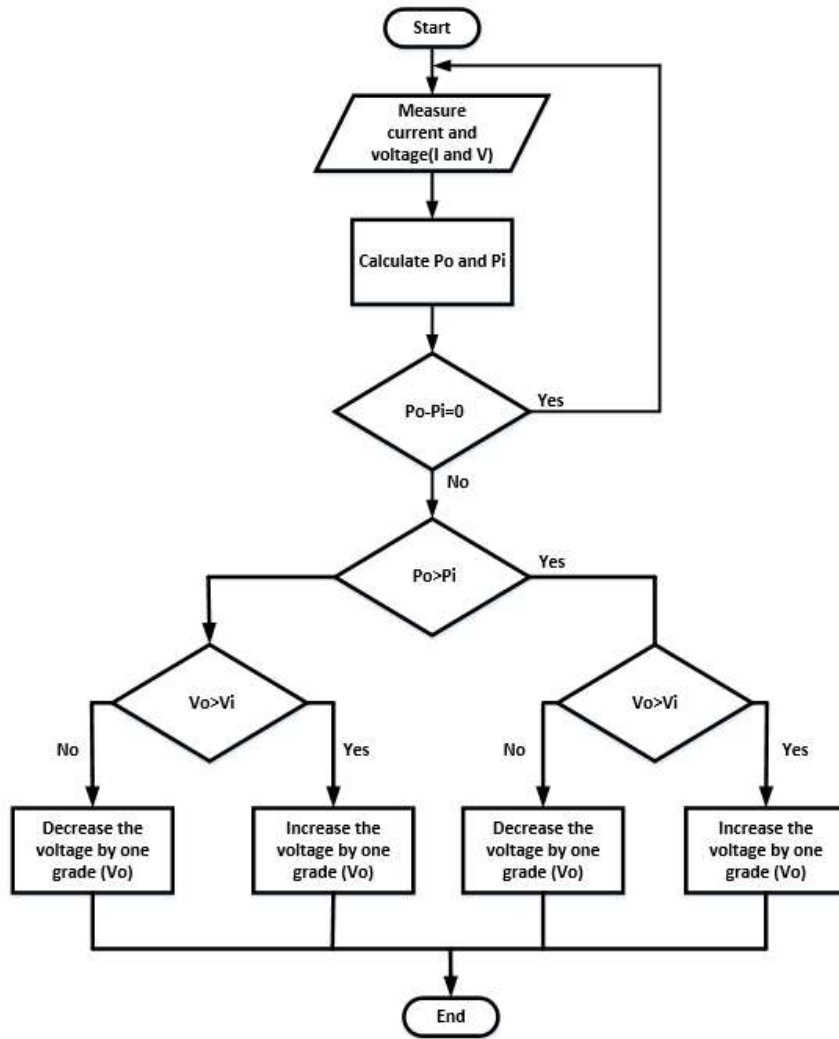


Figure 4.3. The flowchart of the proposed power tracking algorithm

Table 4.1 provides the parameter values for the solar PV panel utilized in the simulation, and Figure 4.4 displays the power changes that solar PV plants experience as a result of shifting weather conditions. The parallel string design of the solar PV array keeps the voltage constant. As solar radiation declines, current losses in solar cells alter power output proportionally.

Overall, the proposed power tracking algorithm and system design aim to optimize power generation and ensure efficient operation of the solar PV plant, even in the face of varying environmental conditions. This contributes to the stability and reliability of the smart home's power supply, maximizing the benefits for the homeowners.

Table 4.1. The solar PV panel electrical parameters

Electrical Characteristics	Value	Electrical Characteristics	Value
Power Rating	415 W	Power Density	192.13 W/m ²
Short Circuit Current (I_{sc})	6.21 A	Power Tolerances	0% - +5%
Open Circuit Voltage (V_{oc})	85.6 V	Peak Efficiency	19.66%
Maximum System Voltage	600 V	Nominal Operating Cell Temperature	45.8 °C

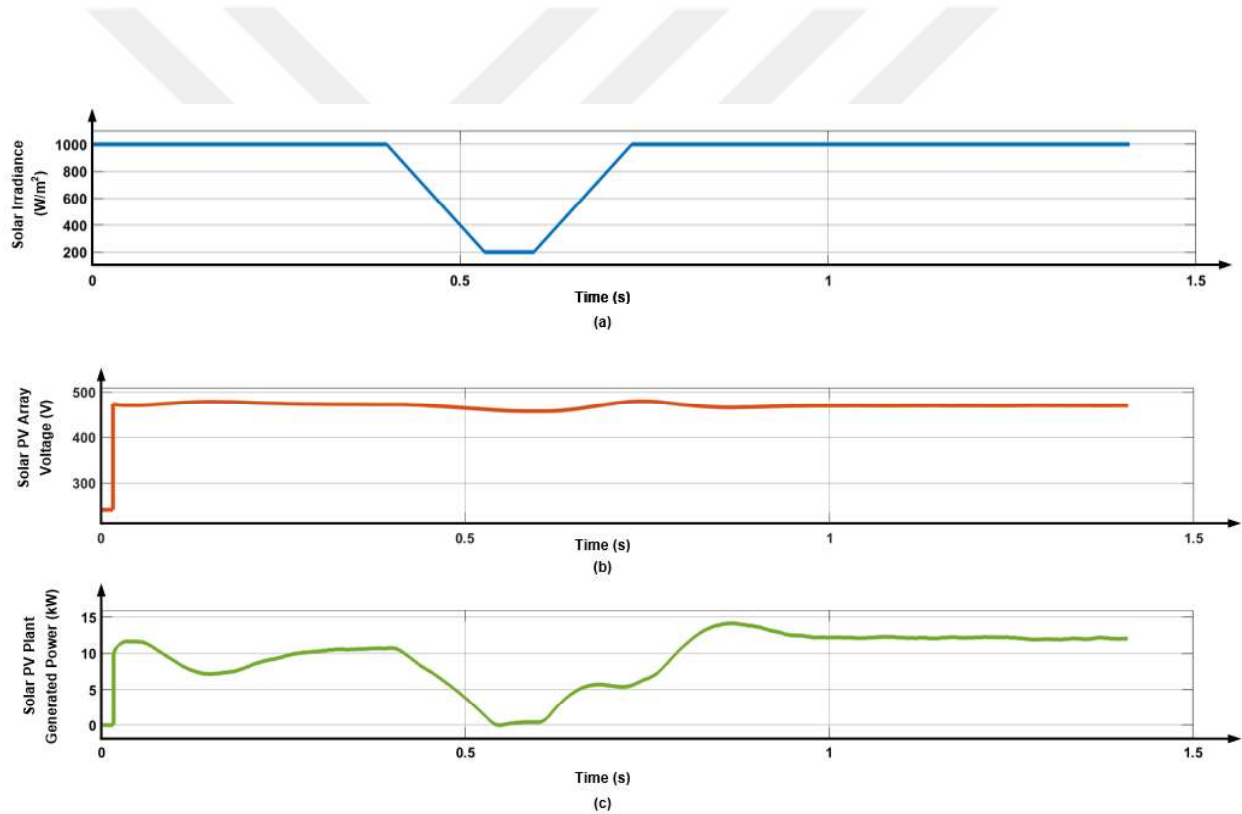


Figure 4.4. Solar PV power plant a) solar irradiance b) PV array voltage c) generated power

PWM rectifiers are highly favored for their ability to maintain high input power factors, keep line currents close to sinusoidal waveforms, and facilitate power injection back into the grid. Compared to uncontrolled rectifiers, controlled rectifiers are increasingly adopted due to their advantages, including the reduction of harmonic distortion sent to the grid and the capability

for regenerative operation. Among the various control strategies for controlled rectifiers, PWM stands out as an effective method.

The underlying principle of PWM control is to ensure that line currents reach desired values within fixed time intervals at a specific switching frequency. Command voltages are calculated to regulate the line currents, which are then converted into appropriate switching signals. By employing this switching scheme, power factor correction can be achieved, and the current waveforms can closely approximate sine waves. Implementation of PWM rectifiers often involves the use of IGBTs as the switching devices.

To convert 380V AC energy to a boosted 750V DC voltage, a modular PWM rectifier topology is selected, primarily due to the high voltage blocking capabilities required at MV levels. The detailed structure of the PWM converter can be observed in Figure 4.5. Through the three-phase PWM control system, variables such as load current, voltage, and power can be effectively controlled. Various types of controllers are employed for this purpose, including linear, nonlinear, robust, predictive, and intelligent controllers.

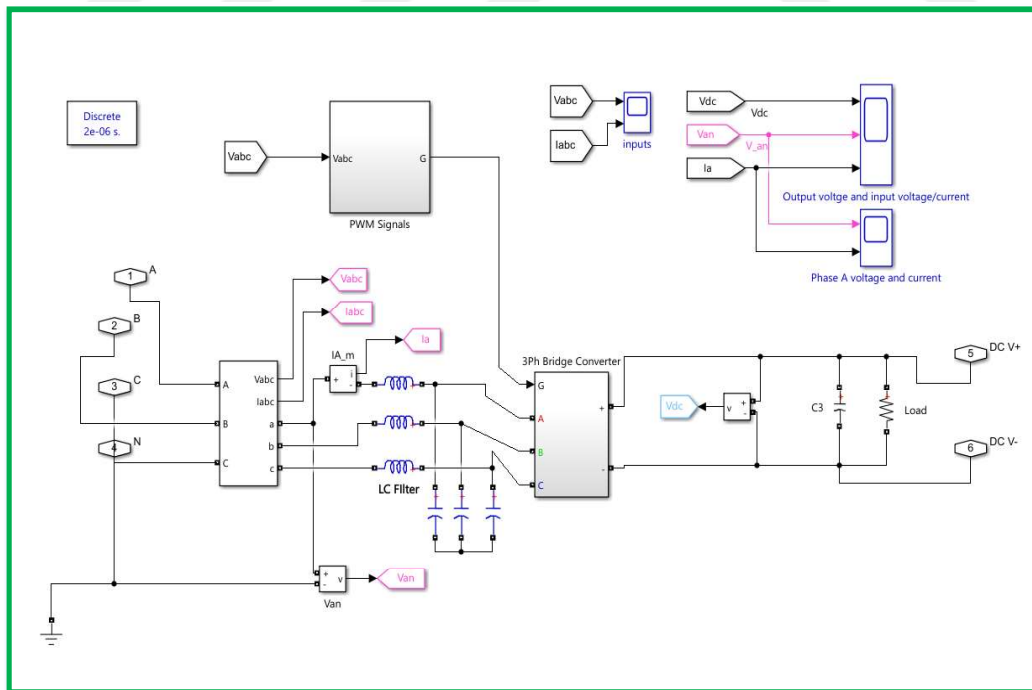


Figure 4.5. PWM rectifier for AC-DC stage of the SST

While controllers other than linear controllers may offer a superior dynamic response, they often introduce time delays into the system. Hence, linear controllers are generally preferred. In this study, the PI controller, which is widely recognized and utilized, was chosen as the linear controller for the PWM rectifier system.

The PI controller plays a crucial role in the PWM rectifier system by converting the error signal between the applied reference voltage and the output voltage into command currents, considering the switching delay. These command currents are then used to derive the corresponding command voltages, which are subsequently converted into appropriate switching signals. By imposing these switching signals, the line currents are compelled to track the command currents, ultimately resulting in a power factor of one. The determination of the switching functions for the system involves comparing the desired command voltages with a triangular waveform. To ensure system stability under unit negative feedback, the root square analysis is conducted, enabling the identification of suitable PI controller parameters.

Due to their low component count and superior controllability, dc-dc bidirectional DAB converters do exceptionally well when it comes to power conversion. Phase-shift modulation is used to enable both forward and backward power flow in the proposed DAB converter topology shown in Figure 4.6. The phase-shifted square waves' duty cycles can be adjusted to change the power flow direction.

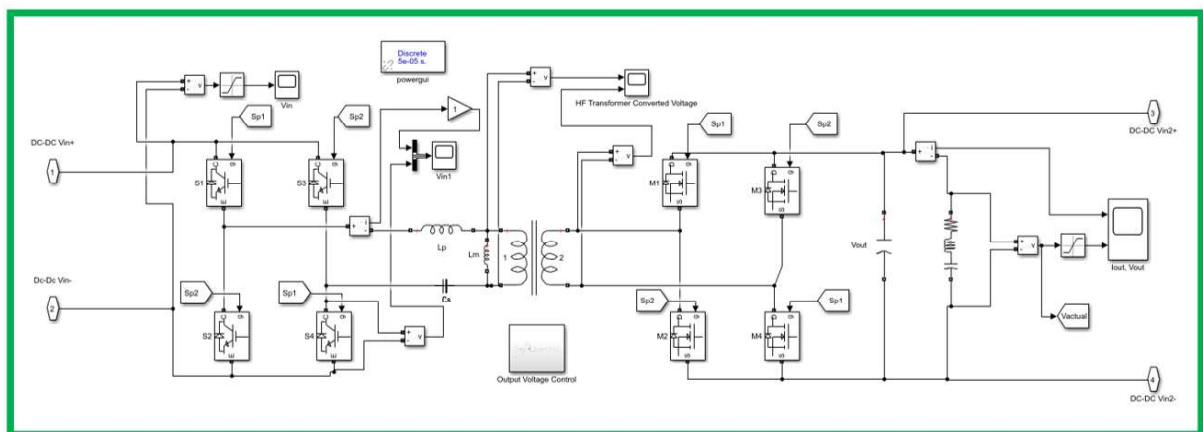


Figure 4.6. DAB converter for isolated DC-DC stage of the SST

The proposed converter incorporates two essential inductors: the primary leakage inductance (L_p) and the magnetizing inductance (L_m). L_p serves to mitigate sudden voltage spikes and minimize current harmonics. On the other hand, L_m plays a role in reducing electromagnetic interferences arising from circulating currents in the high-frequency transformer. Consequently, this design lowers power losses in the power switches and HF transformer, leading to improved converter efficiency.

With a DAB topology and great power conversion capabilities, the recommended dc-dc converter works well. The minimal component count and strong controllability are just two of its many benefits. Phase-shift modulation, which is used by the converter, allows for both forward and backward power flow. The power flow direction can be changed based on the operation needed by using phase-shifted duty square waves.

To ensure efficient operation and minimize power losses, the converter employs 1.2-kV IGBTs on the MV-side. These high-voltage IGBTs facilitate ZVS operation, effectively reducing turn-on and turn-off losses.

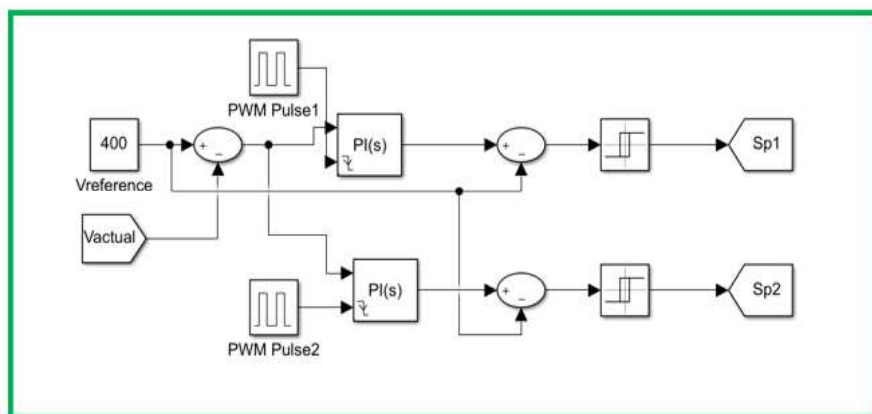
Primary leakage inductance (L_p) and magnetizing inductance (L_m) are two crucial inductors that are included in the converter. L_p is used to reduce current harmonics and mitigate unexpected voltage spikes. The electromagnetic interferences brought on by the high-frequency transformer's circulating currents are lessened significantly by L_m , on the other hand. Reducing these interferences lowers power losses on the power switches and HF transformer, which eventually improves the converter's overall efficiency.

Table 4.2 lists the precise specifications of the proposed dc-dc converter. Two H-bridges integrated with an HF transformer make up the converter, which enables efficient power conversion and control.

Table 4.2. The proposed DAB converter parameters

Parameters of DAB Module	Value	Parameters of DAB Module	Value
Power	12.5 kW	Input Capacitor	0.88 mF
Input Voltage	750 V	Output Capacitor	0.67 mF
Output Voltage & Current	400 V, 30 A	Switching Frequency	20 kHz
Resonant Tank Magnetizing Inductance (L_m) and Leakage Inductance (L_p)	140 mH -2.2 mH	Magnetization Resistance (Ω) and Inductance (H)	1.8 k Ω – 3H
Resonant Tank Capacitor (C_s)	1mF	Turn ratio (N)	2:1

The DAB converter makes use of PWM signals in the power switches to ensure effective functioning and the desired output voltage. The control process is essential for preserving system stability and increasing overall effectiveness. The output voltage is compared to a reference voltage in the voltage loop control technique shown in Figure 4.7 to calculate the error value. To produce the required phase-shift angle and PWM signals, this error value first goes through PI control and is then sent to a limiter.

**Figure 4.7.** Output voltage control of the recommended DAB converter

The converter can effectively regulate the output voltage by using this control technique. Both the LV-side PWM and the MV-side bridges are subjected to the changing phase angle, enabling accurate voltage control and modification. This guarantees that throughout the operation, the output voltage is constant and meets the desired parameters. Under low power conditions, 1.2-kV IGBTs are used on the MV side to reduce power switch losses and enable ZVS, while 600-V MOSFETs are used in the LV levels. These component selections help the DAB converter operate more effectively and efficiently.

Figure 4.8 presents the behavior of the MV and LV-side capacitor voltages, which serve as an indicator of the voltage control performance. Despite the varying power conditions caused by changes in weather conditions in the solar PV power plant, the voltage stability is effectively maintained. To further regulate the MV-side capacitor voltage, an LC filter is incorporated outside the PWM rectifier. This additional filter helps to ensure consistent and controlled voltage levels, contributing to the overall stability and reliability of the system.

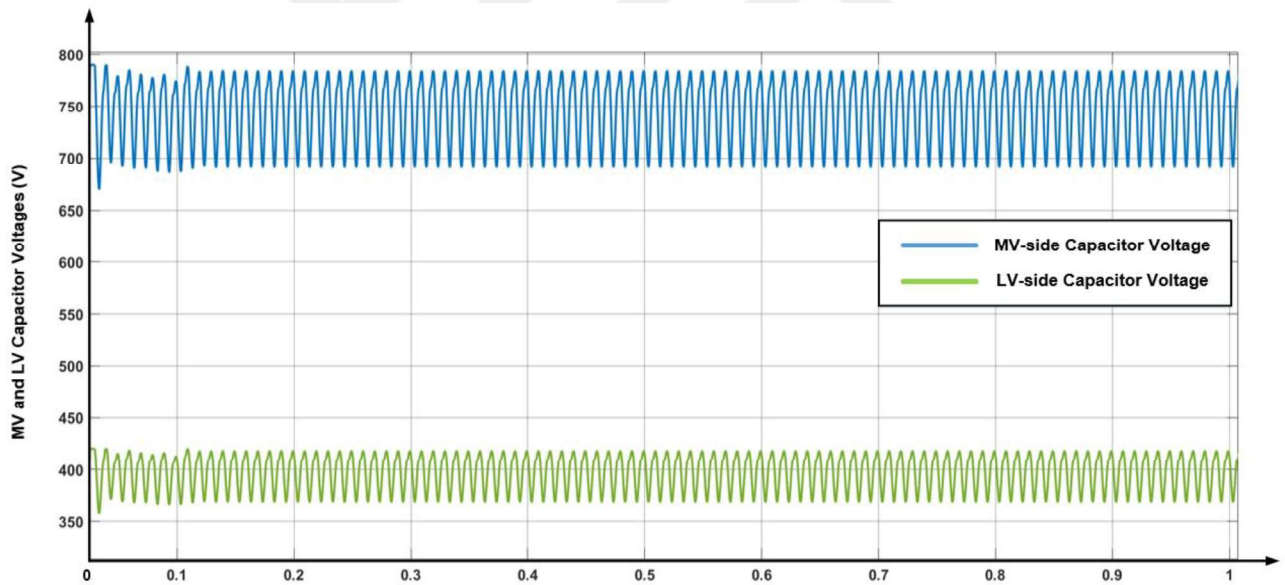


Figure 4.8. Input and output side voltages of the dc-dc bi-directional DAB converter

By using ZVS operation, the DAB converter topology is specifically created to reduce switching losses. An illustration of ZVS's successful integration into the suggested converter is shown in Figure 4.9. To achieve ZVS, it is essential to choose the right power switches and control strategies.

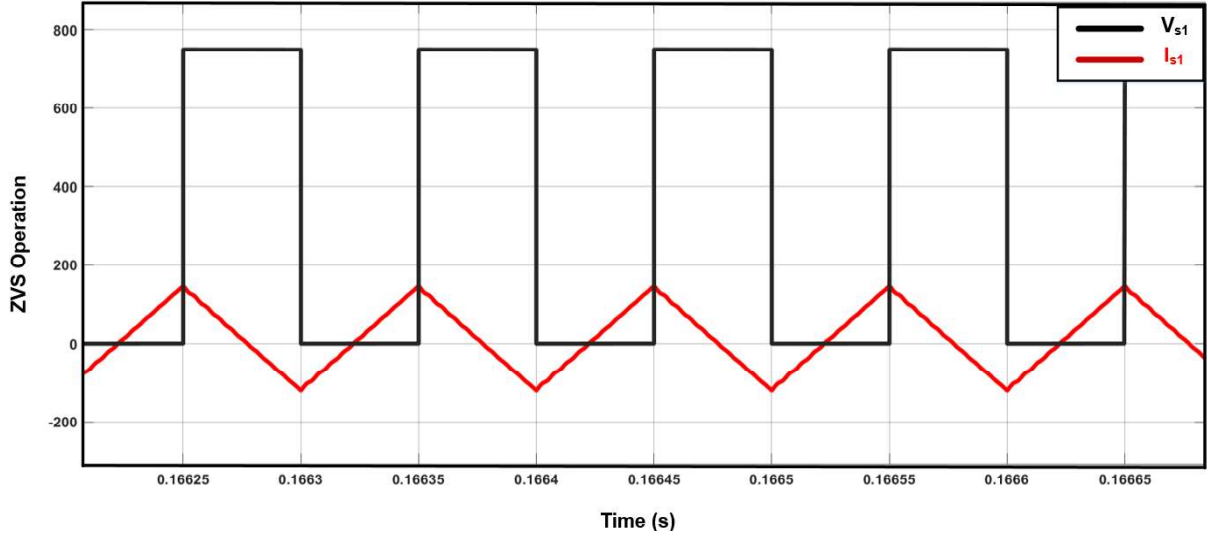


Figure 4.9. ZVS capability of the proposed DAB converter

ZVS turn-on is accomplished at the primary side of this isolated converter by operating at the resonance point. The capacitors (C_1 and C_2) are connected in parallel with the switches and the transformer leakage inductance generate a resonant mode. Any time the polarity of the MV-side H-bridge input voltage (V_{PI}) and V_{SI} is altered, this resonance manifests. For instance, in Figure 4.6, switches S_2 and S_3 are fully switched on before the time hits zero but S_1 and S_4 stay off. S_3 is still on at this time, but S_2 has been turned off at ZVS. The switch capacitors facilitate the capacitive energy transfer from S_2 to S_1 by resonance between L_p , C_1 , and C_2 . Due to the fact that the resonant operation only takes place for a short amount of the switching time, this topology is known as a quasi-resonant ZVS.

Following this, the voltage across switch S_2 becomes V_{in1} , and diode D_1 conducts the load current. The resonant inductor is charged by the DC voltage, and as the anti-parallel diode is still conducting, the ZVS turn-on is ensured for the active switch. When the load current becomes positive, S_1 conducts and can be turned off by the control logic to initiate the next resonance operation.

In inverters, the working principle is based on applying a DC input voltage in the positive direction during the first half period and in the negative direction during the second half period.

The sum of these two half-periods determines the operating frequency of the circuit. By switching the source between positive and negative, it is relatively easy to generate a square wave. Depending on the circuit configuration, a square wave can be converted into a sine wave signal using PWM (Pulse Width Modulation) and filtering methods. The waveform can be further refined until an ideal sine signal is achieved. Typically, positive and negative signals are generated for each phase and applied to their respective positive and negative elements. However, in this case, the signals are generated independently from the AC mains and as needed. Control signals are obtained by comparing the alternating control voltages, which can be in the form of square waves or sine waves, with an alternating carrier voltage in the form of a triangle wave at higher frequencies. This allows for control of the desired frequency and number of phases. In the PWM method, the inverters switch the DC voltage on and off for specific durations, varying the pulse width of each switching period to approximate the effect of a sine wave at the output.

Figure 4.10 presents the key components of a DC-AC inverter. The input filter serves to eliminate any ripples or frequency distortions present in the DC power source, ensuring a clean voltage supply to the inverter circuit. The inverter itself forms the main power circuit and is responsible for converting the DC voltage into a multilevel PWM waveform as required. The output filter plays a crucial role in attenuating high-frequency components present in the PWM waveform, thereby contributing to the generation of a signal that closely resembles a sinusoidal waveform.

In this design, the Sinusoidal Pulse Width Modulation (SPWM) technique is employed for inverter control. This technique enables direct control over the output voltage and frequency, aligned with sine functions. SPWM involves generating constant amplitude pulses with varying cycle times in each period. The pulse width of these pulses is modulated to control the output voltage of the converter and minimize harmonics. In the sinusoidal pulse width modulation technique, three sine waves and high-frequency triangle carrier waves are utilized to generate the PWM signal.

The amplitude modulation rate determines the output voltage amplitude in sinusoidal pulse width modulation. As the modulation ratio approaches 1 for higher output voltages, overmodulation occurs, causing distortion in the sinusoidal PWM output. The inverter must simultaneously adjust both the frequency and amplitude of the output voltage, and this adjustment is achieved through pulse width modulation. The objective is to closely approximate a sinusoidal voltage at the desired frequency. To achieve a sinusoidal voltage output at the desired frequency, the frequency of the control signal must match the desired frequency at the output.

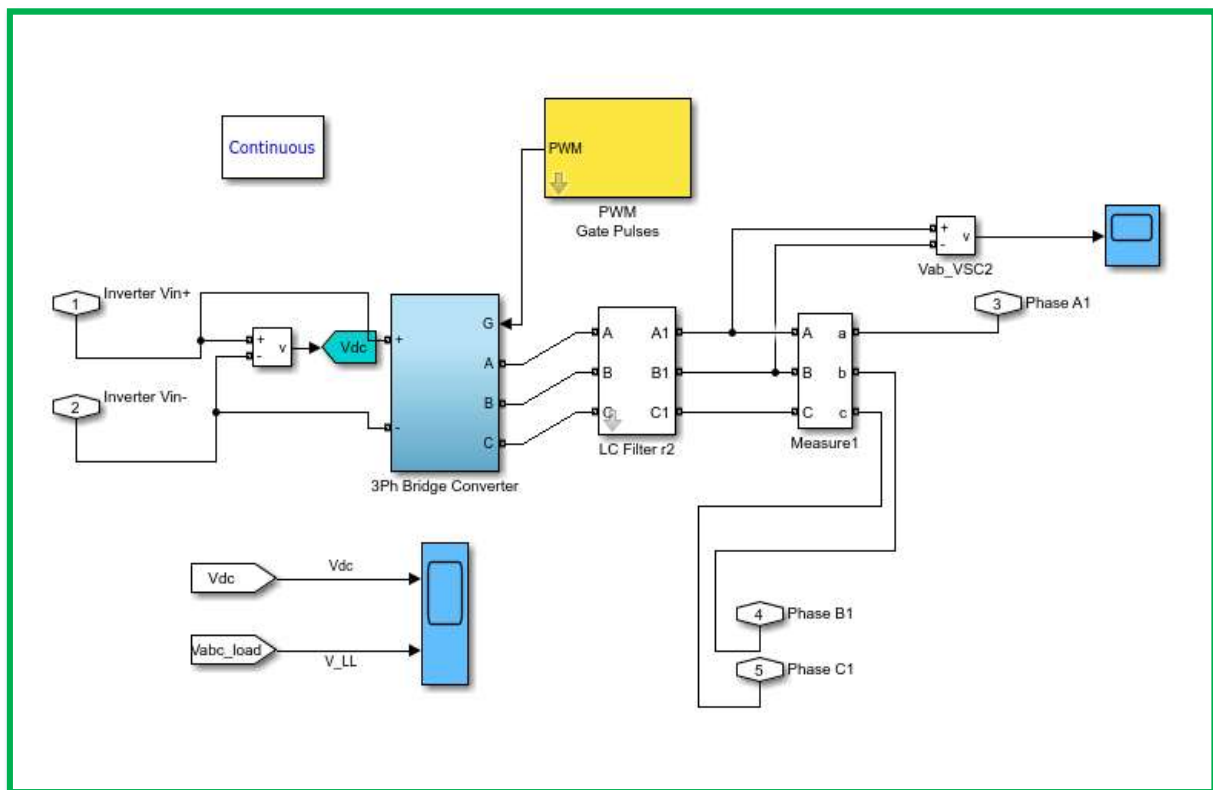


Figure 4.10. PWM inverter for DC-AC stage of the SST

The proposed DC-DC converter structure plays a crucial role in enabling efficient and high-quality electricity utilization for residential purposes. By employing this converter, the system ensures improved power control and enhanced power quality, resulting in an efficient output of electricity. Typically, obtaining pure electrical waveforms in renewable energy-connected electric grids can be challenging. However, the proposed DAB converter-based SST integrated

within a renewable energy grid system, as depicted in Figure 4.11, makes it possible to achieve this desired outcome.

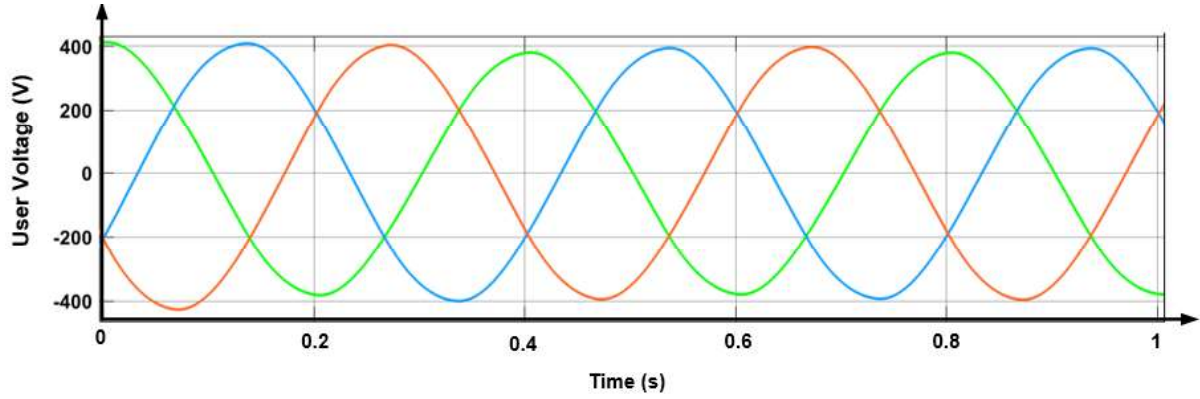


Figure 4.11. End-user regulated voltage

4.2. Experimental Studies for Proposed SST Application

The proposed MVdc SST topology (Medium Voltage DC Solid-State Transformer) along with its control strategy is given in Figure 4.12. This SST is designed to interface with either MVdc or LVac loads or sources within smart grids. The core component of the system is the dc-dc converter module, which utilizes a soft-switching smart transformer topology. This topology offers reduced conduction losses and incorporates voltage, current, and PWM control strategies for optimal performance. All of the primary components may achieve ZVS conditions thanks to these resonant circuits, which boosts performance and efficiency.

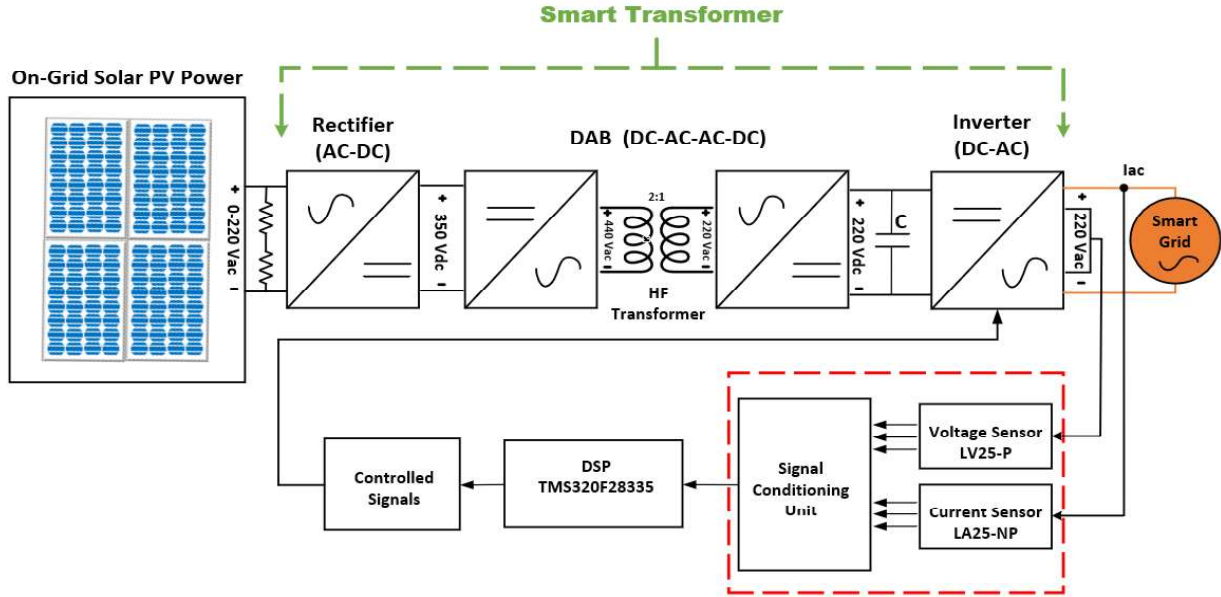


Figure 4.12. The proposed SST topology for the smart power grid

4.2.1. PV Solar System

Grid-connected solar systems are designed to utilize electricity generated by solar panels, allowing for immediate consumption of the generated power or storage for later use. Any surplus energy can be fed back into the main power grid. Effective energy management inside power networks has become essential as the assembly of such solar PV systems rises. In order to explore the input power of a smart transformer, a 12.5 kWp on-grid solar PV power plant was used in this study. 30 units of 415 W solar PV panels and 6 sets of 2 kW on-grid inverter modules made up the system. The solar PV power plant was installed on a rooftop to meet the energy demands of a residential property. The system experiences fluctuations in voltage due to varying weather conditions. Experimental tests were conducted on this power plant using the proposed smart transformer to ensure the stability and reliability of the smart grid.

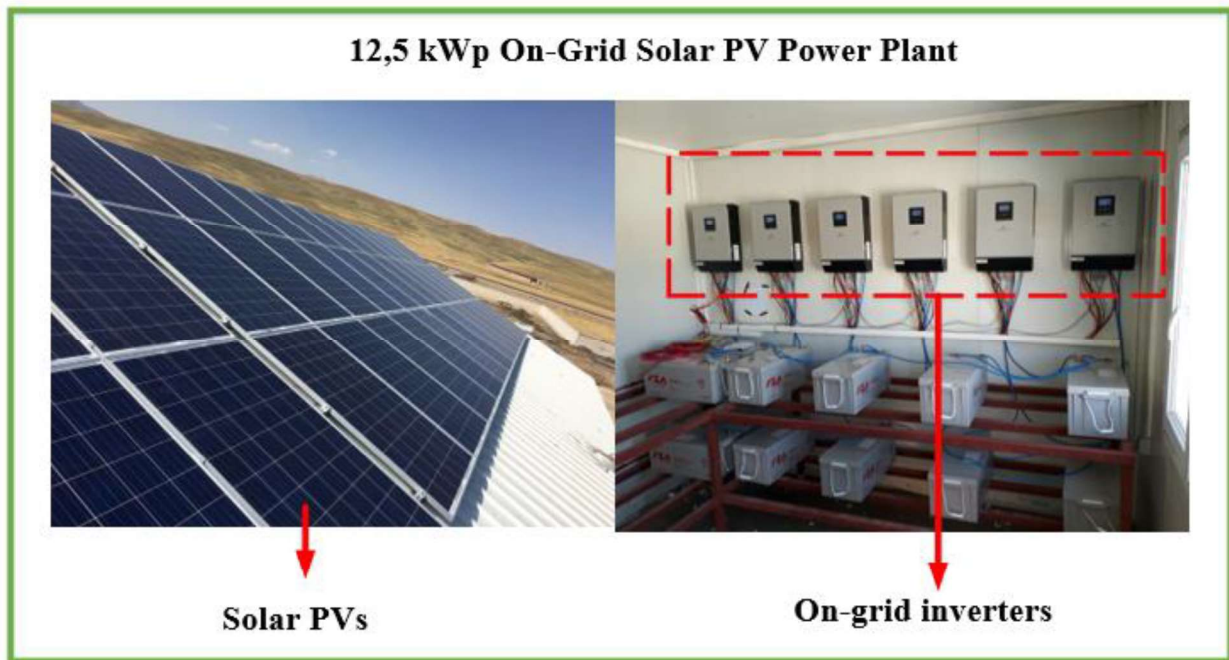


Figure 4.13. 12.5 kWp on-grid solar PV power plant

4.2.2. Bidirectional Isolated DAB Converter Design, Control and Manufacturing

As the overall efficiency of the power system heavily depends on the effectiveness of this stage, the dc-dc conversion stage is essential to the operation of three-level SSTs. As they serve as DC transformers, high-frequency isolated dc-dc converters are crucial for future smart grid applications. However, to achieve high conversion efficiency and galvanic isolation, the dc-dc conversion stage of the SST poses substantial problems and demands.

Fast switching speeds, zero load current, minimal hard turn-on losses, low di/dt characteristics, high-frequency isolation, high-rated power, soft switching, and high power density are some of these requirements. To address these needs in varied applications, researchers have used a variety of converter topologies.

The isolated DAB converter, which supports ZVS operation and bidirectional power flow, is one often used topology. Because switching losses are decreased during ZVS operation, efficiency, and power density are increased. DAB converters are thus appropriate for battery management systems and DC microgrid applications. These converters must overcome

problems including large circulating currents at heavier loads and ZVS failure at lighter loads. Low load operation results in switching losses, and harmonics can develop as a result of high dv/dt , low load operation, and electromagnetic interference in the HF transformer. Therefore, choosing the right power switches and designing the HF transformer are crucial for the converter to achieve high conversion efficiency.

A high-frequency transformer and two h-bridge bridges that operate at low voltage and medium voltage levels make up the planned DAB dc-dc converter. The DAB converter enables voltage reduction from 800 Vdc to 400 Vdc and has a power rating of 10 kVA. The HF transformer runs at 10 kVA of power and 20 kHz of switching frequency. The converter uses voltage and current control methods as well as phase-shifted duty square waves to regulate power flow.

The magnetizing inductance and the principal leakage inductance are two of the inductors used in the DAB converter. While the magnetizing inductance helps reduce electromagnetic interferences brought on by circulating currents in the HF transformer, the leakage inductance reduces current harmonics and prevents unanticipated voltage spikes. Reduced power losses in the HF transformer and power switches as a result of eliminating these interferences lead to increased converter efficiency.

The DAB converter can link with both the LVdc and LVac grids and provides bidirectional power transmission, making it appropriate for smart grid applications. It implements ZVS control while employing an ideal control approach to achieve improved efficiency. The proposed dc-dc converter module also has full-range ZVS capability, reduced dv/dt (rate of change of voltage), single-stage high-frequency link power conversion architecture, and robustness with protective measures against faults like short circuit current limiting from the dc-link inductor, desaturation protection in gate drivers, and mitigation of parasitic-inductance-induced voltage spikes.

4.2.2.1. Operation Principle of DAB Converter

Isolated bidirectional DAB converters consist of two full bridge circuits, an HF transformer, and two complete bridge circuits, as shown in Figure 3.8. The high-frequency transformer's leakage inductance is displayed. As for DAB's voltage gain conversion ratio:

$$d = \frac{aC_{MV}}{C_{LV}} \quad (4.1)$$

The switching frequency is specified as f_s , while the phase difference between the two bridge circuits is given as φ . L_{sst} is the MV side bridge circuit's leakage inductance, and T_{hs} is one-half of the switching time. When two h-bridge circuits are connected, the transmission power is:

$$P_T = \frac{aC_{MV}C_{LV}}{\omega_s L_{sst}} \varphi \left(1 - \frac{\varphi}{\pi}\right) \quad (4.2)$$

Additionally, I_{sst1} and I_{sst2} represent the leakage inductance currents of two bridge circuits.

$$I_{sst1} = -\frac{T_{hs}C_{MV}}{2\pi L_{sst}} (2d\varphi + \pi(1-d)) \quad (4.3)$$

$$I_{sst2} = \frac{T_{hs}C_{MV}}{2\pi L_{sst}} (2\varphi + \pi(d-1)) \quad (4.4)$$

The currents flowing through the DAB converter have a significant impact on the switching losses of the circuit. To achieve ZVS in both bridges, specific conditions must be met. In one bridge, the current (I_{sst1}) should be less than zero, while in the other bridge, the current (I_{sst2}) should be greater than zero. During this scenario, the body diode of the switch that needs to be turned on receives the current from the leakage inductance.

To ensure reliable operation of the high voltage and high power converters, the DAB converter incorporates a deadband, which is a specific time interval where no switches are turned on. The deadband protects the converter from undesirable effects. It is crucial to study the converter at this point to determine the amount of current from the leakage inductance that is necessary for ZVS turn-on operation at the beginning of the deadband.

Figure 4.14 presents the key waveforms of the DAB converter during the deadband period ($[t_0, t_2]$), providing insight into its behavior. Figure 4.15 presents the equivalent circuits of the DAB converter, which include the current path in various modes during the deadband.

In summary, the currents, ZVS operation, and deadband play essential roles in the performance and protection of the DAB converter, and analyzing the waveforms and equivalent circuits during the deadband provides valuable information about its behavior.

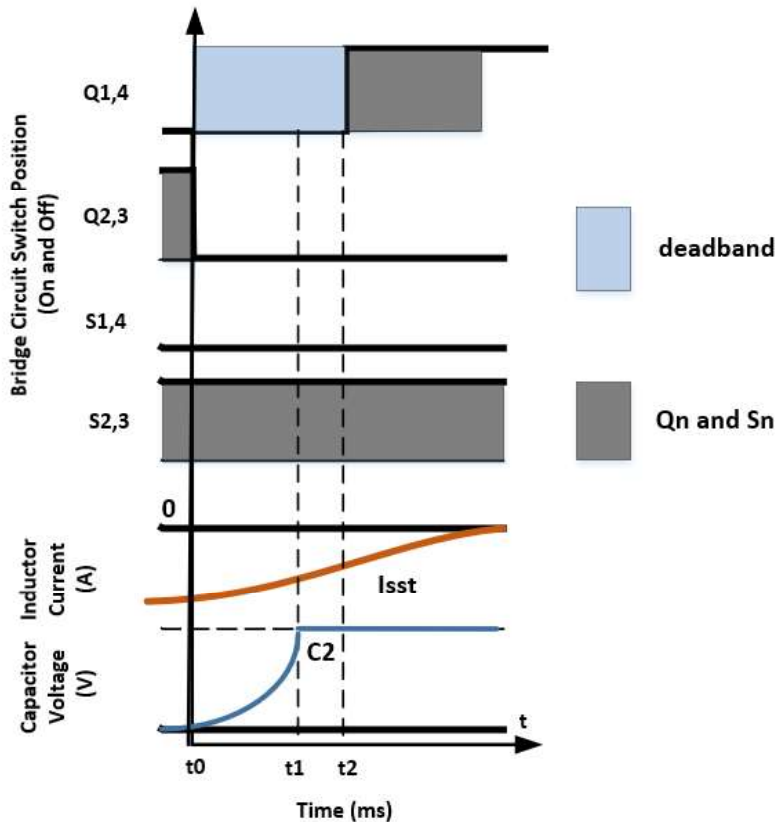
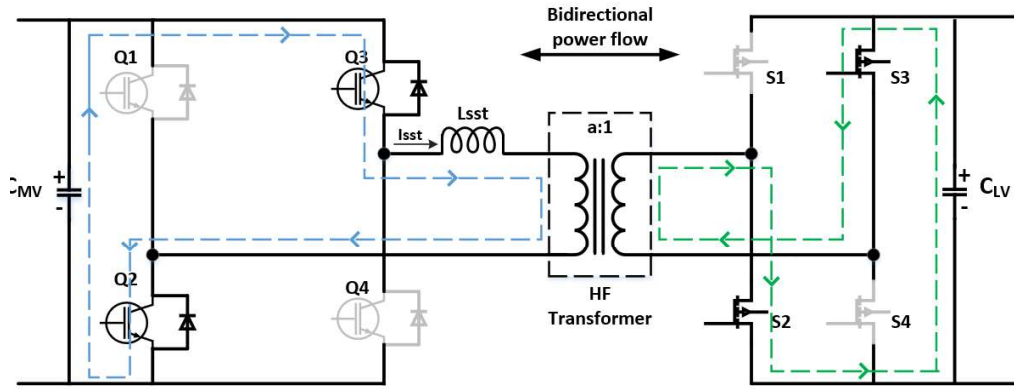
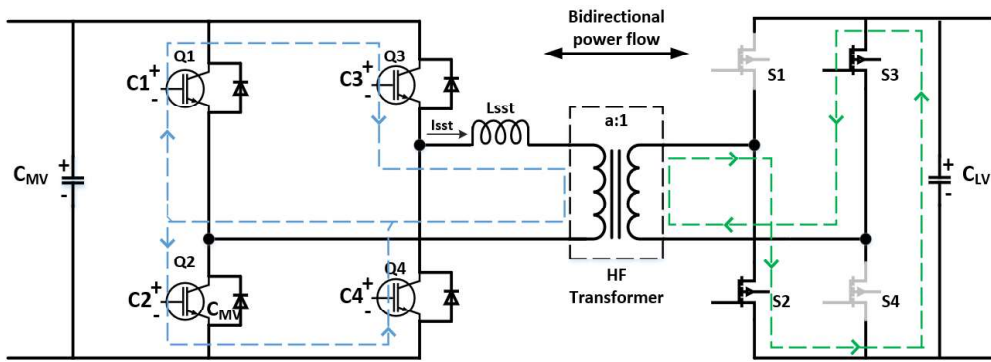


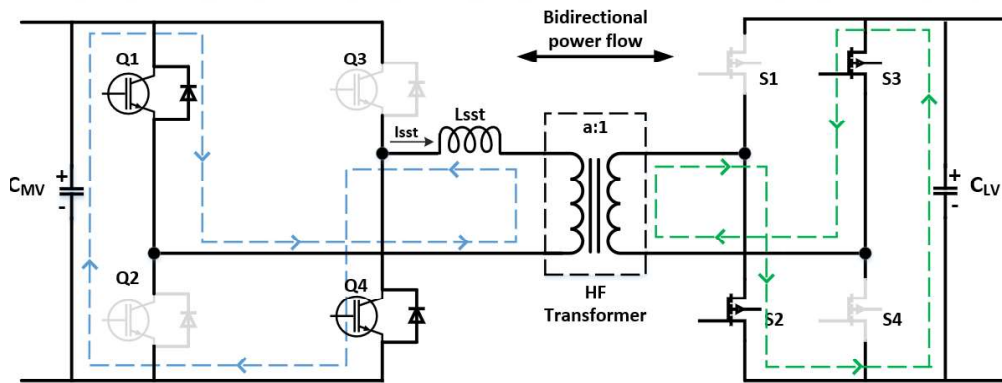
Figure 4.14. The voltage of the ZVS capacitor of C_2 under deadband, the leakage inductance current I_{sst} , and the PWM gate signals of power switches



(a)



(b)



(c)

Figure 4.15. The current paths of DAB converter in various power switch positions (a) mode 0 ($[t < t_0]$) b) mode 1 ($[t_0 t_1]$) c) mode 2 ($[t_1 t_2]$)

4.2.2.2. Control Scheme

Phase shift modulation techniques are commonly employed due to their ease of application and control capabilities. However, it is important to note that the phase shift modulation strategy alone is suitable for steady-state conditions. To ensure optimal performance and protection, additional control algorithms are utilized for managing overcurrent, and overvoltage, and achieving soft-switching operation.

The extended phase shift control method, a single phase shift control method based on a PI controller, and the output voltage or current control methods are a few other control techniques that are applied. These algorithms are especially helpful in solar PV systems that are connected to the grid since they allow for efficient converter current and voltage balance regulation.

To regulate the PWM signals of the inverter module in the aforementioned study, a DSP control card and MATLAB/Simulink application are used in tandem. An LV25-P sensor and a LA25-NP sensor are used to measure the output voltage and current, respectively. As shown in Figure 4.16, these measurements are then scaled and processed using Simulink blocks to give the PWM signals accurate control and dependable output signals. Figure 4.17 also shows the DAB's duty cycle control, demonstrating the degree of control over the converter's switching behavior.

In summary, while phase shift modulation techniques form the basis of control, the inclusion of additional control algorithms, combined with advanced control hardware and software, enables comprehensive management of various operating conditions and ensures the generation of reliable output signals.

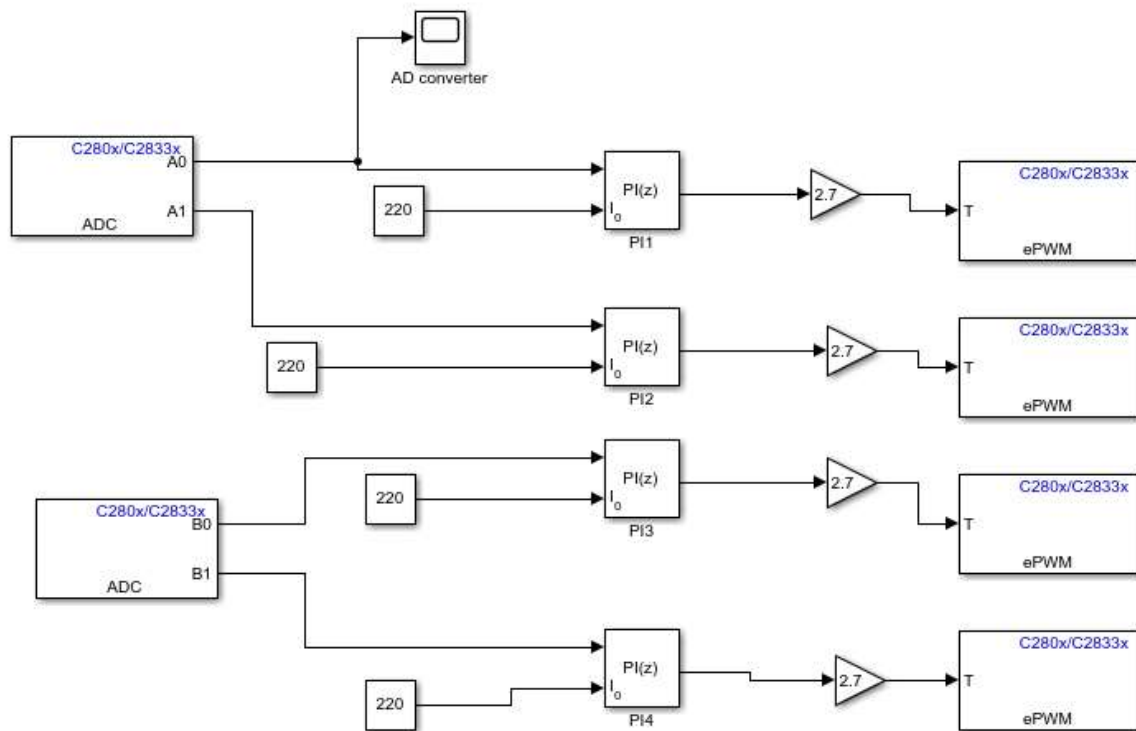


Figure 4.16. Controlled PWM signals producing for the PWM inverter

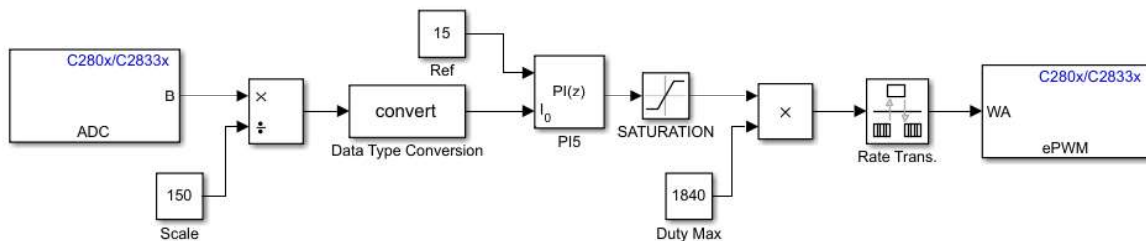


Figure 4.17. Controlled duty for DAB converter using DSP card and simulink block

To measure the voltage in a single-phase power system, a LEM LV 25-P voltage sensor is utilized. This voltage sensor has a conversion ratio of 220/5 volts, meaning that it converts the high voltage of 220V to a lower voltage of 5V. The voltage sensor is supplied with a positive voltage of +15V and a negative voltage of -15V. The output of the voltage sensor is then connected to a zero-crossing detector, which detects the points where the voltage crosses zero.

Figure 4.18 depicts the schematic for the voltage sensor that is linked to the circuit in parallel. The measurement resistance, R_m , is shown in this diagram while the principal input resistance, R_i , is shown. In this investigation, a single-phase 220V system operating at a frequency of 50Hz is being used as the power supply voltage.

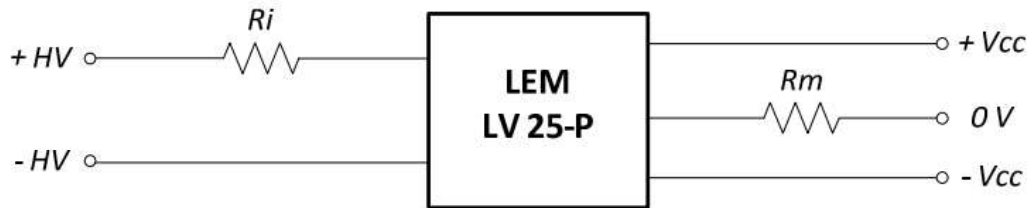


Figure 4.18. The schematic diagram for the LV 25-P voltage converter [64]

The voltage sensor's output waveform needs to fall within the required range for compatibility with the ADC range used by the DSP TMS320F28335. The LV 25-P voltage converter's datasheet states that its conversion ratio is 2500:1000, which means that the input voltage level of 220V is lowered to 5V at AC RMS value [64].

The appropriate resistors are chosen using Ohm's Law and the primary rated RMS current of 10mA. The secondary side of the voltage converter is designed to preserve the conversion ratio stated in the datasheet, and the secondary output current is pulled at 25mA when the input current is 10mA. Using Ohm's Law, the major input resistance (R_i) and measurement resistance (R_m) are computed. R_i is set to 22k.

In summary, the LEM LV 25-P voltage sensor is used to measure the voltage in a single-phase power system. It converts the high voltage of 220V to a lower voltage of 5V, which is compatible with the ADC range of the DSP. By using appropriate resistors based on the conversion ratio, the system voltage level is effectively changed to 5V at AC RMS value, enabling accurate voltage measurement.

To measure the current in the system, a current sensor is required to convert the input current into a corresponding voltage value. In this study, the LA 25-NP current transducer is employed

for this purpose. The preferred connection configuration for the LA 25-NP transducer is depicted in detail in Figure 4.19.

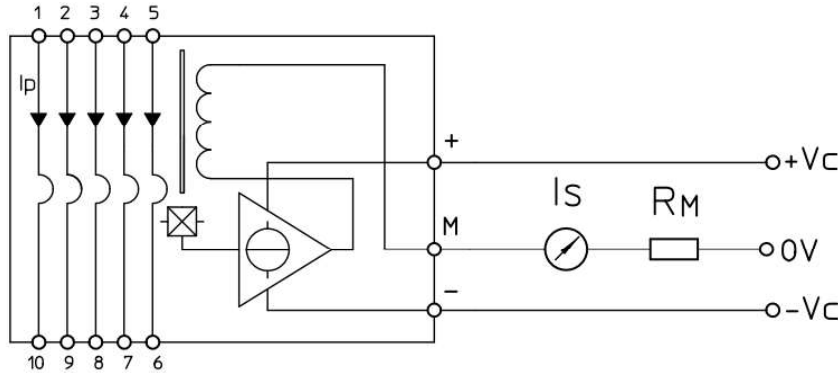


Figure 4.19. The schematic diagram for the LA 25-NP current converter [65]

To accurately measure both voltage and current in the single-phase power system, a power measurement board is utilized, which includes both voltage and current transducers. However, the voltage values obtained from these transducers need to be adjusted to match the ADC voltage level specified by the DSP.

To achieve this adjustment, a conversion circuit is employed. This circuit rescales the measured voltage values from the transducers, ensuring compatibility with the ADC voltage level of the designated DSP. By applying the appropriate rescaling techniques, the voltage values can be effectively adjusted to match the required ADC voltage level, enabling accurate measurement of both voltage and current in the system.

To accommodate the input requirements of the DSP, which operates within a 0-3 V input data range, the low-level bipolar voltage signals in the ± 5 V range need to be adjusted to fit within the 0-3 V range. This signal conversion is achieved through the implementation of a specific circuit, as depicted in Figure 4.20.

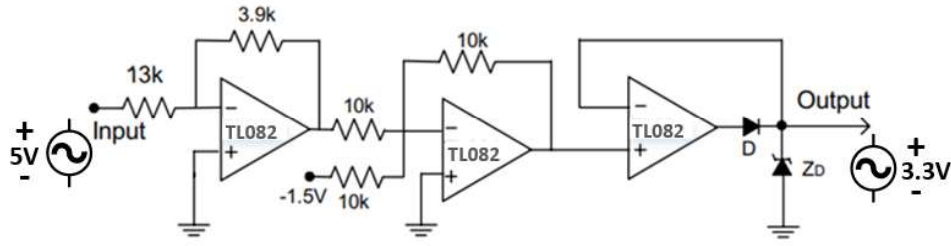


Figure 4.20. Signal conversion circuit for ADC port of DSP card: Bipolar ($\pm 5\text{V}$) to unipolar (0-3V)

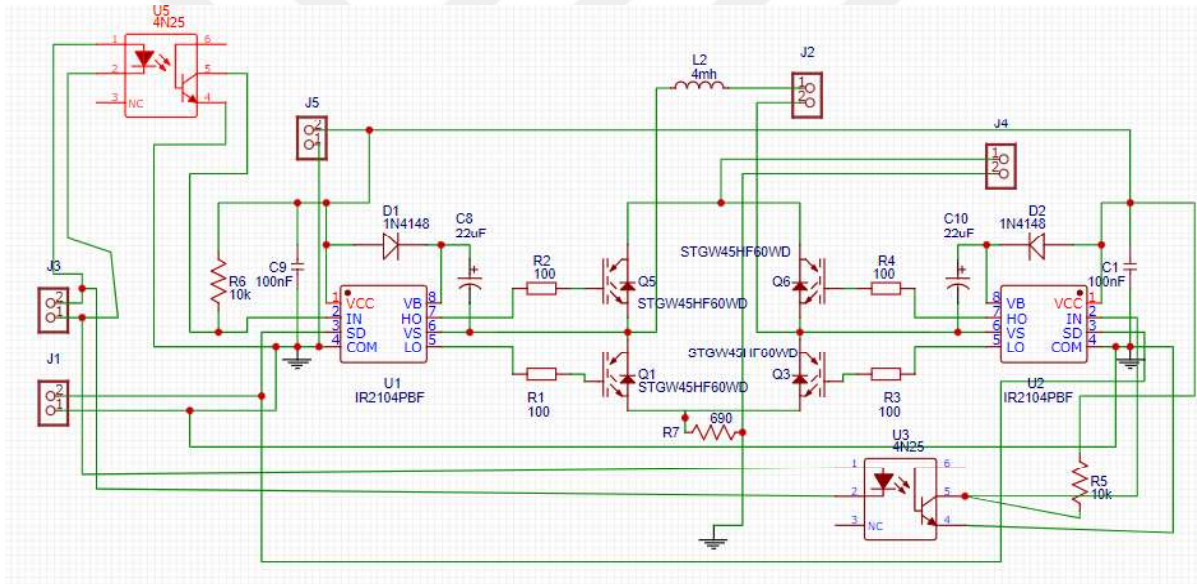
The circuit configuration ensures that the voltage signals, originally spanning the $\pm 5\text{ V}$ range, are effectively transformed and scaled down to the desired 0-3 V range. By utilizing appropriate components and techniques, the circuit enables seamless integration of the voltage signals with the DSP, facilitating accurate and reliable data processing within the specified voltage limits.

4.2.2.3. *Experimental Model of DAB Converter*

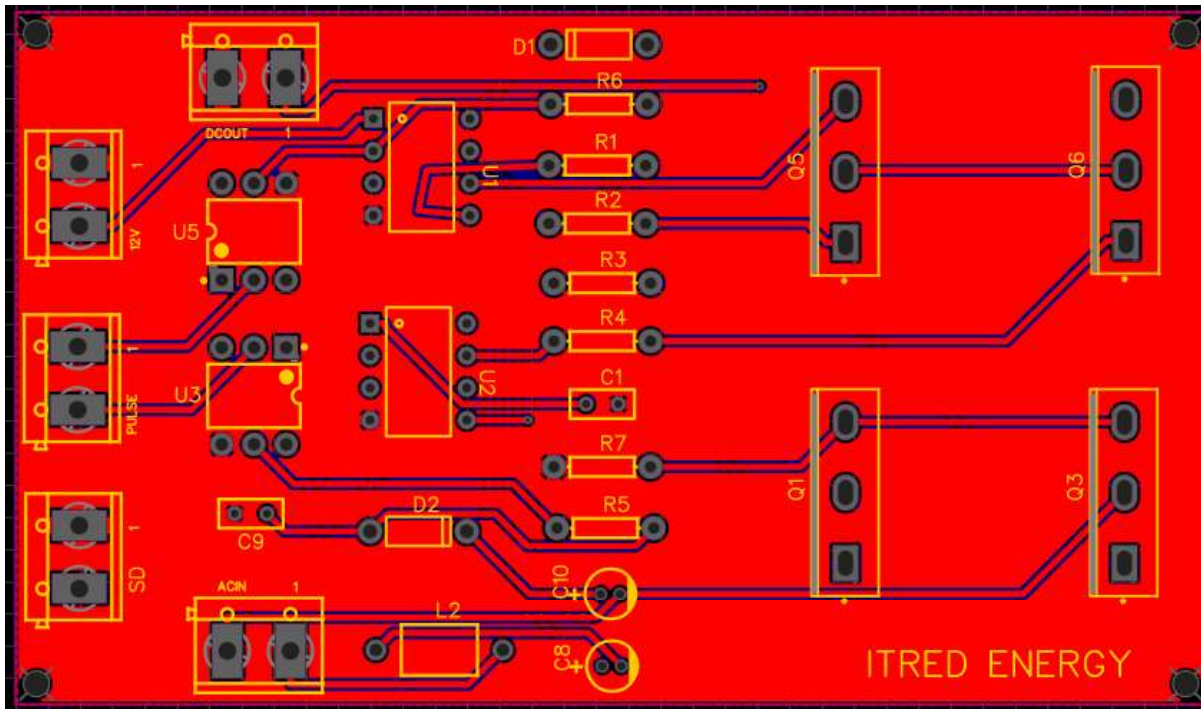
Gate driver circuits are vital because they act as the link between control circuitry and high-power semiconductor devices in power electronics systems. The efficiency of the gate driver circuits has a significant impact on the behavior and performance of power electronics converters like DA-DA converters and SMPS. As a result, the creation of gate driver circuits is crucial to the entire design of power electronics converters.

High-power IGBTs are driven in this thesis using the HCPL-3120 driving integrated circuit. An integrated circuit for the power output stage is optically coupled to the GaAsP LED found inside the HCPL-3120 gate driver optocouplers. The necessary driving power for MOSFETs and IGBTs used in motor control applications turns out to be perfectly supplied by this design. The output stage of the HCPL-3120 is a good option for driving IGBTs rated up to 650 V/40 A because of its wide working voltage range, which supplies the necessary drive voltages for gate-controlled devices. For driving high-power MOSFETs, an IR2104 IC in conjunction with a 4N25 isolated optocoupler is utilized to create a driver circuit.

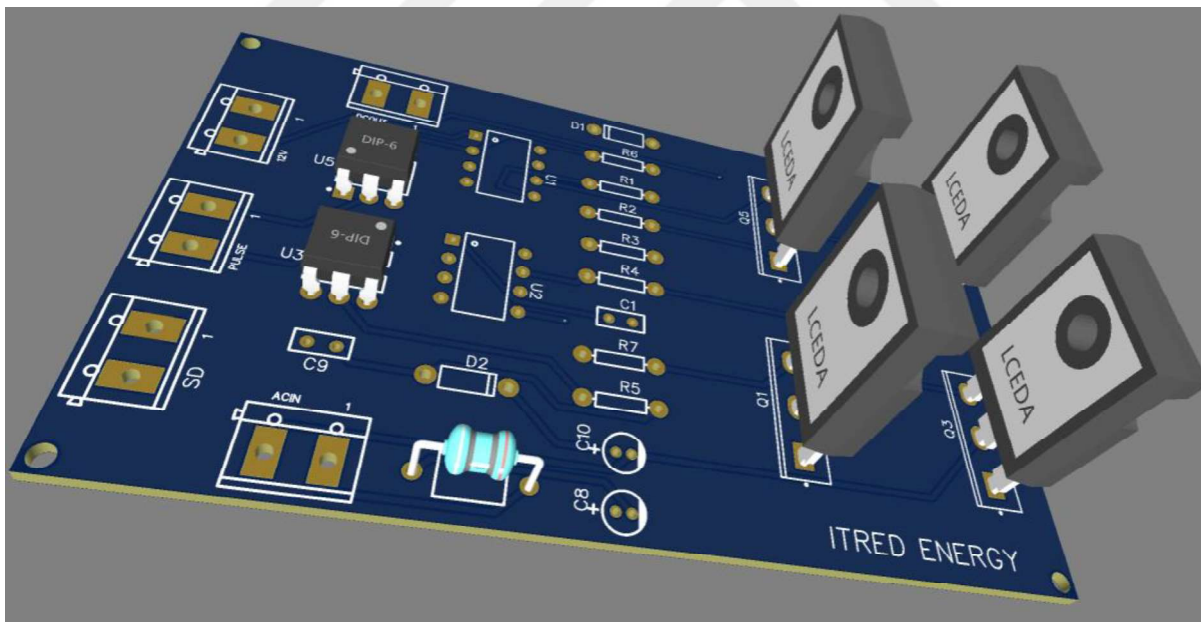
The DAB bidirectional DC/DC converter topology has various advantages, including a small number of devices, soft-switched commutations, low cost, and high efficiency. Particularly for situations where dependability, cost, weight, and power density are crucial considerations, it is advocated. In the DA-DA converter's control method, trigger signals are provided by the Phase Shift Modulation technique, which is commonly suggested in the literature. This modulation method is appropriate for situations involving fluctuating DC connection voltages and allows for efficient power transmission between converters. Maximum efficiency is ensured when the DC link voltages are consistent and closely matched to the transformer conversion ratio. An H-bridge arrangement is used in the design and implementation of the suggested dual-direction DAB converter. The DAB converter card's design and manufacturing process is shown in Figure 4.21.



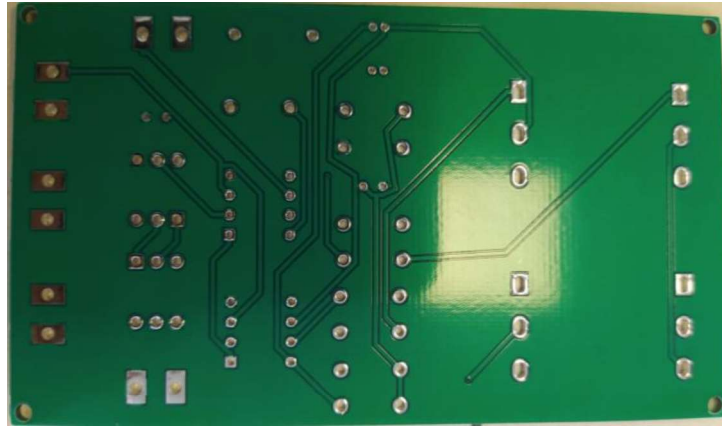
(a)



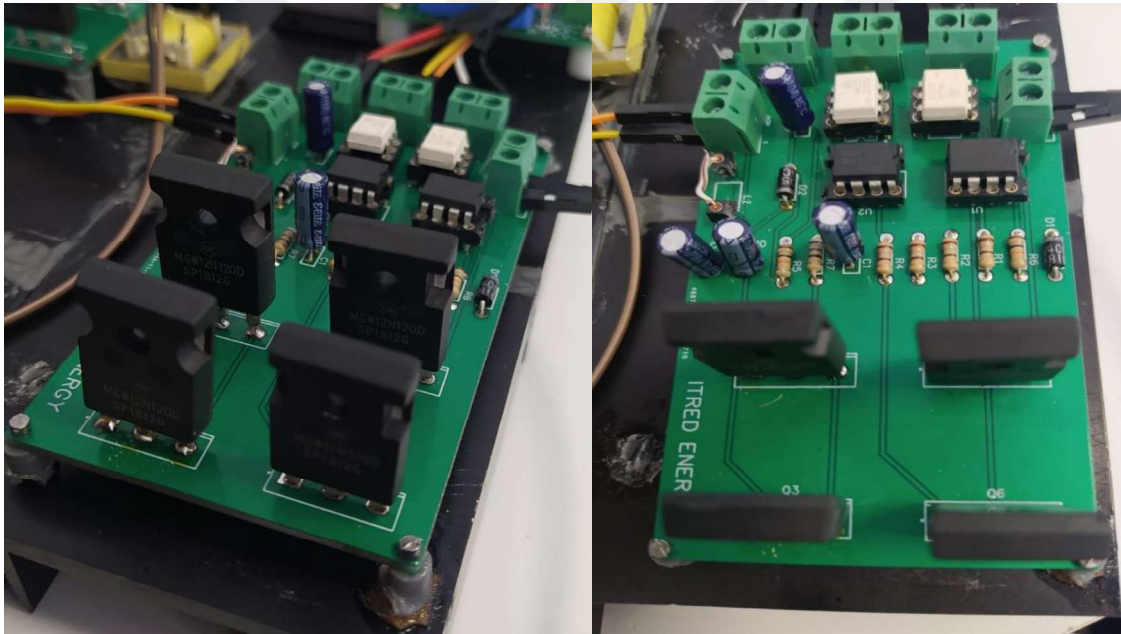
(b)



(c)



(d)



(e)

Figure 4.21. The designed DAB converter a) circuit diagram, b) 2D board view, c) 3D board view d) rear view e) physically produced and assembled converter

The gate signals of the PWM signals driven in the designed converter are seen in Figure 4.22. The gate signals were converted with a 50% duty cycle and $\pm 20\text{V}$ supply voltage by

successfully driving the power switches. In full bridge converters located on the input and output side of the transformer, cross-trigger signals are square waves with a 50% duty cycle. Efficient operation of converters is ensured by giving phase shifts in input and output full bridges. At the same time, PWM gate signals are realized by the controller.

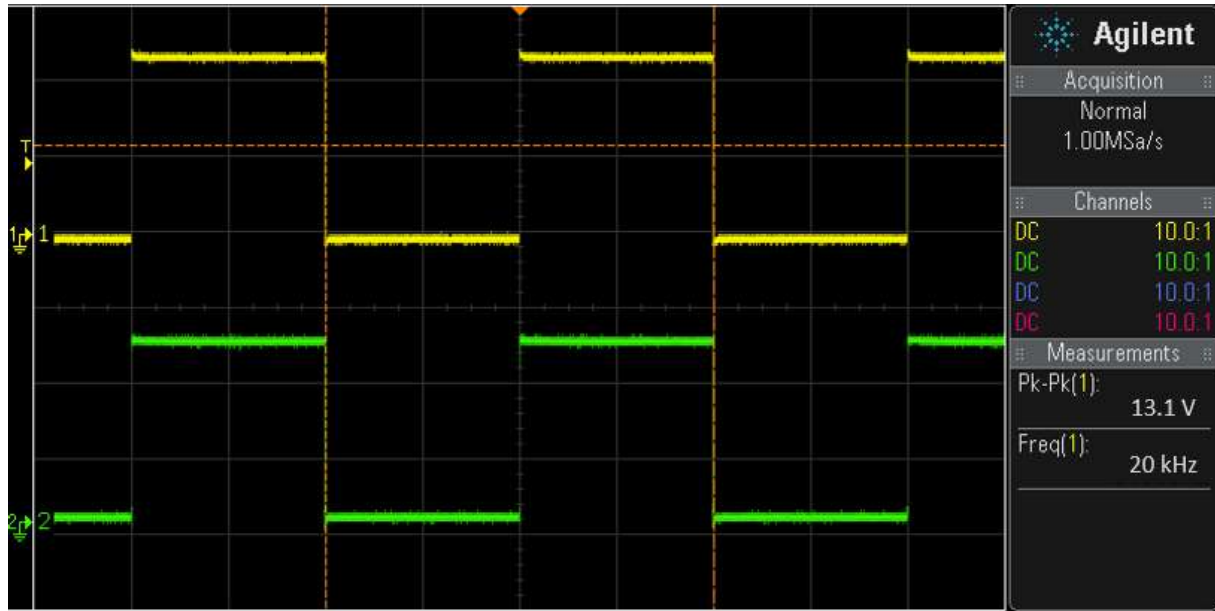


Figure 4.22. Gate signals of PWM signals driven in the designed converter

The electronic control system of the converter requires a microprocessor to implement the developed control methods, convert voltage and current signals into numerical values, and generate ignition signals for the IGBTs. To ensure real-time control and high-performance compensation functions, a microprocessor with a high-resolution cycle speed is essential. DSPs are the preferred choice for achieving the required sampling and resolution cycle rates in the control methods of the proposed system.

In this regard, DSP-based microcontrollers are considered more suitable as they integrate ADC, PWM, and communication modules into the electronic control system. The F28335 processor from Texas Instruments is a notable example of such a microcontroller. It supports programming in C/C++ and assembly languages, providing flexibility for control applications in power electronics systems.

Traditionally, writing programs for control applications in power electronics systems could be a time-consuming and challenging task. However, with the advancement of new-generation DSP development tools, programming has become less of an obstacle, allowing engineers to focus more on control methods. One widely used software development tool is MATLAB/Simulink. By designing the system model in MATLAB/Simulink to adhere to discrete-time and real-time operation, it is possible to generate C code from the model, simplifying the programming process.

The Code Composer Studio program facilitates the loading of programming language code or automatically generated codes from the simulation program into the processor. This integrated development environment further streamlines the programming and deployment process for the microprocessor.

To ensure the proper functioning of the Mosfet and IGBT drivers in the converter, converter, and inverter card designs, dedicated DC feeds are required. Each card is designed separately to prevent electromagnetic interference and is supplied with different voltage DC feeds in isolation. Isolation is crucial to address issues related to grounding, harmonic fluctuations, and short-circuit protection within the cards. These challenges are effectively mitigated by implementing a multi-output isolated DC supply design.

The design incorporates insulated DC-DC (5V-5V) converters, which allow the 5V voltage obtained from the control card to be supplied directly to the converter cards without the need for external DC voltages. This voltage is first routed through the insulated DC-DC converters to ensure proper ground isolation. Figure 4.23 illustrates the circuit design of the isolated DC-DC converter.

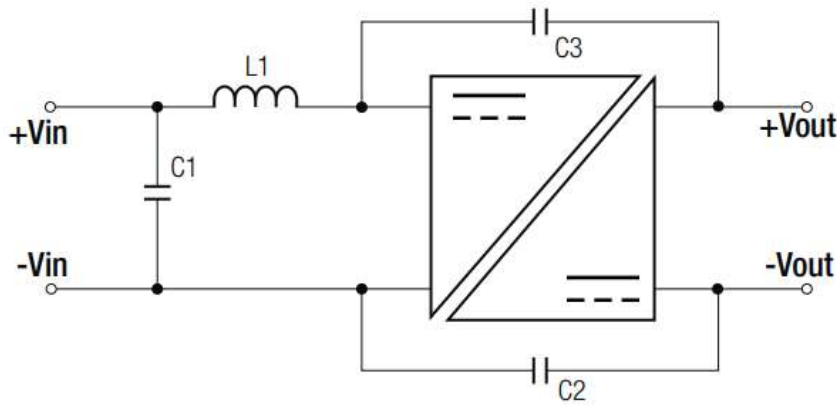


Figure 4.23. Isolated converter circuit diagram (5v-5v)

The isolated DC/DC converter used in this design is specifically chosen for its cost-effectiveness and suitability for general-purpose power isolation and voltage-matching applications. Despite its affordability, the converter offers high-quality features, including 1kVDC isolation and reliable operation within an industrial temperature range of -40°C to $+85^{\circ}\text{C}$.

XL6009 IC offers adjustable supply voltage ranging from 5V to 50V. The DC/DC converter with a wide input range, current mode, and the ability to produce either positive or negative output voltages is the XL6009 regulator. It can be set up as an inverting converter, flyback, SEPIC, or boost converter. Current mode architecture and an integrated N-channel power MOSFET in the XL6009 enable steady performance over a broad supply and output voltage range. In Figure 4.24, the application circuit is shown. Figure 4.25 shows the design for isolated DC card feeds.

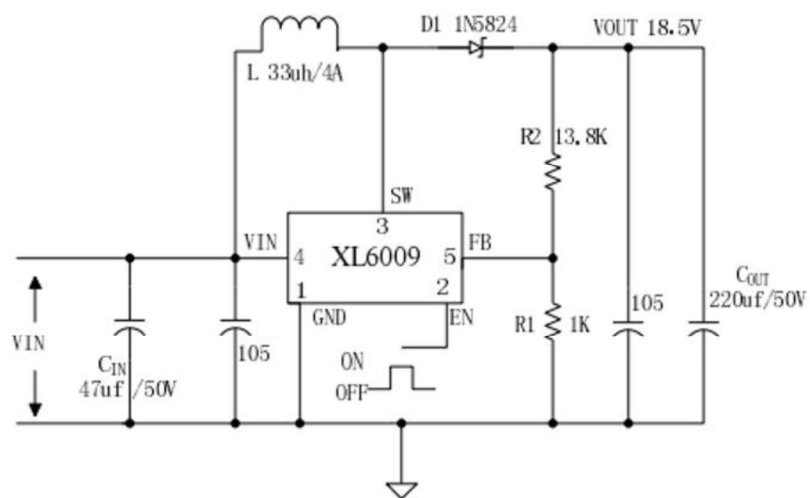


Figure 4.24. DC amplifier design circuit (5v-50v) [66]

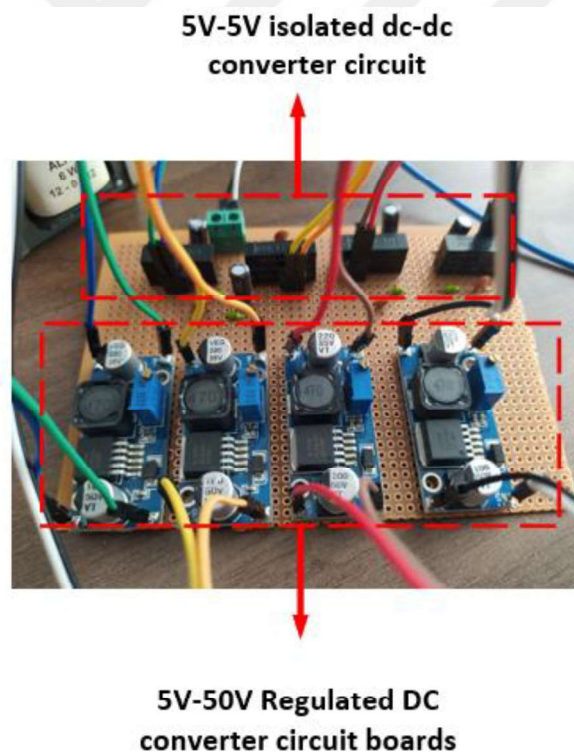


Figure 4.25. Isolated dc card feeds design

4.2.2.4. *HF Transformer Design*

The design and analysis of a 10kVA 800/400 V Yoke-Ferrite transformer have been carried out in this study. High-frequency transformers (HFTs) have gained significant attention due to the development of strong soft magnetic materials with low core loss in semiconductor power switches. Optimizing the design of HFTs is crucial for achieving high-performance energy conversion systems.

The design process involved mathematical calculations and 3D ANSYS/Maxwell simulations, which considered electromagnetic behavior and loss analysis. The simulations accounted for hysteresis losses, eddy current losses, copper losses, and magnetic flux density, providing accurate results to validate the mathematical model calculations. Various parameters were analyzed, including core magnetic flux density, core and copper losses, leakage and magnetizing inductances, winding parasitic capacitances, and input/output voltage and current ratings. The design parameters were optimized accordingly.

Furthermore, the overall efficiency of the proposed HF transformer was calculated and presented. The mathematical model considered a high-frequency transformer with the isolated discontinuous current return, taking into account pre-selected parameters that were chosen based on the specific power application. The magnetic core material properties were considered, including initial permeability, flux density, temperature, and frequency. For this particular design, a flux density of 0.30 T was chosen based on the magnetic core material properties.

The secondary winding of the transformer includes a voltage diode, typically considered as having a voltage drop (V_d) of 1.0 V. The ratio of the AC equivalent resistance to the DC equivalent resistance is assumed to be 1. Efficiency and cost are two crucial factors in the design of electrical machines. Therefore, initial designs are based on specific parameters and subsequently optimized for size and efficiency.

In this section, we provide a detailed mathematical model of the high-frequency flyback transformer. The shell depth (δ), also known as the skin depth, refers to the reduction in current-carrying capacity at higher frequencies due to the formation of eddy currents that primarily flow

on the outer surface of the conductor. The skin depth, denoted as δ , can be mathematically expressed for a given frequency F . It is important to note that as the frequency increases, the skin depth decreases, resulting in a higher concentration of current near the conductor's surface.

Overall, the mathematical model presented in Equation 4.5 offers valuable insights into the behavior and characteristics of the high-frequency flyback transformer, taking into account parameters such as the voltage diode, equivalent resistances, and skin depth phenomenon.

$$\delta = \frac{6.62}{\sqrt{f}} \text{ [cm]} \quad (4.5)$$

In high-frequency transformer design, the diameter of the wire denoted as D_{AWG} , plays a significant role. It is an important factor that affects the resistance and current characteristics of the transformer. The wire diameter needs to be carefully determined, taking into account these parameters for optimal high-frequency transformer design. The wire diameter (D_{AWG}) is expressed as twice skin depth. This equation allows for the calculation of the wire diameter based on the wire's cross-sectional area. By considering the properties and desired current characteristics of the transformer, the appropriate wire diameter can be determined to ensure efficient and reliable operation.

The area of the base wire denoted as A_w , is an important parameter in high-frequency transformer design. It represents the cross-sectional area of the wire and can vary depending on the frequency of operation. The value of A_w is directly proportional to the diameter of the wire (D_{AWG}).

When determining the value of A_w , reference is made to the AWG datasheet. The AWG parameter corresponding to the frequency value is used to calculate A_w . The general equation for calculating the area of the bare wire, as given in Equation 4.6, takes into account these considerations.

$$A_w = \frac{\pi * (D_{AWG})^2}{4} [cm^2] \quad (4.6)$$

By determining the appropriate value of A_w based on the chosen AWG and frequency, the cross-sectional area of the wire can be accurately determined. This information is crucial for designing high-frequency transformers that meet the desired performance and efficiency requirements.

The AWG system provides standardized sizes and specifications for wire, including specific frequency values. These frequency values are used to determine the appropriate AWG for a given application. By referring to the AWG table, the corresponding AWG value can be identified based on the desired frequency.

Once the AWG is determined, specific parameters such as wire diameter, bare wire area, and resistance values can be selected according to the standard. These parameters are chosen from Table 4.3, which presents the recommended values corresponding to the selected AWG and frequency. By following this standardized approach, the appropriate wire specifications can be established, ensuring reliable performance and compatibility with the desired frequency range.

Table 4.3. Cable parameters selected according to AWG standard

AWG	Diameter (mm)	Area (mm ²)	Resistance (ohm/km)	Max. current (A)	Max Frequency for 100% Skin Depth	Bare Area, (10 ⁻³ cm ²)	Resistance, ($\frac{10^{-6}\Omega}{cm}$)	Diameter, (cm)
<i>Selected Wire Size</i>	0.91186	0.653	26.407	1.8	21kHz	6.531	263.9	0.0948

In the design process, the time period (t) is determined based on the frequency (f). The maximum time for a transistor to be "on" (t_{on}) is also considered as given in Equation 4.7. Additionally, the maximum duty ratio (D_{max}) is selected as 50% in this design.

$$t_{on} = T * D_{max} \quad [sec] \quad 4.7$$

To put it differently, the design takes into account the time period associated with the chosen frequency and the maximum time that a transistor can remain in the "on" state. Furthermore, the maximum duty ratio is set at 50% to ensure a balanced operation of the system.

The secondary side of the transformer is divided into two parts, and the power losses in these sections are denoted as P_{o1} and P_{o2} . P_{o1} represents the power output in the first part of the secondary side, and it can be calculated using Equation 4.8. In simpler terms, P_{o1} refers to the power delivered by the first section of the transformer's secondary side.

$$P_{o1} = I_{o1} * (V_{o1} + V_d) \quad [Watts] \quad 4.8$$

The power output of the second part of the secondary side of the transformer is represented as P_{o2} and can be calculated using Equation 4.9. P_{o2} depends on the output voltage of the second section (V_{o2}), the output current of the second section (I_{o2}), and the diode voltage (V_d) typically taken as 1V. In simpler terms, P_{o2} refers to the power delivered by the second section of the transformer's secondary side, and it is determined by considering the output voltage, output current, and diode voltage.

$$P_{o2} = I_{o2} * (V_{o2} + V_d) \quad [Watts] \quad 4.9$$

P_{o1} represents the output power in the secondary LV choke, while P_{o2} represents the output power in the secondary HV choke. The maximum primary current, $I_{pri,max}$ shown in Equation 4.10, is determined by dividing the total power by the product of efficiency and the minimum input voltage.

$$I_{pri,max} = \frac{P_T}{\eta * V_{in_{min}}} \quad 4.10$$

The output voltage of the second part of the secondary side of the transformer is denoted as V_{o2} and is set to 25V based on safety standards and the desired output channel quantity. The output current of the second part, I_{o2} , is determined as 0.5A. The diode voltage, V_d , is typically considered as 1V. P_T represents the total secondary load power shown in Equation 4.11, η is the efficiency of the converter (usually taken as 0.9), V_{inmin} is the minimum input voltage. $I_{pri,pk}$ is defined as the primary peak current as given in Equation 4.12. In a discontinuous flyback transformer operating at maximum load and minimum input line, a 50% duty cycle is typically used. Although a lower duty cycle can be employed, it would result in higher peak and RMS currents. This would necessitate a larger duty cycle, more complex stability mechanisms, and a lower bandwidth. In this section, the RMS and peak values of the current are calculated for a duty cycle of 0.50.

$$P_T = P_{omax} = P_{o1} + P_{o2} \quad 4.11$$

$$I_{pri,pk} = \frac{2 * P_T * T}{\eta * V_{inmin} * t_{on}} \quad 4.12$$

The maximum input power, P_{in_max} , is a crucial parameter in the design of the medium frequency isolation transformer. It is determined based on the efficiency (η) of the converter, which is set to a minimum of 80%. P_T represents the maximum output power. Equation 4.13 provides the calculation for the maximum input power, P_{in_max} .

$$P_{in_max} = \frac{P_T}{\eta} \quad 4.13$$

The equivalent input resistance, denoted as R_{in_eq} , is a parameter that characterizes the input impedance of the system. With the known value of P_{in_max} , the value of R_{in_eq} can be determined using Equation 4.14.

$$R_{in_eq} = \frac{V_{in_min}^2}{P_{in_max}} \quad 4.14$$

The estimation of leakage inductance is a crucial aspect of the design process. It represents the inductance that influences the current peak during the duty cycle of the switching mechanism. The value of the primary inductance can be calculated using Equation 4.15.

$$L = \frac{R_{in_eq} * T * (D_{max})^2}{2} \quad 4.15$$

A flyback transformer, being a type of coupled inductor, operates by transferring the input voltage to the primary winding during each cycle. The energy is stored within the core's hollow cavity. Subsequently, this energy is transmitted to the secondary winding, which then powers the load. In the case of a filter inductor or a flyback transformer, a significant portion of the required energy storage is achieved by incorporating an air gap (or a non-magnetic material with a relative permeability, M_r , equal to 1) in series with the high-permeability core material. The magnetic energy stored in the inductor, denoted as E , represents its energy processing capacity, as depicted in Equation 4.16.

$$E = \frac{L * I_{pri, pk}^2}{2} \quad 4.16$$

The current flowing through the wire is represented as $I_{pri, pk}$. The constant K_e , which characterizes the electrical state, is influenced by the magnetic and electrical operating parameters, as defined in Equation 4.17. On the other hand, the kernel geometry constant K_g is determined by Equation 4.18.

$$K_e = (0.145) * P_T * B_m^2 * 10^{-4} \quad 4.17$$

$$K_g = \frac{(E)^2}{K_e * \alpha} \quad 4.18$$

A core with an ETD geometry, which has a similar kernel geometry to K_g , was selected for the design. The core geometry for ETD ferrite cores is illustrated in Figure 4.26. The specific dimensions and design data for ETD ferrite cores can be found in Table 4.4 and Table 4.5, respectively. In this particular case, the EDT-39 core type was chosen based on the core geometry (K_g) parameters. This core type is well-suited for high-frequency flyback transformer applications and is considered a suitable option for the design.

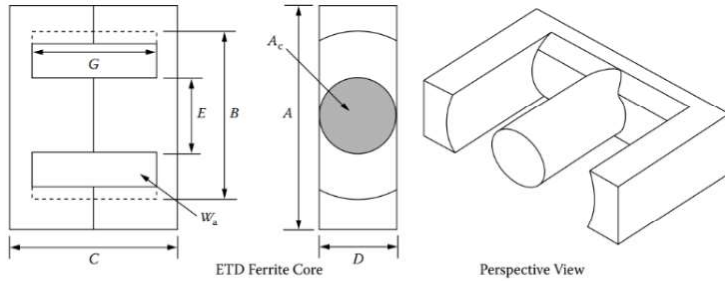


Figure 4.26. Core geometry for ETD ferrite cores [67]

Table 4.4. Dimension data for ETD ferrite cores

ETD, Ferrite Cores (Ferroxcube)						
Part No.	A (cm)	B (cm)	C (cm)	D (cm)	E (cm)	G (cm)
ETD-34	3.500	2.560	3.460	1.110	1.110	2.360
ETD-39	4.000	2.930	3.960	1.280	1.280	2.840
ETD-44	4.500	3.250	4.460	1.520	1.520	3.220

Table 4.5. Design data for ETD ferrite cores

ETD, Ferrite Cores (Ferroxcube)											
Part No.	W_{tcu} (grams)	W_{tfe} (grams)	MLT (cm)	MPL (cm)	W_a A_c	A_c (cm ²)	W_a (cm ²)	A_p (cm ⁴)	K_g (cm ⁵)	A_t (cm ²)	*AL (mh/1K)
ETD-34	43.4	40.0	7.1	7.87	1.757	0.974	1.711	1.6665	0.0914	53.4	1182
ETD-39	69.3	60.0	8.3	9.22	1.871	1.252	2.343	2.9334	0.1770	69.9	1318
ETD-44	93.2	94.0	9.4	10.30	1.599	1.742	2.785	4.8515	0.3596	87.9	1682

The current density, denoted as J , represents the amount of electric current flowing through a unit cross-sectional area of the conductor. It is used to determine the appropriate conductor area for the primary and secondary windings of the transformer. A higher current density results in a smaller conductor area, which reduces both the cost and size of the transformer. However, it also leads to increased I^2R losses, as the resistance of the conductor (R) increases with a smaller conductor area. Additionally, a higher current density can cause a temperature rise, potentially damaging the insulation of the winding. Therefore, it is important to select an optimal current density level. The calculation of J can be performed using Equation 4.19.

$$J = \frac{2 * (E) * (10^4)}{B_m * A_p * K_u} \quad 4.19$$

In Equation 4.20, $A_{pw}(B)$ represents the area of the primary wire in the transformer. The value of $A_{pw}(B)$ is determined based on the selected current density. The primary wire area, $A_{pw}(B)$, varies as the current density changes.

$$A_{pw}(B) = \frac{I_{pri_{RMS}}}{J} \quad 4.20$$

In a Litz-wire transformer winding, the optimal number of primary wires, denoted as S_{np} , is determined to minimize losses. The AC resistance factor decreases with a finer twist of the wires. However, the DC resistance increases as the insulation occupies more space in the winding window. To account for the insulation thickness and proximity effect losses, a power law model is employed. The required number of primary wire sequences is calculated based on the bare area value provided in Table 4.3. This calculation is performed using Equation 4.21

$$S_{np} = \frac{A_{pw}(B)}{(bare\ area)} \quad 4.21$$

The number of primary turns is represented by N_p , which indicates the total number of turns in the primary winding. The available window area dedicated to the primary winding is denoted as W_{ap} . The thread count, S_{np} , refers to the number of threads or wires in the winding. The bare space accounts for any gaps or insulation thickness in the winding. The equation for calculating the number of turns in the primary winding is as follows:

$$N_p = \frac{K_u * W_{ap}}{2 * (Bare\ Area)} \quad 4.22$$

The presence of an air gap in the core of a transformer has a significant demagnetizing effect, leading to a decrease in the permeability of high-permeability materials and disruption of the hysteresis loop. This effect remains consistent even with direct current excitation. However, the impact of a small air gap on core bias is relatively minor compared to magnetization properties. The magnitude of the air gap effect is influenced by factors such as the average magnetic path length and the properties of the core material without the gap.

A larger magnetic flux path mitigates the reduction in permeability, while a core with high permeability and low coercive force experiences a more pronounced decrease. The air gap extends the length of the magnetic path. In a transformer with highly inductive circuitry, minimal current flows when the voltage (E) applied across the primary winding (N_p) is affected. The maximum inductance is achieved when the air gap is smallest for a given core size.

After determining the number of primary turns, the calculation proceeds by considering the core geometry data. The required size of the air gap (l_g) is calculated using Equation 4.23.

$$l_g = \frac{0.4 * \pi * N^2 * A_c * (10^{-8})}{L} - \left(\frac{MPL}{\mu_m} \right) \quad 4.23$$

The movement of magnetic flux within magnetic core results in its dispersion or scattering into the surrounding medium, which is known as flux fringing (F). Flux fringing is particularly relevant in magnetic cores of flyback transformers that include an air gap. In the case of a hollow magnetic core, the fringe effect reduces the reluctance of the magnetic path, thereby increasing the inductance of the winding.

The fringe flux factor (F) that contributes to the increased inductance is determined by the shape of the magnetic core. An approximation of the fringe flux factor (F) can be obtained using Equation 4.24:

$$F = 1 + \frac{l_g}{\sqrt{A_c}} \ln \left(\frac{2 * G}{l_g} \right) \quad 4.24$$

The calculation of the additional turns in the primary winding denoted as N_{np} , takes into account the fringe flux factor (F) and the magnetic core geometry values. The equation for obtaining the number of additional turns in the primary winding is given by Equation 4.25.

$$N_{np} = \sqrt{\frac{l_g * L}{0.4 * \pi * A_c * F * (10^{-8})}} \quad 4.25$$

To estimate core losses and prevent saturation, it is necessary to determine the peak flux density (B_{pk}). This peak flux density is calculated by using the number of additional turns in the primary winding (N_{np}) and the magnetic core geometry data, as shown in Equation 4.26.

$$B_{pk} = \frac{0.4 * \pi * N_{np} * F * (I_{pri,pk}) * (10^{-4})}{l_g + \left(\frac{MPL}{\mu_m}\right)} \quad 4.26$$

The resistance value ($\mu\Omega/\text{cm}$) of the primary new AWG is calculated using Equation 4.27.

$$\text{New AWG resistance} = \frac{\mu \frac{\Omega}{\text{cm}}}{S_{np}} \quad 4.27$$

The resistance (R_p) of the primary winding can be calculated by considering the information from the magnetic core geometry and the newly calculated number of windings. In general, the primary winding resistance can be calculated using Equation 4.28.

$$R_p = MLT * (N_{np}) * \left(\frac{\mu\Omega}{\text{cm}}\right) * 10^{-6} \quad 4.28$$

The heat generated by electric currents in the primary winding conductors is referred to as primary copper loss (P_p) in a transformer given in Equation 4.29. Copper losses, occur due to Joule heating. According to Joule's First Law, the amount of energy or power lost per second is proportional to the square of the current flowing through the windings and the electrical resistance of the conductors. At high-frequency currents, the proximity effect and skin effect cause uneven distribution of current along the conductor, resulting in increased effective resistance. This complicates the calculation of losses. Litz wire is a type of wire designed to ensure even current distribution and reduce Joule heating.

$$P_p = I_{pri_{RMS}}^2 * R_p \quad 4.29$$

The secondary side of the flyback transformer is divided into two parts, and thus, the number of turns on the secondary side is calculated separately. Equation 4.30 provides the calculation for the turns of the first part of the secondary side, denoted as N_{s01} .

$$N_{s01} = \frac{N_{np} * (V_{01} + V_d) * (1 - D_{max} - D_{well})}{V_{in,min} * D_{max}} \quad 4.30$$

Equation 4.31 provides the calculation for the peak current on the secondary side for the first part, denoted as $I_{s01(pk)}$. It is calculated by considering the fringe flux, the number of new turns (N_{np}), the output voltage of the first section (V_{01}), the minimum voltage of the primary side (V_{in_min}), the maximum duty ratio (D_{max}), and the dwell time.

$$I_{s01(pk)} = \frac{2I_{01}}{(1 - D_{max} - D_{well})} \quad 4.31$$

Equation 4.32 provides the calculation for the RMS current on the secondary side for the first part, denoted as $I_{s01(pk)}$. It is calculated based on the output current of the secondary side for the first part (I_{01}), the maximum duty ratio (D_{max}), the dwell time duty ratio, and the peak current of the secondary side.

$$I_{s01(rms)} = I_{s01(pk)} * \sqrt{\frac{(1 - D_{max} - D_{well})}{3}} \quad 4.32$$

Equation 4.33 provides the calculation for the area of the secondary wire for the first part in the transformer, denoted as $A_{sw01(B)}$. The area of the secondary wire is dependent on the current density.

$$A_{sw01(B)} = \frac{I_{s01(rms)}}{J} \quad 4.33$$

The calculation of S_{ns01} , which represents the number of secondary strips required for the first part, is determined based on the bare area value AWG. The specific equation for calculating S_{ns01} is not provided in the given information.

$$S_{ns01} = \frac{A_{sw01(B)}}{A_{w,Bare}} \quad 4.34$$

S_{01} secondary, resistance value in $\mu\Omega/\text{cm}$ AWG and S_{01} calculation is given in Equation 4.35.

$$(S_{01})\mu\Omega/\text{cm} = \frac{\frac{\mu\Omega}{\text{cm}}}{S_{ns01}} \quad 4.35$$

The resistance of the secondary winding (R_{s01}) can be calculated using Equation 4.36, taking into account the units of microohm/centimeter and the parameter MLT. This equation provides a quantitative measure of the resistance present in the secondary winding.

$$R_{s01} = MLT * (N_{s01}) * \frac{\mu\Omega}{\text{cm}} * 10^{-6} \quad 4.36$$

In the MF transformer, the copper loss includes both AC and DC losses. However, since AC resistance is typically negligible at low frequencies, the DC loss can be used as an approximation for the copper loss. It is important to consider eddy current losses resulting from skin and proximity effects, especially in medium and high-frequency transformers. These losses should be estimated for both the primary and secondary windings. In the case of the first part, the secondary copper loss is denoted as P_{s01} in Equation 4.37.

$$P_{s01} = I_{s01}^2 * R_{s01} \quad 4.37$$

N_{s02} refers to the number of turns in the second part of the secondary winding of the flyback transformer. It is calculated separately for each section of the secondary side.

$$N_{s02} = \frac{N_{np} * (V_{02} + V_d) * (1 - D_{max} - D_{well})}{V_{in_{min}} * D_{max}} \quad 4.38$$

The peak current flowing through the second part of the secondary side ($I_{s02(pk)}$) can be calculated based on the input of fringe flux, the number of new turns (N_{np}), the output voltage of the second part (V_{02}), the minimum voltage of the primary side (V_{in_min}), the maximum duty ratio (D_{maxis}), the dwell time duty ratio, and the diode voltage (V_d). Equation 4.39 is used to express the calculation of $I_{s02(pk)}$ for the second part.

$$I_{s02(pk)} = \frac{2I_{02}}{(1 - D_{max} - D_{well})} \quad 4.39$$

The output current of the secondary side for the second part is denoted as I_{02} . The rms current of the secondary side for the second part is represented as $I_{s02(rms)}$. The maximum duty ratio is calculated using the dwell time duty ratio and the peak current of the secondary side.

$$I_{s02(rms)} = I_{s02(pk)} * \sqrt{\frac{(1 - D_{max} - D_{well})}{3}} \quad 4.40$$

$A_{sw02(B)}$ refers to the area of the secondary wire for the second part in the transformer. The area of the secondary wire is influenced by the current density. The calculation of $A_{sw02(B)}$ is given by Equation 4.41.

$$A_{sw02(B)} = \frac{I_{s02(rms)}}{J} \quad 4.41$$

Here, $I_{s02(rms)}$ represents the RMS current of the secondary side, and J represents the current density for the second part. The number of secondary strips S_{ns02} required for the second part is calculated based on the bare area value. The calculation of S_{ns02} is given by Equation 4.42.

$$S_{ns02} = \frac{A_{sw02(B)}}{A_{w,Bare}} \quad 4.42$$

In Equation 4.43, S_{02} represents the secondary resistance in $\mu\Omega/\text{cm}$ AWG (American Wire Gauge). The resistance value is obtained from Table 4.3.

$$(S_{02})\mu\Omega/\text{cm} = \frac{\mu \frac{\Omega}{\text{cm}}}{S_{ns02}} \quad 4.43$$

The calculation of secondary winding resistance (R_{s02}) is given in Equation 4.44.

$$R_{s02} = MLT * (N_{s02}) * \frac{\mu\Omega}{\text{cm}} * 10^{-6} \quad 4.44$$

In Equation 4.45, P_{s02} represents the copper loss of the second part of the secondary side. It is calculated based on the secondary current, secondary resistance, and number of turns for the second part.

$$P_{s02} = I_{s02}^2 R_{s02} \quad 4.45$$

The window utilization factor (K_u) represents the amount of copper that occupies the window area of the transformer or inductor. It is calculated based on the areas of the primary winding (S_{np}) and the secondary windings (S_{ns01} and S_{ns02}). The equation to calculate the window utilization factor is not explicitly provided, but it can be determined by dividing the total cross-sectional area of the windings by the total window area.

$$\text{Primary turns} = (N_{np} S_{np}) \quad 4.46$$

$$\text{Secondary turns}_1 = (N_{s01} S_{ns01}) \quad 4.47$$

$$\text{Secondary turns}_2 = (N_{s02} S_{ns02}) \quad 4.48$$

$$K_u = \frac{N_t A_w}{W_a} \quad 4.49$$

When a transformer operates, there is a power difference between the input and output sides. This power difference is converted into heat due to various losses within the transformer. These losses can be categorized into two main types: core losses (P_{fe}) and copper losses (P_{cu}).

The core losses are constant losses that occur in the transformer core and are independent of the load current. They result from the magnetic properties of the core material and include hysteresis and eddy current losses. Core losses remain relatively constant regardless of the load conditions.

On the other hand, copper losses are variable losses that occur in the transformer windings. They are directly related to the load current and cause power dissipation in the form of heat. Copper losses depend on the resistance of the winding conductors and increase with higher load currents.

To calculate the total copper loss (P_{cu}), the losses from both the primary and secondary windings of the high-frequency transformer are added together. This accounts for the power dissipated as heat in the windings due to their resistance.

The efficiency of a transformer is determined by the ratio of the output power to the input power. It takes into account both the core losses and the copper losses. Higher efficiency means less power is wasted as heat.

Regulation refers to the ability of the transformer to maintain a relatively constant output voltage even when the load conditions change. It is influenced by the transformer design and the balance between core and copper losses.

The temperature rise of a transformer is directly related to the power losses. Excessive temperature rise can be detrimental to the transformer's performance and lifespan, so it is important to keep it within acceptable limits through proper design and cooling measures.

$$P_{cu} = P_p + P_{s01} + P_{s02} \quad 4.50$$

The regulation indicates how well the transformer can handle variations in the load without significantly affecting the output voltage. When the transformer operates at its maximum efficiency, which occurs when the constant loss equals the quadratic loss at the rated voltage, the regulation is calculated using Equation 4.51.

Essentially, the regulation quantifies the voltage stability of the transformer, and a lower regulation value implies better voltage regulation under varying load conditions.

$$\alpha = \frac{P_{cu}}{P_t} \times 100 \quad 4.51$$

In high-frequency transformers, there is a power loss between the input and output due to factors like core loss and copper loss. This power loss represents the difference between the input power (P_{in}) and the output power (P_{out}) of the transformer. The efficiency of a transformer is an important measure of its performance. When the core loss is equal to the copper loss, the total loss is minimized, and the efficiency is maximized.

Copper losses are directly related to the properties of the material used in the transformer. The design of a transformer is influenced by various factors such as size, material availability, temperature rise, and cost. The use of ferrous alloys and ferrites in the construction of power inductors offers significant advantages. The properties of magnetic core materials, including ferrous alloys and ferrites, are provided in Table 4.6, which helps in selecting the appropriate material for a given application.

Table 4.6 compares the effect of different materials on the flux density. Based on the comparison, ferrite is chosen as the most suitable material among the options considered. This selection is made considering factors such as magnetic properties, performance, and other design considerations.

Table 4.6. Magnetic core material properties

MATERIAL NAME	COMPOSITION	INITIAL PERMEABILITY μ_i	FLUX DENSITY B_s	CURIE TEMPERATURE °C	DENSITY GRAMS/cm ³ δ
SILICON	3-97 SiFe	1500	1.5-1.8	750	7.63
ORTHONOL	50-50 NiFe	2000	1.42-.158	500	8.24
PERMALLOY	80-20 NiFe	25000	0.66-0.82	460	8.73
AMORPHOUS	81-3.5 FeSi	1500	1.5-1.6	370	7.32
AMORPHOUS	66-4 CoFe	800	0.57	250	7.59
AMORPHOUS (μ)	73-15 FeSi	30000	1.0-1.2	460	7.73
FERRITE	<i>MnZn</i>	<i>2500</i>	<i>0.5</i>	<i>>230</i>	<i>4.8</i>

AC flux density B_{ac} is calculated as in Equation 4.52.

$$B_{ac} = \frac{(0.4) * \pi * N_{np} * F * (\frac{I_{pri(pk)}}{2})(10^{-4})}{l_g + (\frac{MPL}{\mu_m})} \quad 4.52$$

The formula for watts per kilogram (WK), which realizes the core loss confirmed by the manufacturers, is presented below. Equation 4.53 contains the WK calculation.

$$\frac{\text{watts}}{\text{kilogram}} (WK) = k(f)^{(m)} (B_{ac})^{(n)} \quad 4.53$$

The values were extracted from a table of core loss coefficients for magnetic ferrite materials. Material P with $k = 1.983 \times 10^{-3}$, $m = 1.36$, and $n = 2.86$, frequency 100 kHz. The most important ac characteristic is core loss (P_{fe}). Ac core loss is influenced by a magnetic material, magnetic material thickness, magnetic flux density, frequency, and operating temperature. Because of this, selecting a magnetic material involves obtaining the best property while taking into account common trade-offs like price, size, and efficiency. The manufacturer does not utilize the same units when expressing core loss. The user should note different core loss units while contrasting various magnetic materials. As can be seen in Equation 4.54, computations are done in milliwatts per gram in this case.

$$P_{fe} = \left(\frac{\text{milliwatts}}{\text{gram}} \right) W_{tfe} * 10^{-3} \quad 4.54$$

The efficiency of a transformer serves as a valuable metric for assessing the efficiency of its design. It is defined as the ratio of output power (P_o) to input power (P_{in}). The difference between P_o and P_{in} is primarily attributed to losses occurring in the transformer. These losses include both core losses (P_{fe}) and copper losses (P_{cu}) in the windings of the high-frequency transformer. The total loss (P_{Σ}) is the sum of these losses, measured in watts, and can be calculated using Equation 4.55.

$$P_{\Sigma} = P_{fe} + P_{cu} \quad 4.55$$

Here, P_{fe} is the core loss and P_{cu} is the copper loss of the transformer. The parameters of a 10KVA 800/400 V flyback type high-frequency transformer are obtained by the mathematical model of the transformer. In Table 4.7, the weight, electrical and electromagnetic parameters of the transformer are given in detail.

Table 4.7. 800/400 V single phase flyback type high frequency transformer design parameters obtained using the mathematical model of the transformer.

Symbol	Parameters	Unit	Values
δ	Skin Depth	cm	0.0468
D_{AWG}	Wire Diameter	cm	0.936
A_w	Base Wire Area	mΩ/cm	263,9
T	Time Period	sec	$5 * 10^{-5}$
t_{on}	Maximum Transistor on Time	sec	$2.5 * 10^{-5}$
P_{o1}	Power out in secondary No.1	kW	10
P_{o2}	Power out in secondary No.2	kW	5
P_T or P_{omax}	Total Secondary Power	kW	12
$I_{pri,max}$	Max Primary Current	amps	7,21
$R_{in,eq}$	Equivalent Input Resistance	ohms	56.25
L	Primary Inductance	μH	351
W-S	Energy Handling Capability	w – s	0.0022
W-S	Energy Handling Capability	w – s	0.0022
K_e	Electrical Condition	-	$5.22 * 10^{-5}$
K_g	Core Geometry	cm ⁵	0.0946
J	Current Density	amps/cm ²	174.15
$A_{pw}(B)$	Primary Wire Area	cm ²	$8.3 * 10^{-3}$
S_{np}	The Required Number of Primary Strands	-	8

N_p	The Number of Primary Turns	<i>turns</i>	112
l_g	The Required Gap	<i>cm</i>	0.0266
<i>mils</i>	The Equivalent Gap in Mils	-	10.472
F	Fringing Flux Factor	-	1.1275
B_{pk}	The Peak Flux Density	<i>teslas</i>	0.382
$\mu\Omega/cm$	The primary, The new $\mu\Omega/cm$	-	131.95
R_p	Primary Winding Resistance	<i>ohms</i>	0.025
N_{s01}	The Secondary Turns	turns	5
$I_{s01(pk)}$	Secondary Peak Current	amps	7.5
$A_{sw01(B)}$	The Secondary Wire Area	cm^2	0.0210
S_{ns01}	Required Number of Secondary Strands	-	3
$(S_{01})\mu\Omega/cm$	The S_{01} secondary, $\mu\Omega/cm$	-	82.2
R_{s01}	The Winding Resistance	<i>ohms</i>	0.0037
N_{s02}	The Secondary Turns	turns	52
$I_{s02(pk)}$	The Secondary Peak Current	amps	15
$A_{sw02(B)}$	The Secondary Wire Area	cm^2	0.00524
S_{ns02}	The Required Number of Secondary Strands	-	1
$(S_{02})\mu\Omega/cm$	The, S_{02} , Secondary, $\mu\Omega/cm$	-	263.9
R_{s02}	The Winding Resistance	<i>ohms</i>	0.0219
K_u	The Window Utilization	-	0.2086
P_{cu}	The Total Copper Loss	<i>watts</i>	360

α	The Regulation, α , For this design	%	0.30
B_{ac}	The AC Flux Density	teslas	0.191
WK	The Watts per Kilogram	$\frac{watts}{kilogram}$	12.4
P_{fe}	The Core Loss	watts	440.43
P_{Σ}	The Total Loss, Core P_{fe} and Copper P_{cu} , in watts	watts	860.43
Ψ	The Watt Density	watts/cm ²	12.4
T_r	The Temperature Rise, $T_r, in, ^\circ C$	$^\circ C$	11.94

The electromagnetic analysis of a discontinuous high-frequency flyback transformer involves several steps. Initially, a mathematical model is utilized to determine various parameters of the transformer. These parameters encompass losses within the transformer (such as copper losses, core losses, and stray losses), as well as the weight of different components (such as the core weight, primary and secondary winding weight), and the dimensions of the core and windings.

Once the parameters are obtained through the mathematical model, the next step involves optimizing them with a focus on improving both the weight and efficiency of the transformer. This optimization process seeks to find the best combination of parameters that minimizes the weight while maximizing efficiency.

Following the optimization process, the design of the transformer is then converted into the ANSYS/Maxwell program. This program is specifically designed for the design, analysis, and simulation of electrical machines, including the electromagnetic and thermal aspects.

Within the ANSYS/Maxwell program, detailed design steps are performed to determine the appropriate dimensions for the transformer's core and windings. This involves considering factors such as the desired performance, operating conditions, and constraints. By utilizing the

capabilities of ANSYS/Maxwell, the design process can be accurately simulated and analyzed to ensure the transformer meets the required specifications.

To analyze the high-frequency flyback transformer in a discontinuous mode, the ANSYS/Maxwell simulation program is employed. The process begins by selecting the appropriate simulation type, which in this case is magnetic-transient. This choice is made because transformers are electromagnetic devices, and the analysis results are time-dependent.

Next, a 3D representation of the transformer's geometry is created within the program. This involves modeling the core and windings as solid objects, capturing their physical dimensions and configurations. The designed transformer model is shown in Figure 4.27.

After the geometry is established, the materials used in the transformer are assigned within the simulation. This step ensures an accurate representation of the electromagnetic properties and characteristics of each component.

To define the simulation boundaries, specific regions, and conditions are specified. This enables the program to simulate the behavior of the transformer under specific operating conditions, taking into account factors such as the surrounding environment and external influences.

Excitations and coil terminals are then assigned to the transformer model, representing the input and output connections of the windings. These define the electrical inputs and outputs that drive the transformer's operation.

The next step involves mesh operations, where the transformer model is divided into discrete elements or cells. This discretization facilitates numerical computations and allows for accurate calculations of electromagnetic fields within the transformer.

Once the meshing is complete, the solution setup is determined. This includes specifying various parameters and settings related to the simulation, such as solver options, convergence criteria, and time step intervals.

The simulation is then executed, and the program calculates the time-dependent behavior of the transformer. This analysis involves solving the electromagnetic equations governing the transformer's operation, taking into account the specified boundary conditions and excitations.

After the simulation completes, the results and diagrams are obtained, providing valuable insights into the transformer's performance. These results include various electromagnetic quantities, such as magnetic fields, flux densities, and currents, which can be visualized and analyzed to evaluate the transformer's behavior and efficiency.

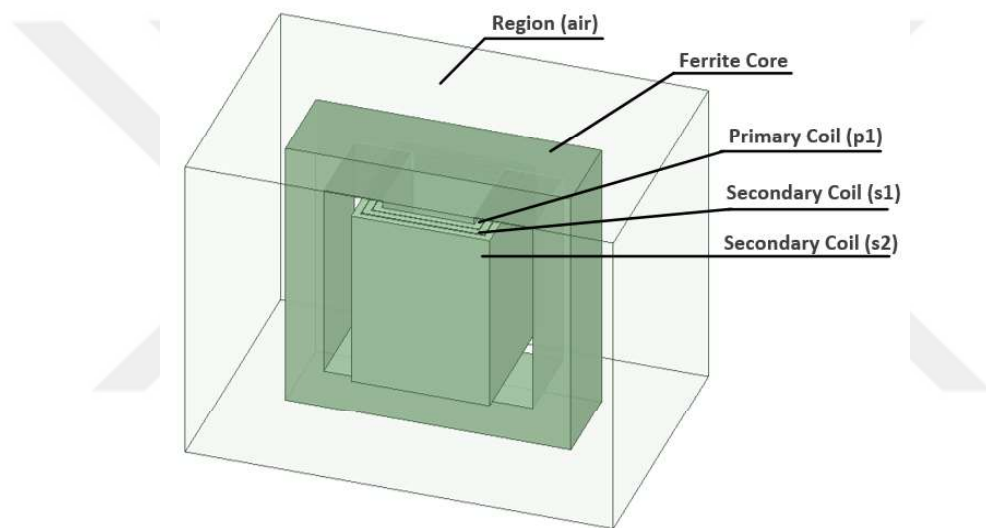


Figure 4.27. ANSYS Maxwell design of flyback high-frequency transformer

The design of the transformer progresses by creating a 3D model using Maxwell-3D within the ANSYS/Maxwell simulation program. This modeling step involves accurately representing the transformer core, as well as the primary and secondary windings.

By utilizing the capabilities of Maxwell-3D, the physical dimensions, and configurations of the transformer's core and windings are carefully designed. This allows for a detailed and precise representation of the transformer's geometry.

The resulting 3D model of the transformer, as presented in Figure 4.28, provides a visual representation of how the core and windings are structured within the design. This representation aids in understanding the physical layout and arrangement of the components.

By employing Maxwell-3D for the transformer design, the simulation program facilitates the analysis of electromagnetic fields and behaviors within the transformer. This enables comprehensive evaluations of factors such as magnetic fields, flux densities, and currents, offering insights into the performance and efficiency of the transformer design.

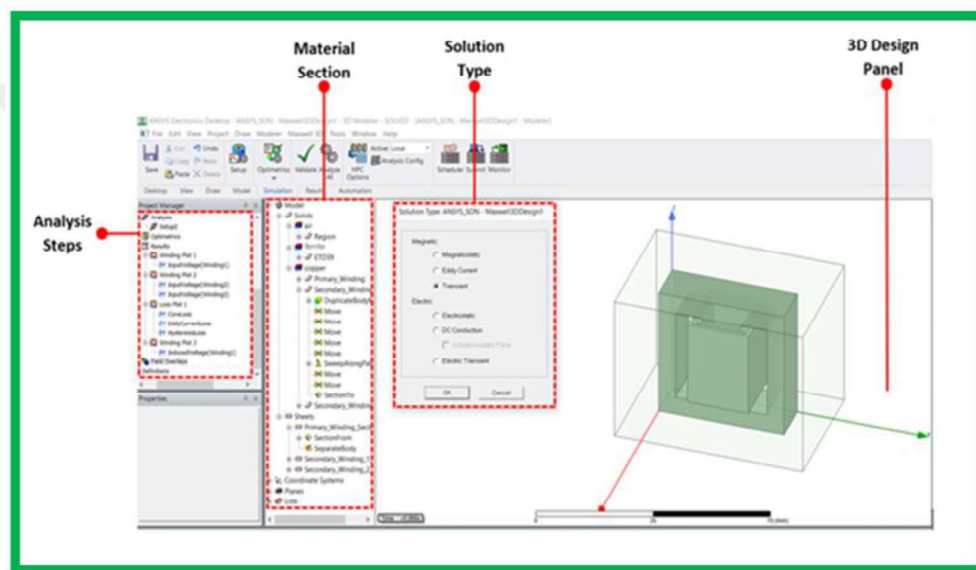


Figure 4.28. ANSYS/Maxwell 3D design operation panel

Following the modeling of the transformer core and windings, the next step involves selecting the appropriate material types for these components. This selection is carried out using the material property panel within the ANSYS/Maxwell simulation program, as depicted in Figure 4.29.

The material property panel provides a user-friendly interface for specifying the electromagnetic properties and characteristics of the transformer's core and windings. By assigning the correct material types, the program can accurately simulate the behavior of these components under different operating conditions.

Careful consideration is given to factors such as magnetic permeability, electrical conductivity, and other relevant material properties when making the material selections. These properties greatly influence the electromagnetic performance and efficiency of the transformer.

Through the material selection process, the ANSYS/Maxwell program ensures that the simulated transformer model reflects the real-world behavior of the chosen materials, allowing for accurate analysis and prediction of the transformer's electromagnetic response.

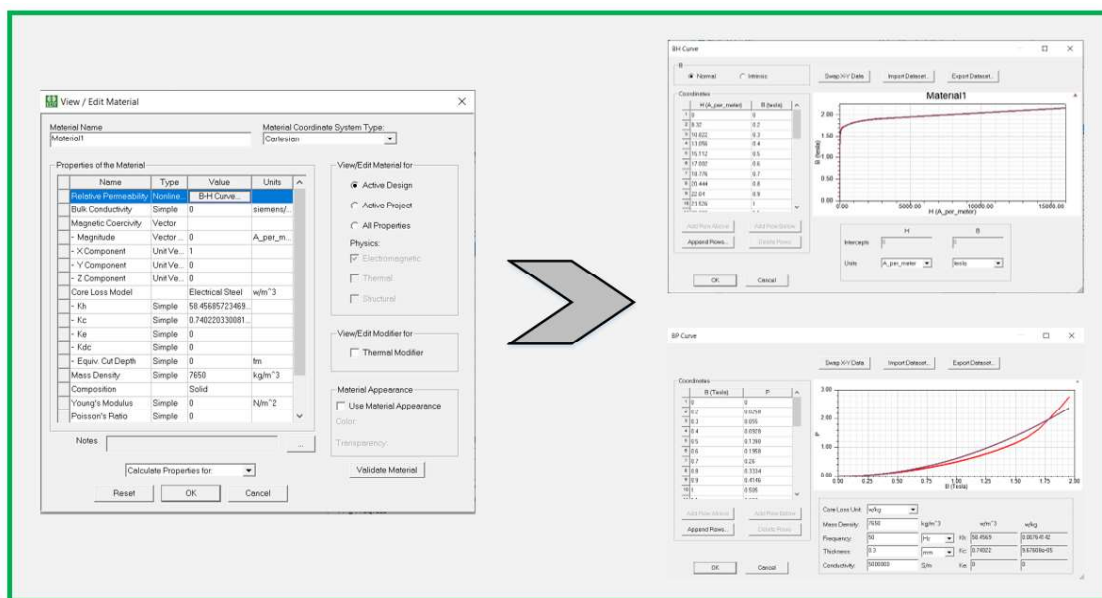


Figure 4.29. Creating specific material properties and determining B-H and B-P curves of the core

During the design process, the core material for the transformer is chosen to be ferrite. This selection is based on the characteristic features of ferrite. Ferrite is known for its specific electromagnetic properties that make it suitable for transformer cores.

Additionally, the primary and secondary windings of the transformer are selected to be made of copper material. Copper is a commonly used material for transformer windings due to its excellent electrical conductivity.

After assigning the appropriate materials to the core and windings, the next step involves exciting and energizing the primary and secondary windings based on the specified transformer parameters. These parameters include the turn ratio, ohmic resistance, and excitation terminal voltage values. By considering these factors, the windings are designed to meet the desired electrical characteristics of the transformer.

The assignment of excitations to the windings is depicted in Figure 4.30, providing a visual representation of how the windings are connected and energized within the transformer design.

Furthermore, to account for the specific characteristics of the transformer, the air material is assigned as a region or boundary. This is done to accurately represent the presence of air gaps or regions within the transformer structure.

By assigning the appropriate materials and excitations, the ANSYS/Maxwell program enables a comprehensive analysis of the transformer's electromagnetic behavior, considering factors such as magnetic fields, electrical currents, and voltage distributions.

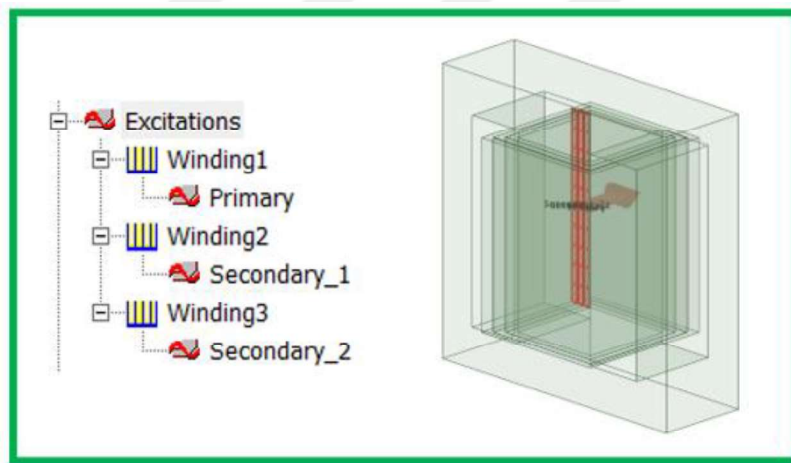


Figure 4.30. Assignment of excitation to the coil terminals of the transformer

In finite element analysis (FEA), it is essential to ensure that the results obtained are independent of the mesh size. A convergence study is conducted to achieve this, which not only accelerates the solution process but also enhances the accuracy of the results.

FEA is a valuable method, particularly when making critical and costly decisions. Therefore, it is crucial to establish the verifiability of the results obtained through finite element analysis.

To demonstrate the accuracy of the results, the simulation study incorporates a mesh operation. This operation involves dividing the designed transformer into a large number of particles, specifically around 80,000 to perform electromagnetic analyses on each particle. By employing this mesh operation, the designed model undergoes a more precise and detailed analysis.

The mesh operation, as shown in Figure 4.31, visually represents how the transformer model is discretized into numerous particles. This process ensures that every particle is electromagnetically analyzed, leading to a more accurate evaluation of the transformer's electromagnetic behavior.

Through the mesh operation, the ANSYS/Maxwell simulation program provides a comprehensive analysis, capturing intricate details of the electromagnetic fields, flux densities, and currents within the transformer model. This aids in obtaining reliable and trustworthy results, bolstering confidence in the analysis outcomes.

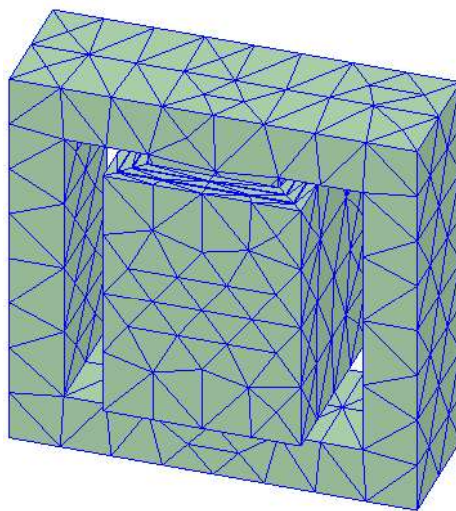


Figure 4.31. Mesh operation applied to the core and windings of the transformer

Upon completing the preceding processes, the transformer can undergo analysis to study its electromagnetic behavior. The analysis time is determined by utilizing the setup panel in the ANSYS/Maxwell program.

The setup panel allows for the specification of various parameters that are crucial for obtaining an accurate electromagnetic transient analysis of the transformer. Among these parameters, the most important ones are the analysis solution step type, start time, and stop time. These parameters directly influence the precision of the calculated currents and voltages within the transformer.

To ensure the fidelity of the analysis results, careful consideration is given to selecting appropriate values for these parameters. In this study, small ranges are chosen for the analysis step size, aiming to obtain pure signals devoid of unnecessary noise or distortion.

The analysis settings for the transformer, as depicted in Figure 4.32, visually illustrate the chosen parameters and settings within the setup panel. This allows for a clear representation of how the analysis time is defined and how the analysis solution step is configured.

By appropriately setting up the analysis parameters, the ANSYS/Maxwell program facilitates the accurate simulation and evaluation of the transformer's electromagnetic transient behavior. This enables the examination of key electrical quantities, such as currents and voltages, throughout the defined analysis time range.

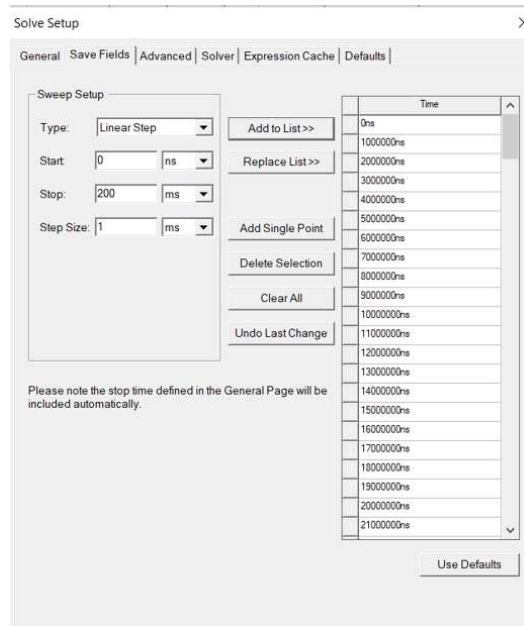


Figure 4.32. Analysis settings of the high-frequency transformer in ANSYS/Maxwell

Subsequently, the analysis results are obtained based on the defined analysis time. The electromagnetic transient analysis reveals the following findings:

When a voltage is applied to the primary side of the transformer coils, it gives rise to a dynamic magnetic field within both the transformer core and the coils. This changing magnetic field induces a voltage across the ends of the primary coil. Consequently, the primary current and magnetic flux within the transformer undergoes variations.

The alterations in the magnetic flux, resulting from the changing primary current, then lead to the induction of a voltage in the secondary side of the transformer coils. This voltage induction is a consequence of the transformer's core and winding interactions.

Figure 4.33 displays the analyzed transformer core, illustrating the distribution of the magnetic flux within the core. This visual representation provides insights into how the magnetic flux is distributed and influenced by the applied voltage and current variations.

Through the electromagnetic transient analysis, the ANSYS/Maxwell program enables a comprehensive examination of the transformer's behavior during dynamic operations. This analysis sheds light on crucial aspects such as voltage induction, current variation, and the resulting magnetic flux distribution. Such insights facilitate a deeper understanding of the transformer's performance and aid in making informed decisions regarding its design and operation.

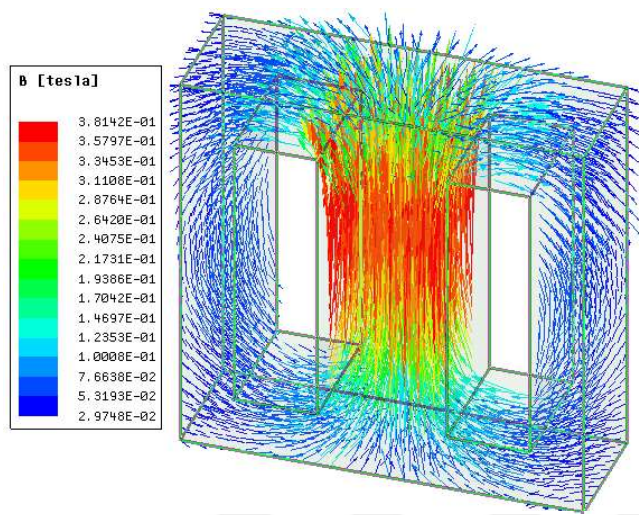


Figure 4.33. Magnetic field distribution on the transformer core

In the analysis of the transformer, a comprehensive power loss analysis is conducted, encompassing both the transformer core and coils. Figure 4.34 presents the distribution of power losses within the transformer.

When a variable current flows through a coil wound on a core, it induces a voltage in the core. This voltage, in turn, gives rise to the formation of multiple closed loops of current within the core. This phenomenon, known as eddy currents or Foucault currents, occurs not only on the surface of the core but also inside it.

Each closed current path, formed as tiny closed rings, carries a current strength directly proportional to the induced voltage. The intensity of these currents is inversely proportional to the electrical resistance of the respective current path. However, these eddy currents result in

the generation of heat within the core, leading to energy loss. This heat generation due to eddy currents contributes to what is known as eddy current loss.

Additionally, certain ferromagnetic materials, such as iron, exhibit a temporary or permanent magnetization when exposed to an external magnetic field. This magnetization opposes the magnetic field present in the transformer, resulting in energy loss in the form of heat. This type of loss is referred to as hysteresis loss. Hysteresis loss occurs as a result of molecular friction when the core's magnetic direction changes in response to the frequency of the magnetic field.

Furthermore, copper loss is a term commonly used to describe the heat generated by the flow of electric current through transformer windings or conductors in other electrical devices. It accounts for the resistive heating effect and contributes to overall power loss in the transformer.

The electromagnetic transient analysis of the high-frequency transformer provides a comprehensive understanding of these power losses. It allows for a detailed examination and quantification of eddy current loss, hysteresis loss, and copper loss. By considering these losses, the analysis results contribute to assessing the transformer's overall efficiency and addressing potential issues such as overheating.

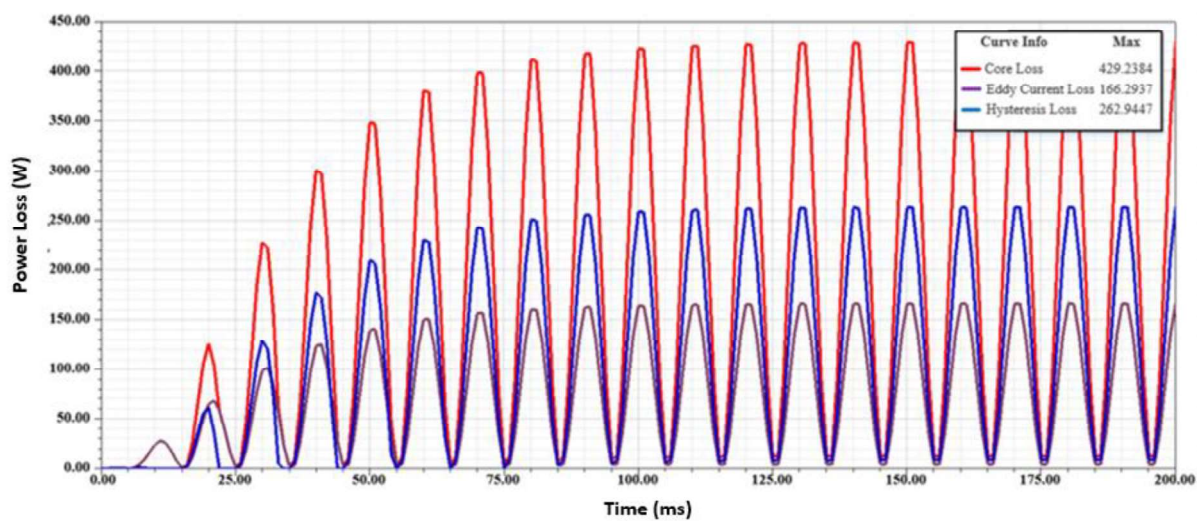


Figure 4.34. Power losses of the HF flyback transformer

Upon examining the losses in the HF transformer, it is observed that the total power loss amounts to approximately 859 W. Comparatively, the mathematical model predicts a power loss of about 860 W. This close agreement between the simulation and mathematical results validates the realism of the analysis.

The accuracy of the analysis can be attributed to several factors, including the careful selection of the transformer's core and coils, the consideration of B-H and B-P characteristics, the meticulous design of dimensions in a one-to-one fashion, and the implementation of the mesh operation.

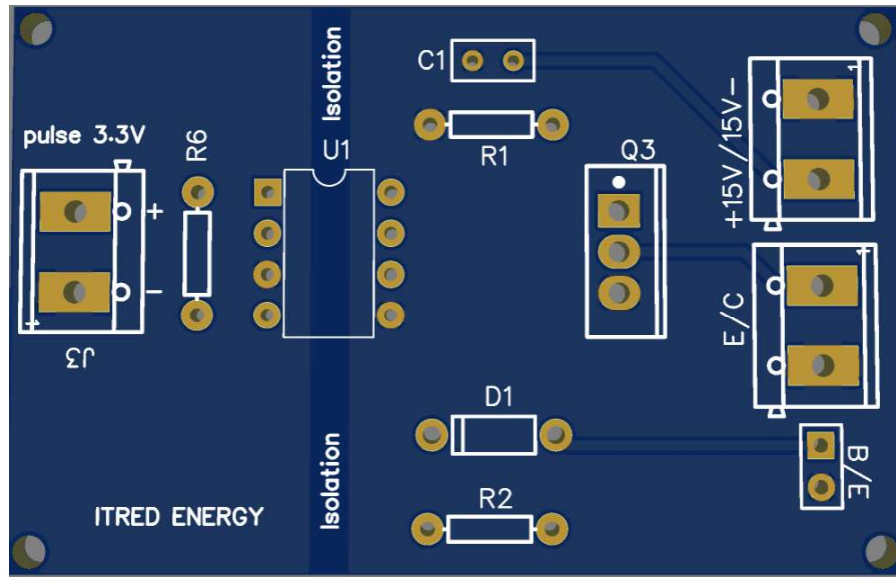
The HF transformer has a total power of 12 kW, and based on the power loss analysis, it achieves an efficiency of 92.8%. This efficiency measure quantifies the transformer's ability to convert electrical power with minimal losses.

Overall, the consistency between the simulated and mathematical results affirms the reliability and effectiveness of the HF transformer design and analysis process.

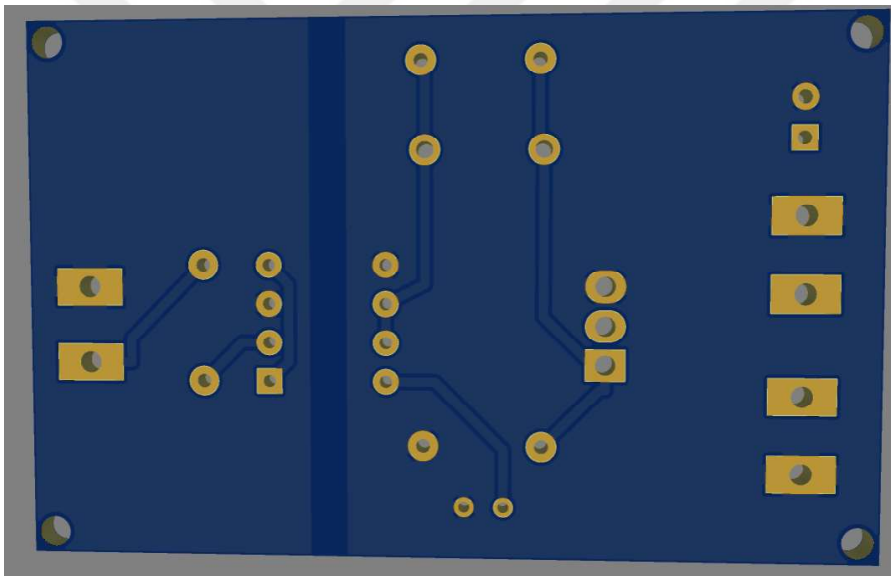
4.2.3. PWM Inverter (DC-AC)

To ensure high-speed operation of the IGBTs, the HCPL-3120 IC is incorporated into the design for PWM inverter driving. This additional design facilitates the driving process and enhances the performance of the IGBTs. It is crucial to use an isolated driver model when operating IGBTs at high current and voltage levels to protect the control card from reverse current flow and prevent potential damage. Each IGBT is driven with gate signals having a 50% duty cycle.

To conduct the driver test circuit for the control card, a dedicated PCB card is designed, implementing an isolated gate driver. This PCB card is utilized on the experimental table for carrying out the control card driver test. Figure 4.35 showcases the designed gate driver model for one module, providing a visual representation of the implemented design.



(a)



(b)

Figure 4.35. Isolated gate driver circuit for one IGBT module a) front view b) rear view

The gate signals necessary to drive the IGBTs are presented in Figure 4.36. The isolated driver circuit, as shown in the circuit diagram, enables the generation of these gate signals. It is noteworthy that the signals exhibit a fast switching speed, achieving a response time of less than 600ns.

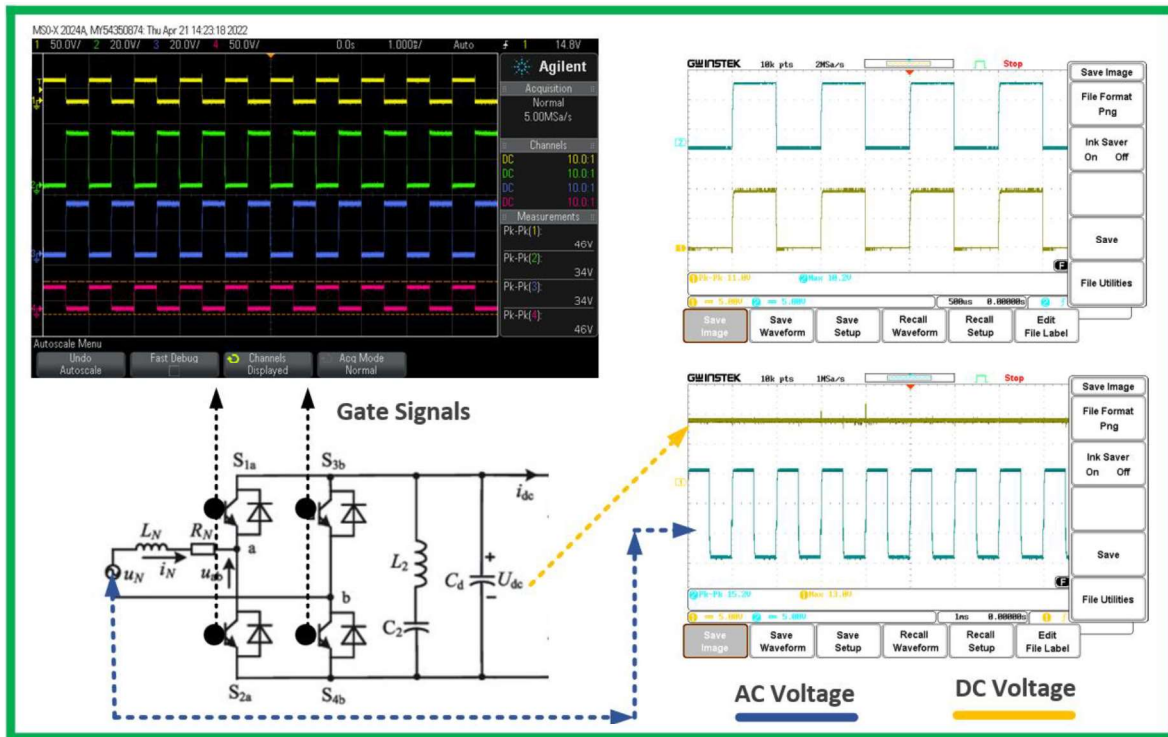


Figure 4.36. Isolated gate PWM inverter structure gate signals, dc input and ac output signals

4.2.4. Full Bridge Rectifier (AC-DC)

The full wave rectifier is an electronic circuit that converts both the positive and negative halves of each waveform cycle into a pulsating DC signal. This conversion is achieved using four rectification diodes.

In a previous tutorial on power diodes, we discussed methods for reducing the ripple or voltage variations in a direct DC voltage by incorporating smoothing capacitors across the load resistance.

The Full Wave Bridge Rectifier, on the other hand, utilizes four individual rectifying diodes arranged in a closed loop configuration, forming a "bridge." This configuration enables the desired output to be produced.

The main advantage of this bridge circuit is that it eliminates the need for a special center-tapped transformer, thereby reducing the size and cost of the circuit. Instead, a single secondary

winding is connected to one side of the diode bridge network, while the load is connected to the other side.

By using this full wave bridge rectifier configuration, both positive and negative portions of the input AC waveform are effectively rectified into a pulsating DC output. This allows for a more efficient utilization of the AC power source and facilitates the generation of a steady DC voltage for various applications.

A full wave rectifier, specifically a Bridge Rectifier, can be implemented using four rectifier diodes. Unlike other rectifier designs, it does not require a center tap in the transformer. As the name implies, the circuit incorporates a bridge configuration.

The four diodes in the circuit are interconnected to form a closed-loop bridge arrangement. This arrangement enables the rectification of both the positive and negative halves of the input AC waveform. The resulting output is a pulsating DC.

One advantage of the Bridge Rectifier is its cost-effectiveness and compact size. The absence of a center tap transformer reduces the overall complexity and size of the circuit. This makes it a preferred choice for applications where space and cost are critical factors.

To illustrate the classical full-wave diode converter structure, refer to Figure 4.37. This visual representation provides a clear depiction of the circuit configuration and helps in understanding the overall design of the Bridge Rectifier.

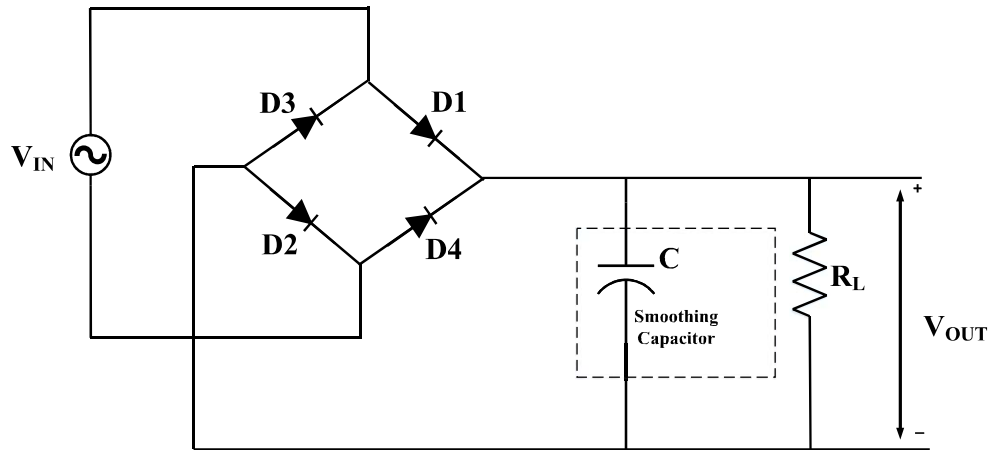


Figure 4.37. The circuit diagram of the full bridge diode rectifier

In the Bridge Rectifier circuit, four diodes, namely $D1$, $D2$, $D3$, and $D4$, are utilized. Depending on whether the upper half-cycle or lower half-cycle of the input signal is fed to the circuit, two diodes conduct simultaneously, either $D1$ and $D3$ or $D2$ and $D4$.

The full-wave rectifier, operating on both halves of the input signal simultaneously, exhibits bidirectional characteristics. This means that it can handle signals flowing in either direction.

The full-wave rectifier can be constructed using either two or four diodes, depending on the specific design requirements.

It's important to note that the peak reverse voltage of the full-wave rectifier is twice the value of the input voltage supplied to the circuit.

Taking into account the features of the proposed full bridge converter structure, it is equipped with a surge current NTC (Negative Temperature Coefficient) to inhibit circuit inrush current, ensuring more stable operation. Additionally, it incorporates fuses for short-circuit protection. The rectifier board also includes a power indicator function and screen printing marks on the board, facilitating ease of operation.

In the presented design, the rectifier board is specifically designed and manufactured for an input voltage range of 0-220VAC, an output voltage of 450VDC, and an output current of 10A.

To visualize the converter produced based on these specifications, refer to Figure 4.38. The figure illustrates the physical realization of the rectifier board, showcasing the components and layout of the circuit.

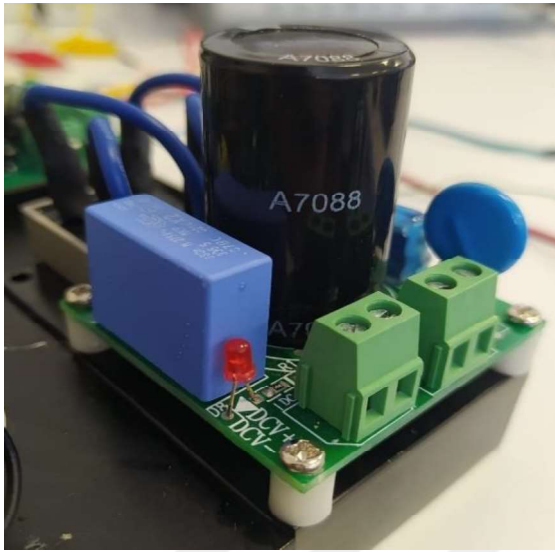


Figure 4.38. Full bridge diode rectifier used for experimental study

4.2.5. Experimental Test Results of the SST

Figure 4.39 showcases a soft switching smart transformer model with a power rating of 10 kVA and a voltage rating of 380-380 V. This model incorporates various components, including a rectifier circuit, a DAB (Dual Active Bridge) converter, and a PWM (Pulse Width Modulation) inverter.

The primary function of this smart transformer is to convert and regulate the variable AC voltage generated by a solar PV power plant, ensuring a user-defined output voltage. The rectifier circuit plays a crucial role in converting the incoming AC voltage to DC, which is then fed into the DAB converter. The DAB converter, utilizing advanced soft switching techniques, efficiently transforms the DC voltage to the desired output voltage.

To achieve precise voltage control, the PWM inverter is employed. It modulates the width of the output pulses to regulate the average voltage value, resulting in a smooth and stable output voltage waveform. This smart transformer model is designed to enable user-defined voltage regulation in the presence of variable AC voltage inputs from solar PV power plants. By effectively managing and controlling the voltage conversion process, it contributes to the overall stability and reliability of the power system.

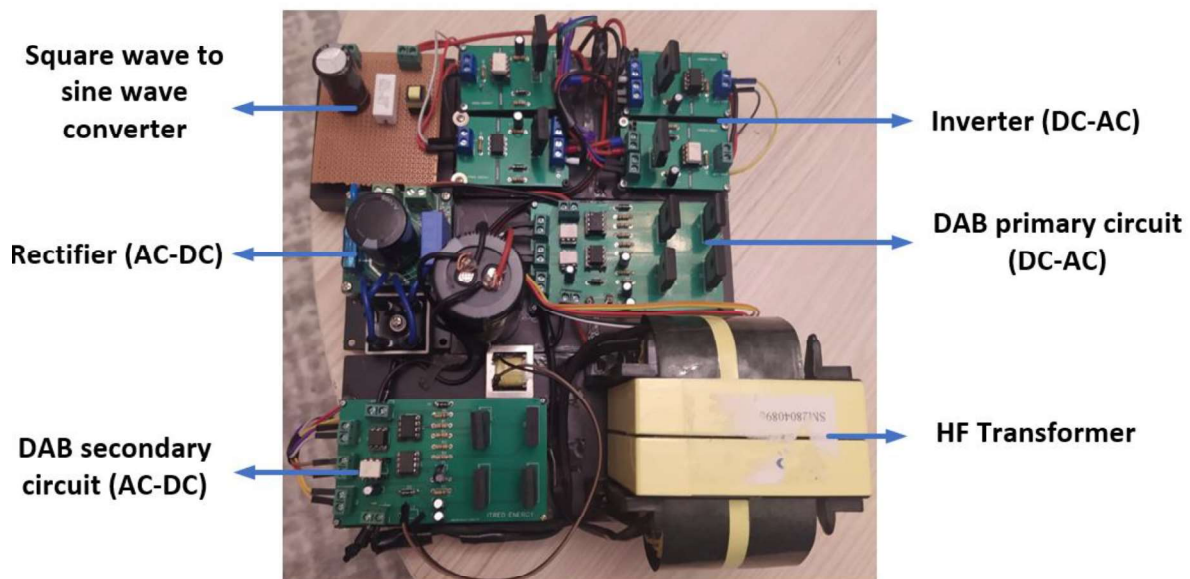


Figure 4.39. Designed and produced smart transformer models for smart power grids

The power measurements and signal conditioning unit, shown in Figure 4.40, consists of amplifiers and supply voltage parts for converters and an inverter. The output power of the solar PV power plant serves as the system's principal voltage source.

The primary voltage is stepped down to 12 Vac using a 10W, 220/12 Vac power transformer to provide the necessary voltage changes. With the aid of a diode rectifier circuit, this 12 Vac voltage is further transformed to 5 Vdc. The 5 Vdc voltage is also used to create numerous isolated dc-dc converters that give the necessary supply voltages. For the 5 Vdc supply voltages to be increased to 50 Vdc, the boost converter is essential. The converters and inverters are then powered by these elevated supply voltages.

Using a current/voltage measurement sensor, the reference voltage of 220 Vac is changed to 5 Vac in the measurement and signal conditioning circuitry. Through the use of a signal scaling circuit, this 5 Vac signal is further converted to a 3.3 Vac unipolar signal appropriate for the DSP control board. Current/voltage measurements and PWM control are made possible by the DSP control card, which is integrated with MATLAB/Simulink. Overall, this configuration allows for accurate power measurements, signal conditioning, and supply voltage generation, enabling efficient control and operation of the converters and inverter.

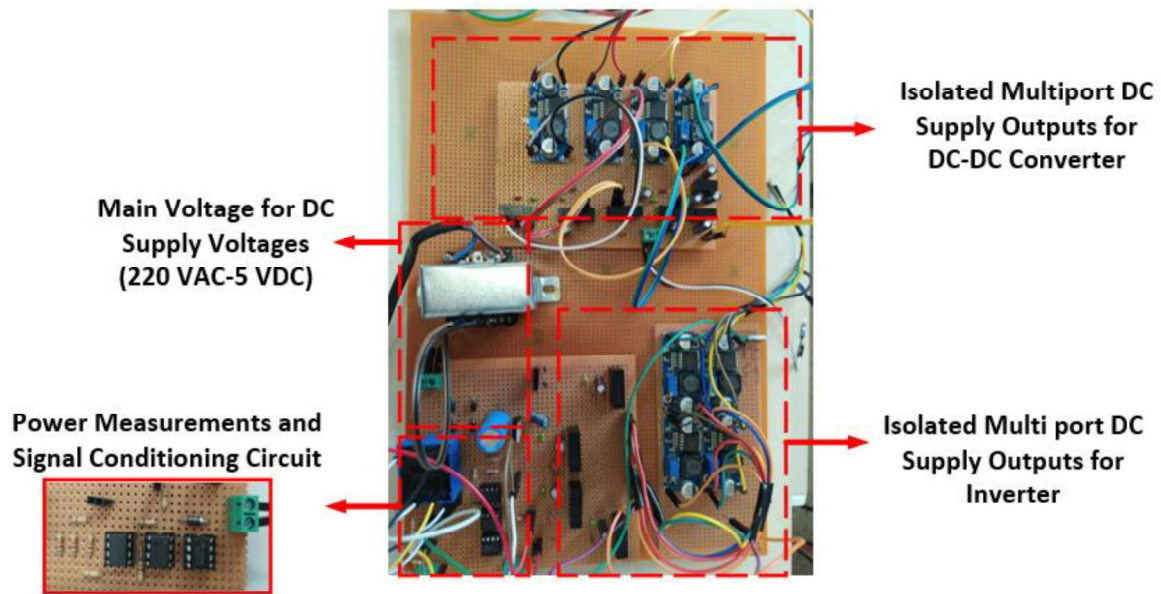
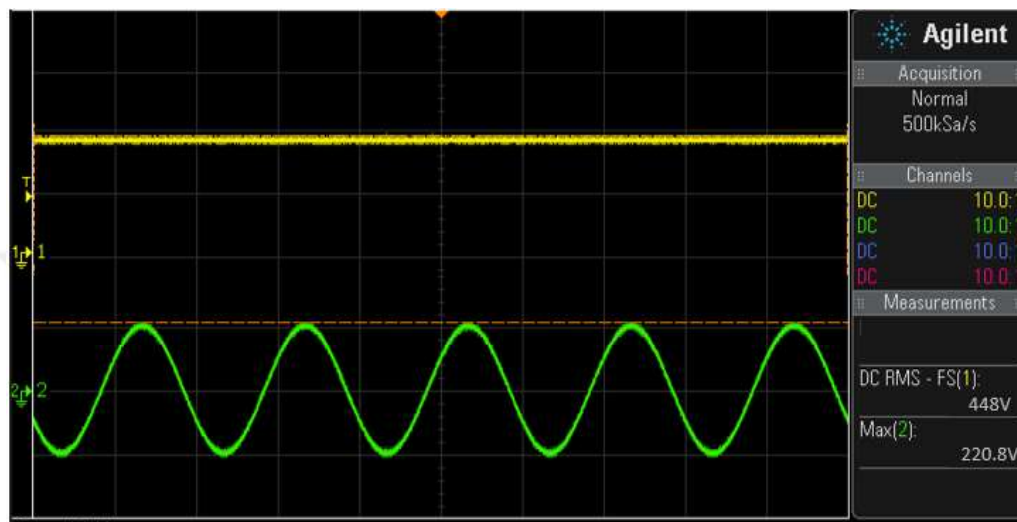


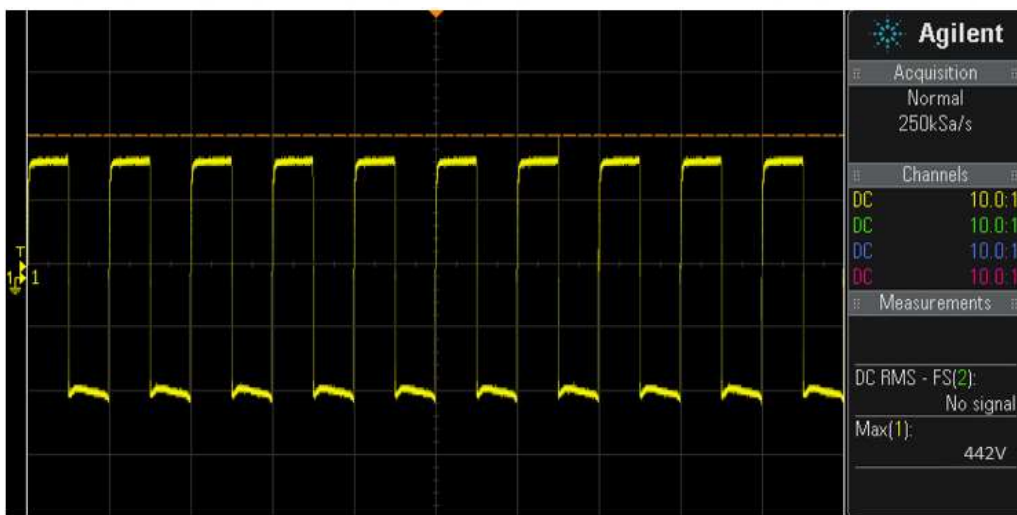
Figure 4.40. The measurement circuit, power supply circuit and signal conditioning circuit for SST

Using a 220 Vac supply voltage, the SST performs an AC-DC conversion, resulting in a voltage of 450 Vdc on the converter side. This 450 Vdc voltage serves as the basis for the final conversion voltage in the inverter section, which is adjusted to 220V based on the transformer's conversion ratio. To assess the efficiency of the system, the conversion efficiencies of both the DAB converter and the SS T are calculated. At a power rating of 10 kVA, the DAB converter exhibits an efficiency of 98.6%, while the ST achieves an efficiency of 96.7%.

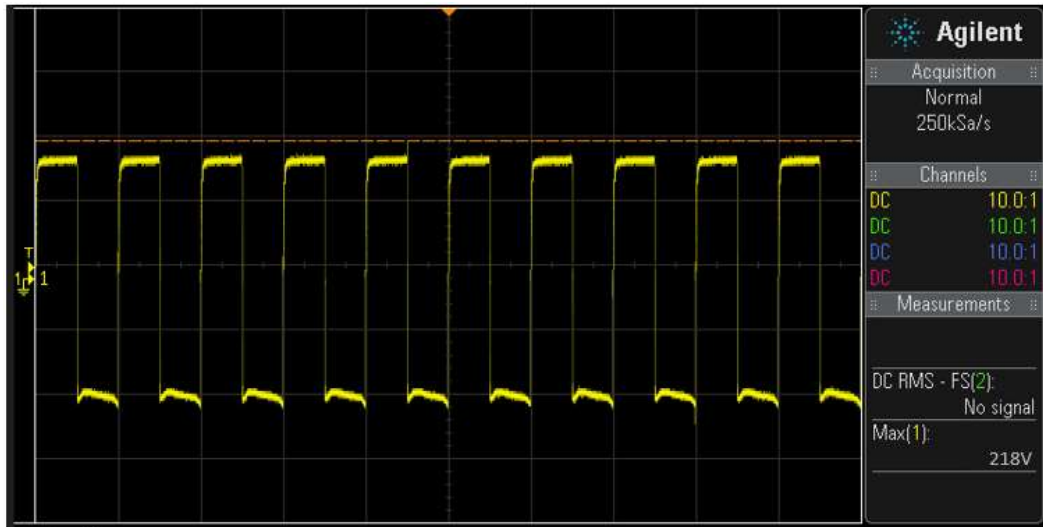
The solar PV power plant provides the input voltage, which is carefully controlled by the smart transformer. Through this control mechanism, the variable input voltage is managed, and the desired power is efficiently transmitted to the end-user smart grid. Figure 4.41 presents the waveforms of the input and output voltages sequentially, showcasing the characteristics and behavior of the voltage signals throughout the conversion process.



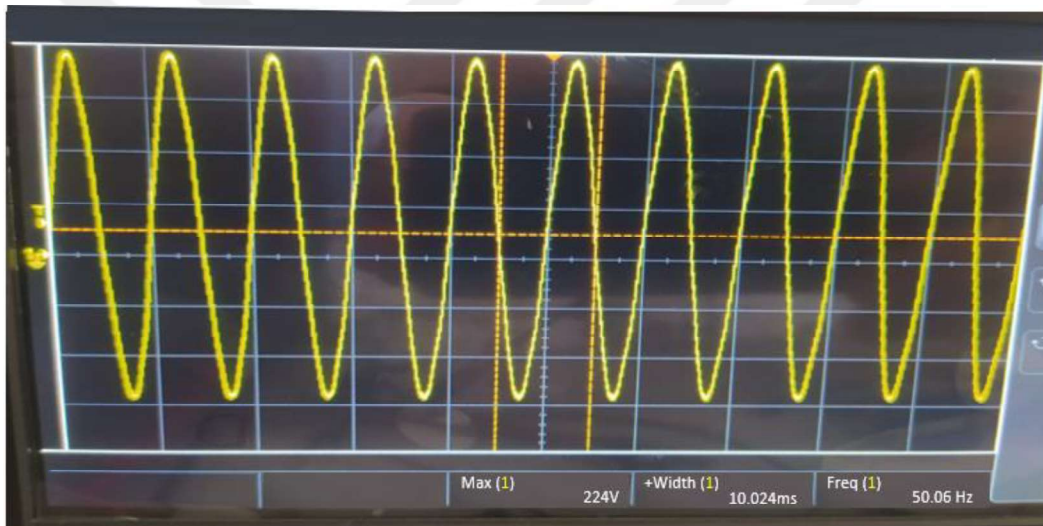
(a)



(b)



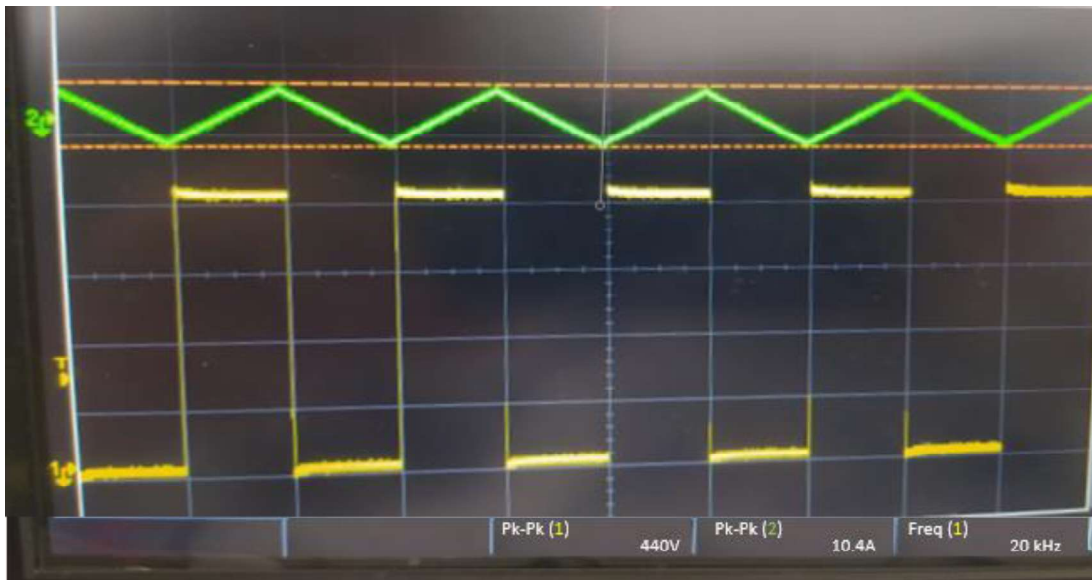
(c)



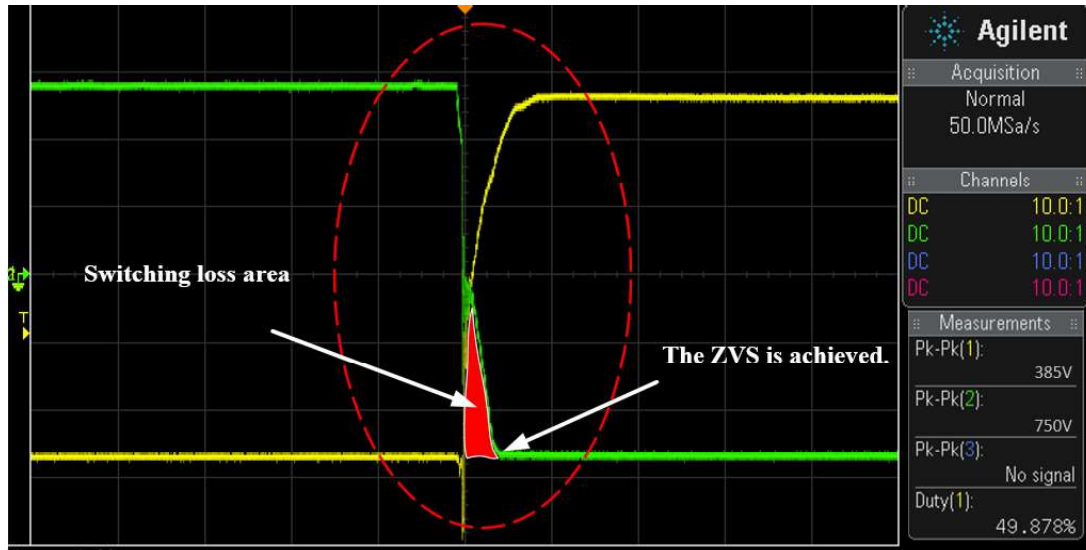
(d)

Figure 4.41. Power waveforms resulting from experimental tests a) the rectifier terminal voltage and converted output voltage b) voltage seen at the primary side of the HF transformer c) voltage seen at the secondary of the HF transformer d) inverter end-user voltage

Soft switching refers to a technique used in dc-to-dc converters to minimize the switching losses associated with turning power switches on or off. By implementing soft-switching capability, the proposed circuit in this context achieves full-range ZVS, which surpasses conventional solutions that have limited ZVS capabilities in DAB converter. The experimental findings are shown in Figure 4.42, with a special emphasis on the ZVS gain seen over the primary side of the HFT. The ZVS gain measurement corresponds to a primary side current of 10.4 A. The significance of achieving ZVS in the proposed circuit lies in its ability to reduce switching losses and enhance overall efficiency. This improvement in switching characteristics contributes to a more efficient and optimized operation of the dc-to-dc converter.



(a)



(b)

Figure 4.42. ZVS capability of the converter a) primary side output of the DAB, b) minimized switching loss area

The high-frequency oscillation problem caused by capacitances in the transformer and the high rate of change of voltage (dv/dt) in the DAB converter becomes more pronounced as the switching speed increases. A theoretical solution to this problem entails setting the DAB dv/dt excitation's zero crossing points at the oscillation frequency. This can be achieved by connecting parallel capacitors to the power components, effectively altering the DAB dv/dt .

The adding of these capacitors not only reduces oscillation losses in the transformer but also creates a more favorable soft-switching environment. In addition to resolving the high-frequency oscillation issue, this technique boosts overall effectiveness. Rapid rise times in the voltage pulse can lead to high dv/dt , resulting in voltage spikes and ringing effects on the pulse's leading edge. This ringing serves as a source of high-frequency EMI and can potentially damage loads if the peak voltage is sufficiently high. To control the dv/dt in the proposed converter, a resonant capacitor is utilized. The DAB, on the other hand, only controls the dv/dt within its ZVS range, while the inverter's dv/dt remains uncontrolled.

Devices made of medium-voltage silicon carbide can exhibit uncontrolled dv/dt rates ranging from 10-100 kV/s. Experimental data confirms a dv/dt rate of 6.8 kV/ μ s under a load current of 10.4 A with a voltage of 344Vac. The dv/dt capability is depicted in Figure 4.43, providing insights into the behavior and performance of the system.



Figure 4.43. The dv/dt capability at the reverse switch positions

The performance of the SST in handling sudden voltage dips and surges in the generated voltage by the 12.5 kW output power on-grid PV solar system is demonstrated in Figure 4.44. The measurements of current and voltage at the output of the ST are presented.

Upon analyzing the signal fluctuations, it is evident that the SST effectively maintains a constant output voltage for the end user, despite the variations in the input power. Additionally, no noticeable harmonics are observed in the load current, indicating a smooth and stable power delivery to the connected devices.

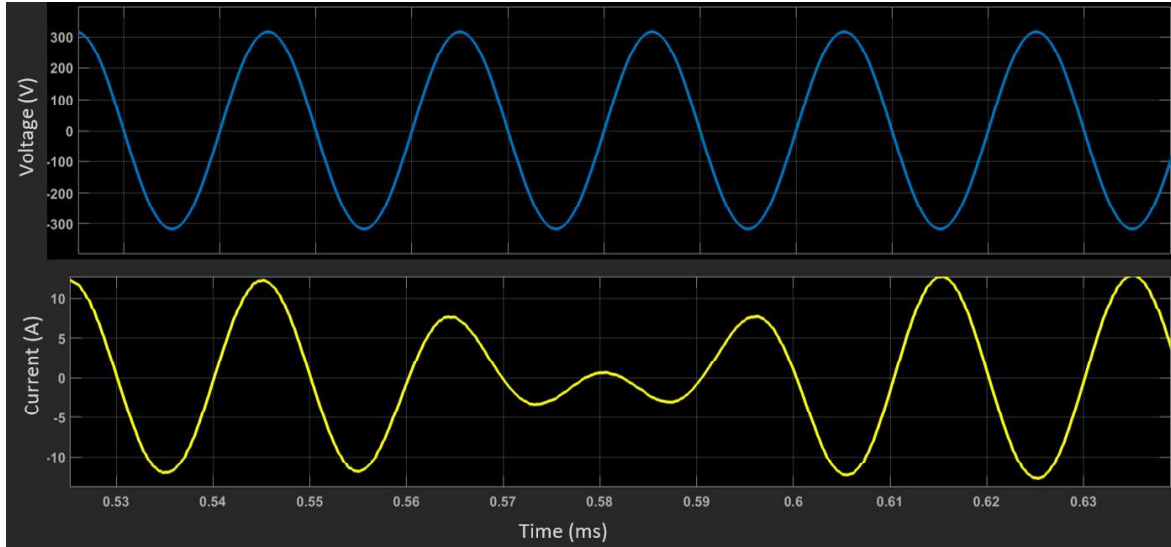


Figure 4.44. The output power of the SST under the sudden sag and swell of the Solar PV power

5. FINDINGS AND DISCUSSION

Advancements in controllability are required to successfully integrate renewable sources into the power grid as the demand for renewable energy keeps rising. Power flow management, reactive power compensation, harmonic control, there level conversion, and simplifying the integration of renewable energy sources are just a few of the applications for which SSTs provide potential solutions. The isolation stage is essential to SSTs since it increases the system's effectiveness and controllability. Numerous studies have been done on the various DC-DC power converter topologies utilized in SST applications. The selection of the best dc-dc power converter architecture specifically for grid-connected solar PV power plants is not covered in much literature, nevertheless.

The recommended unidirectional DAB power converter runs at a constatnt resonant frequency with a 50% duty cycle under low load situations. Increasing the magnetizing inductance (L_m) can reduce backflow power and switch on-off current. However, the magnetizing inductance should be sufficiently small to achieve ZVS operation. It is also important to strike a balance between extending the deadband period and maintaining an optimal ZVS requirement.

While extending the deadband period reduces backflow power, it also decreases the effective time for energy transfer. The performance of ZVS operation in the DAB is determined by finding the right balance between magnetizing inductance and dead time. Switching power losses can be minimized by expanding the ZVS range and employing a suitable control technique. To reduce circulating power flow and enhance system performance, it's essential to choose power switches and HF transformer characteristics carefully.

6. CONCLUSIONS

The performance of a 10 kVA 380/380 Vac SST was examined in experimental study for decreased dv/dt , ZVS achieve, and efficiency of power conversion at various voltages. Through experimental data, a 344Vac and 10.4 A load current was used to obtain a dv/dt rate of 6.8 kV/us. At the primary side of the HF transformer, the ZVS gain was seen, and switch voltage measurements were made to validate its effectiveness. The efficiency of power conversion was estimated to be 96.7% when operating at full load. When the ZVS gain was taken into account, the dynamic analysis showed that the switching losses were almost optimal. Switch losses, however, were observable in the experimental investigation at maximum load capacity. In the experimental investigation, the SST's output power was determined to be 9.66 kW at 750 Vac peak-to-peak voltage and 12.89 A current, whereas in the dynamic analysis, it was 9.9 kW at 750 Vac voltage and 13.2 A current. In the experimental investigation and the dynamic analysis, the SST's efficiency was 96.67% and 99%, respectively. The experimental analysis revealed transmission losses and HF transformer losses in addition to switching losses. Through controlled dv/dt and full-range ZVS capability, the recommended SST showed the potential for less electromagnetic interference. Additionally, the control system performed better at preserving dc voltage stability and guaranteeing safe operation. Voltage balancing between the converter modules was accomplished by the SST using its beneficial control approach in both steady-state and dynamic situations.

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