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Medium-Voltage Medium-Frequency Transformer for Green Hydrogen Applications

Reza Mirzadarani
TU Delft

Medium-Voltage Medium-Frequency Transformer for Green Hydrogen Applications

Medium-Voltage Medium-Frequency Transformer for Green Hydrogen Applications

Dissertation

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at Delft University of Technology
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chair of the Board for Doctorates
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Thursday 18 June 2026 at 10:00

by

Reza MIRZADARANI

This dissertation has been approved by the promotor.

Composition of the doctoral committee:

Rector Magnificus,	chairperson
Prof. ir. P.T.M. Vaessen,	Delft University of Technology, <i>promotor</i>
Prof. dr. ir. P. Bauer,	Delft University of Technology, <i>promotor</i>
Dr. ir. M. Ghaffarian Niasar,	Delft University of Technology, <i>promotor</i>

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To my lovely wife.

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SUMMARY

This thesis focuses on the design and development of a medium-voltage, medium-frequency solid-state transformer (SST) for large-scale green hydrogen production. The work is motivated by the need to improve the efficiency, compactness, and controllability of power conversion systems that connect renewable energy sources, such as offshore wind, to electrolyzers. Conventional 50/60 Hz transformers and rectifiers are well-established but are often heavy, bulky, and limited in performance at high power levels. Solid-state transformers, operating at higher frequencies, offer the potential to reduce system size and weight while improving functionality and efficiency.

The research is conducted within the *FlexH₂* project, a sponsored program that investigates new concepts for integrating offshore wind energy with onshore hydrogen production. The work presented in this thesis contributes to Work Package 2, which focuses on developing an SST-based interface between the medium-voltage AC network and the DC supply of large electrolyzers. Several SST topologies, including the Modular Multilevel Converter (MMC), Resonant Modular Multilevel Converter (MMR), and Input-Series Output-Parallel (ISOP) structures, are analyzed and compared in terms of efficiency, weight, losses, and system complexity.

The main focus of the thesis is on the medium-frequency transformer (MFT), which provides galvanic isolation and voltage conversion within the SST. The study addresses key design challenges, including insulation coordination under non-sinusoidal stress, high-current busbar design, and thermal management in compact, high-power systems. Practical design procedures are proposed for both full-scale and down-scaled transformers. Experimental work on a down-scaled prototype is carried out to verify the analytical and simulation results.

The novel approach using semiconductive coatings is introduced to control electric field distribution and mitigate partial discharges within the transformer. The work also includes guidelines for applying and validating such coatings in dry-type MFT designs.

LIST OF ABBREVIATIONS

ABS	Acrylonitrile Butadiene Styrene
AC	Alternating Current
A_c	Core cross-section area
A_e	Effective cross-sectional area
AI	Artificial Intelligence
AL	Inductance factor
AP	Area product
A	Magnetic vector potential
As	Winding area (secondary)
B_m	Maximum flux density
B_{sat}	Saturation flux density
Bi	Biot number
CF	Carbon fiber
Cdd	Disk-to-disk capacitance
Cp	Primary capacitance
Cps	Primary-to-secondary capacitance
Cst	Secondary-to-tank capacitance
Cstray	Total winding self-capacitance
Ctt	Turn-to-turn capacitance
Cth	Thermal capacitance
Cs	Secondary capacitance
CTC	Continuously Transposed Conductor
D	Winding diameter
DAB	Dual Active Bridge
d	Air gap
d_1	Distance MFT–tank side wall
d_2	Distance between MFTs
d_3	Distance MFT–upper lid
d_4	Distance MFT–bottom
Dd	Delta–delta transformer connection
DC	Direct Current
DRU	Diode-based Rectifier Unit
E	Electric field
EHV	Extremely high voltage
$E_{oil,max}$	Maximum permissible electric field in oil
E_t	Tangential electric field
f	Operating frequency

f_N	Nominal frequency
f_r	Resonant frequency
FFF	Fused Filament Fabrication
FEM	Finite Element Method
FlexH2	FlexH2 Research Project
Fo	Fourier number
G_{min}	Minimum required gap
Gpl	Primary-leg gap
Gps	Primary-secondary gap
Gpy	Primary-yoke gap
Gpp	Primary-primary gap
HFCT	High-Frequency Current Transformer
HP	High power
Hw	Window height
HV	High Voltage
HVDC	High-Voltage Direct Current
I_{phase}	Phase current
I_{line}	Line current
$I_{pri-line-peak}$	Primary line peak current
$I_{sec-line-peak}$	Secondary line peak current
iGSE	Improved Generalized Steinmetz Equation
i2GSE	Improved Improved Generalized Steinmetz Method
ISOP	Input-Series Output-Parallel
J	Current density
k	Empirical Steinmetz coefficient (context-dependent)
k	Thermal conductivity (context-dependent)
k_c	Core stacking factor
k_w	Window utilization factor
LFT	Line-Frequency Transformer
L_{LK}	Leakage inductance
L_a	Air-path inductance
l_c	Core-path inductance
l_e	Effective magnetic path length
LIC	Chopped-wave Lightning Impulse
LI	Full-wave Lightning Impulse
L_σ	Leakage inductance (symbolic)
L_m	Magnetizing inductance
L_{OC}	Open-circuit inductance
ML	Machine Learning
MLT	Mean Length per Turn
MMC	Modular Multilevel Converter
MMF	Magnetomotive Force
MFT	Medium-Frequency Transformer
MMR	Resonant Modular Multi-Level Converter
MV	Medium Voltage

N_{sm}	Number of submodules
n	Turn ratio
N_t	Total number of turns
PA	Polyamide
PD	Partial Discharge
PC	Polycarbonate
PCB	Printed Circuit Board
PET	Polyethylene Terephthalate
PETG	Polyethylene Terephthalate Glycol
PLA	Polylactic Acid
PCC	Point of Common Coupling
P	Rated power
PSM	Phase-Shift Modulation
PWM	Pulse-Width Modulation
PRPD	Phase-Resolved Partial Discharge
PTFE	Polytetrafluoroethylene
q	Electric charge
Q	Internal heat generation
ρ	Electrical resistivity
R	Total AC resistance / Outer radius (context-dependent)
R_{ac}	AC resistance
R_c	Core-loss resistance
R_{dc}	DC resistance
r_p	Primary resistance
r_s	Secondary resistance
RMS	Root Mean Square
S	Apparent power
SE	Steinmetz Equation
SiC	Silicon Carbide
SST	Solid-State Transformer
t	Time
T_{amb}	Ambient temperature
T_{oil}	Oil temperature
T	Period of waveform
THD	Total Harmonic Distortion
t_{tank}	Tank plate thickness
U_{dc}	DC voltage
$U_{electrolyzer}$	Rated electrolyzer input voltage
$U_{IGBT,derated}$	Derated IGBT voltage
$U_{pri-ph-peak}$	Primary phase peak voltage
$U_{sec-ph-peak}$	Secondary phase peak voltage
$U_{sec-LL-peak}$	Secondary line-to-line peak voltage
V_{bd}	Breakdown voltage
V_{dc}	DC voltage
V_{L-L}	Line-to-line voltage

V_{inc}	Partial discharge inception voltage
V_p	Primary voltage level
V_{phase}	Phase voltage (correct in your source)
V_s	Secondary voltage level
$V_{test-max}$	Maximum applied test voltage
V_{tank}	Tank volume
W	Winding width
W_e	Electrostatic energy
W_{mag}	Magnetic energy density
W_w	Window width
W_p	Primary width
X_{tank}	Tank dimension (X-axis)
Y_{tank}	Tank dimension (Y-axis)
Z_{tank}	Tank dimension (Z-axis)
Z	Impedance
ZCS	Zero Current Switching
ZVS	Zero Voltage Switching
η	Efficiency
θ	Dimensionless temperature
μ	Magnetic permeability
ω	Angular frequency
ϵ	Permittivity
ϵ_0	Permittivity of free space
ϵ_r	Relative permittivity
ϵ_{epoxy}	Permittivity of epoxy resin
β	Core-loss flux-dependency coefficient
β	Field enhancement factor

1

INTRODUCTION

*It is not knowledge, but the act of learning,
not possession but the act of getting there,
which grants the greatest enjoyment.*

Carl Friedrich Gauss

This chapter introduces the background and motivation for the research presented in this thesis, focusing on the role of solid-state transformers (SSTs) in large-scale green hydrogen production. It outlines conventional approaches for integrating offshore renewable energy with electrolyzers and describes the limitations of existing transformer solutions. The chapter highlights the potential of SSTs to address these limitations by improving efficiency, reducing system size, and enabling flexible operation. It also presents the research objectives, challenges, and key questions addressed in this PhD project, providing the foundation for the subsequent chapters.

Parts of this chapter have been published in:

R. Mirzadarani, M. Ghaffarian Niasar, Zhengzhao Li, Z. Qin, P. Vaessen, P. Bauer, Lou van Lieshout, "Three-Phase Medium-Voltage Medium-Frequency Transformer for SST in Green Hydrogen Production," IECON 2023- 49th Annual Conference of the IEEE Industrial Electronics Society, Singapore, Singapore, 2023, pp. 1-6 [1]

1.1. RESEARCH BACKGROUND

Green hydrogen production is considered as a solution for some modern world crises like climate change and global warming [2]. The main idea with green hydrogen production is to use renewable energies to power the electrolyzers to produce hydrogen. While renewable energies are vastly available, the challenge here is to harness and distribute the energy effectively.

Off-shore wind farms are considered green and reliable energy sources for green hydrogen production. Because of the long distance between these off-shore wind farms and the hydrogen production facilities, a high-voltage direct current (HVDC) transmission system is required. This HVDC is connected to the Dutch medium-voltage AC electricity grid (50 Hz) via the DC/AC converters, and this medium-voltage (MV) AC grid is used to power the electrolyzers. Traditionally, 6, 12, or 24-pulse rectifiers can be used to convert the AC input to the required DC. However, weight and physical dimensions can be problematic in traditional AC/DC solutions [1, 2] since the bigger the size, the more significant environmental footprint. To reduce the size and weight of the AC/DC section, it is possible to employ solid-state transformers (SSTs). The main idea of an SST is to increase the operating frequency in order to decrease the volume and weight of the transformer.

A typical SST comprises three main parts: 1) an input rectifier, 2) a DC/DC converter, and 3) an output inverter. However, an output inverter is not needed in the green hydrogen application since the electrolyzers work with direct voltage. Solid State Transformers operate at higher frequencies, and as a consequence, they have higher power density. SSTs are also capable to control the power flow more effectively and provide protection, which are vital for modern smart grids.

1.1.1. PROJECT OVERVIEW

This thesis is conducted within the FlexH2 project, a joint research initiative under the GROW program, funded by the RVO MOOI (Missiegedreven Onderzoek Ontwikkeling en Innovation) under Grant Agreement MOOI 52103, for offshore wind innovation in the Netherlands. GROW (Growth through Research, development and demonstration in Offshore Wind) brings together around 17–20 industrial and academic partners, including ABB, DNV, GE Vernova, Shell, TKF, TNO, Van Oord, and Vonk, together with TU Eindhoven and TU Delft, to accelerate technological innovations in offshore wind.

FlexH2 project aims to develop a novel concept for integrating offshore wind energy with onshore hydrogen production. The goal of the project is to improve overall system efficiency and flexibility by introducing new technologies such as grid-forming offshore wind turbines, advanced power conversion systems, and a multi-terminal

hybrid HVDC transmission network. The project involves several partners and is structured into six main work packages, each focusing on a specific aspect of the system.

The work package structure of the project is shown in Figure 1.1. This work is conducted within Work Package 2, which focuses on the design and control of a solid-state transformer for electrolysis. The goal is to develop a medium-voltage power-electronic converter that efficiently connects the AC power from offshore wind to the DC input of electrolyser systems. The proposed SST solution will be compared with state-of-the-art technologies to assess its performance, scalability, and integration feasibility. A hardware-in-the-loop simulation and a medium-voltage demonstrator will be used to validate the design and contribute to raising the technology readiness level of the overall FlexH2 concept.

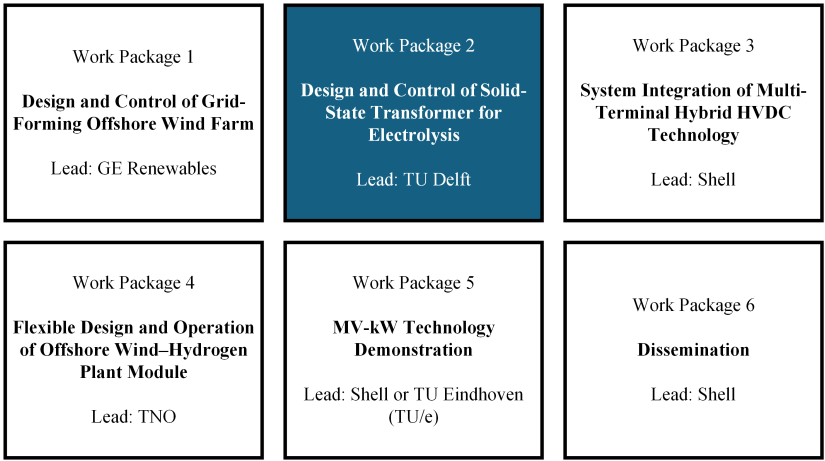


Figure 1.1: Overview of the FlexH2 project work packages. Each work package addresses a specific aspect of the offshore wind-to-hydrogen system design, development, integration, or dissemination. This thesis is conducted within Work Package 2, led by TU Delft.

1.1.2. GREEN HYDROGEN PRODUCTION

Green hydrogen production generates hydrogen using renewable energy sources, through water electrolysis, where energy from sources like wind or solar powers electrolyzers to split water into hydrogen and oxygen [1, 2]. This process avoids greenhouse gas emissions, distinguishing it as a sustainable alternative to traditional methods [3–6]. Recognized globally as crucial for supporting renewable energy expansion and easing grid

congestion, green hydrogen has applications including transportation, industry, and power generation, where it serves as a clean fuel source. Additionally, green hydrogen supports energy storage, enabling excess renewable energy to be stored as hydrogen for later use. Its potential to facilitate decarbonization in hard-to-electrify sectors, such as heavy industry and transportation, reinforces its role in advancing a cleaner, more secure energy system.

Conventional hydrogen production systems rely on 50/60 Hz transformers and rectifiers. These systems step down the grid voltage and convert AC to DC for electrolyzers, but they are typically large and less efficient due to their low-frequency operation, making them less suitable for applications where compact design and high energy density are required.

Solid-State Transformers (SSTs) offer an advanced alternative in green hydrogen production by utilizing medium-frequency transformers with power electronics to achieve high-frequency voltage conversion and galvanic isolation. This approach reduces system size and weight compared to 50/60 Hz, reduces the losses, and enables precise power flow control. The modular structure and high power density of SSTs make them well-suited for large-scale applications, supporting the flexibility and efficiency needed for modern green hydrogen facilities.

Unlike gray hydrogen, which uses fossil fuels, or blue hydrogen, which combines fossil fuels with carbon capture, green hydrogen is fully renewable and emissions-free [7, 8]. Green hydrogen is important because of its sustainability and minimal environmental impact, making it central to the transition to a low-carbon energy system. It is a versatile energy carrier, applicable in transportation, power generation, and industry. Moreover, it strengthens energy security by diversifying energy sources and reducing dependence on fossil fuels.

1.1.3. CONVENTIONAL SOLUTIONS

In a conventional offshore wind farm setup, distances are typically under 80 km; therefore, AC submarine cables can be used to transmit energy directly to the grid. As offshore wind farms expand to distances beyond 80 km, the industry is transitioning to DC submarine cables, still primarily sending power to the grid, which then distributes energy, including to hydrogen electrolyzers. In the FlexH2 model, however, offshore facilities will be more compact, with most of the energy going directly to large-scale hydrogen electrolyzers, reducing reliance on grid capacity. A grid connection will remain, but the energy distribution between the electrolyzers and the grid will be dynamically managed by a control unit, allowing flexible prioritization. These setups are shown in Figure 1.2.

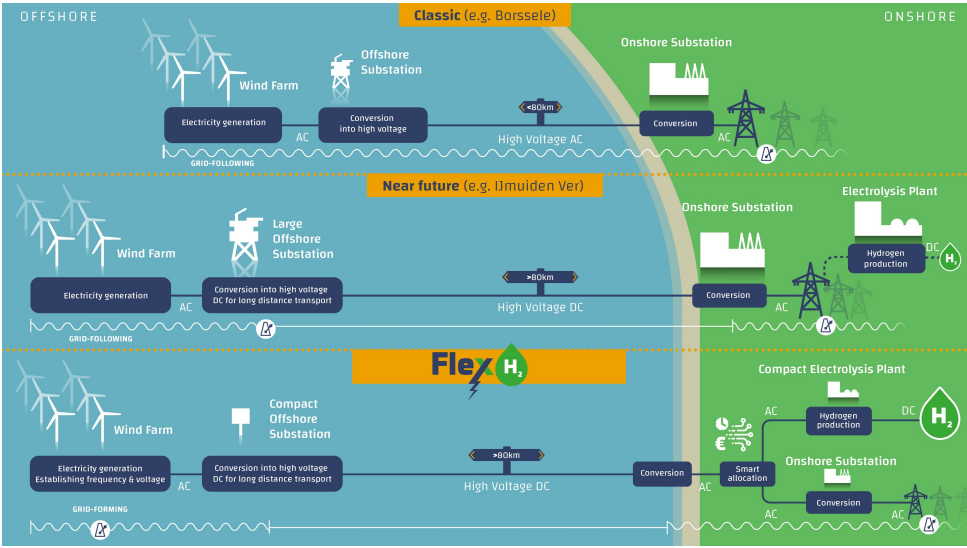


Figure 1.2: Comparison of offshore wind farm systems: Conventional system supplying energy to the grid, and FlexH₂ system prioritizing direct energy flow to hydrogen electrolyzers, with flexible grid-sharing control [5].

1.1.4. IMPLEMENTING SOLID-STATE TRANSFORMERS FOR GREEN HYDROGEN PRODUCTION

Figure 1.3 presents a green hydrogen production system utilizing offshore wind energy and HVDC (High Voltage Direct Current) transmission for power transport. The process begins with a wind farm generating electricity, which is then stepped up by a line-frequency transformer (LFT) to a high voltage suitable for long-distance transmission. This AC power is converted to DC by a rectifier and transmitted via an HVDC submarine cable to an onshore facility. This configuration maximizes efficiency and minimizes losses over long transmission distances, particularly suitable for offshore wind farms. Upon reaching the shore, the DC power is converted back to AC by a Modular Multilevel Converter (MMC). The energy then passes through a three-winding transformer, which adjusts the voltage to match the electrolyzer's requirements. Next, a 12-pulse thyristor rectifier and an interphase transformer further condition the DC power before it reaches the electrolyzer, where it facilitates water splitting to produce green hydrogen.

Figure 1.4 illustrates a green hydrogen production system similar to the conventional one, but with a notable addition: a solid-state transformer (SST) in the onshore facilities. Like the conventional setup, this system is powered by an offshore wind farm, using HVDC submarine

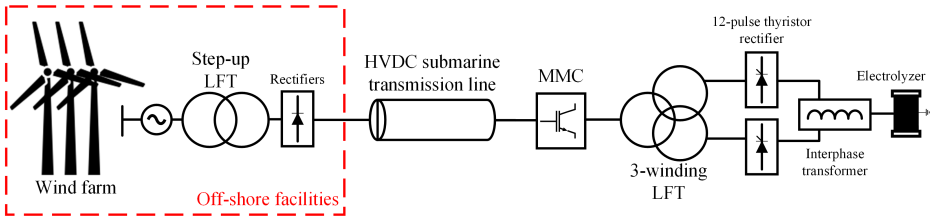


Figure 1.3: Block diagram of a green hydrogen production system powered by an offshore wind farm. The system utilizes HVDC transmission and modular multilevel conversion for efficient power transport to an onshore electrolyzer, enabling sustainable hydrogen generation [1].

transmission to transfer power to shore. The offshore section includes a step-up low-frequency transformer (LFT) and rectifier, followed by a modular multilevel converter (MMC). The onshore section incorporates a 3-winding autotransformer feeding a 12-pulse diode rectifier and an interphase transformer, which provides smoother DC output for the electrolyzer. Here, the SST module, consisting of an MMC and a medium-frequency transformer (MFT), further steps down the voltage before feeding into the electrolyzer through an output rectifier. The inclusion of the SST enables more flexible and efficient voltage conversion compared to conventional LFT-only systems, providing a compact and efficient design for green hydrogen production.

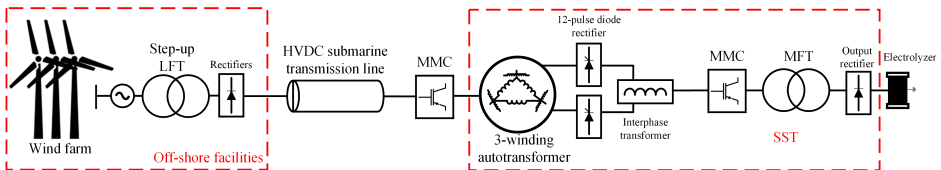


Figure 1.4: Green hydrogen production system using an SST in onshore facilities, integrating offshore wind power transmitted via HVDC submarine cables to an MMC, autotransformer, and electrolyzer for efficient voltage conversion [1].

Traditional solutions for connecting electrolyzers to the power grid involve LFTs to step down voltage and rectifiers to convert AC to DC. While relatively cost-effective in components, these systems are large and heavy due to presence of bulky of LFTs. SSTs present a promising alternative, offering several key remarks:

- a. **Reduced Size and Weight:** Operating at higher frequencies, SSTs can use smaller, lighter medium-frequency transformers (MFTs).

For example, a 25 MVA, 400 Hz MFT designed for green hydrogen production reduces active part weight by 69% compared to a conventional LFT.

- b. **Comparable Efficiency:** Although conventional power transformers of this rating have efficiencies above 99%, SSTs can achieve comparable efficiency by minimizing losses in power conversion due to smaller magnetic components.
- c. **Enhanced Control:** SSTs allow better power flow management, voltage regulation, and power factor correction.
- d. **Fault Isolation and Protection:** SSTs provide robust fault isolation and protection, supporting grid stability and protecting the electrolyzers.
- e. **Modularity:** Several parts of SSTs such as MMC based inverter are made of modules. This modular design makes SSTs adaptable to various power levels and applications.

1.2. MOTIVATION

The transition toward a low-carbon energy system necessitates innovative solutions for large-scale integration of renewable energy sources. Among these, green hydrogen has emerged as a promising energy carrier capable of addressing critical challenges in decarbonizing hard-to-electrify sectors such as industry, heavy transportation, and long-term energy storage. However, realizing its full potential requires the deployment of efficient, scalable, and compact power conversion systems that can interface renewable energy sources with high-capacity electrolyzers.

Traditional line-frequency transformer (LFT) based systems are not well-suited for such applications due to their large size, heavy weight, and limited dynamic control. Also, smaller active parts required for SST solutions means smaller transformers which can be delivered faster by the manufacturers in today's saturated transformer manufactory capabilities, if manufacturers establish required techniques for building SSTs. As green hydrogen facilities scale up to tens of megawatts, these limitations become more pronounced, especially in offshore or space-constrained installations. Furthermore, the increasing use of medium-voltage grids for electrolyzer integration requires power electronic converters that can provide isolation by transformers, voltage transformation, and power quality management.

Solid-State Transformers (SSTs), which leverage high-frequency operation and advanced power electronics, offer a compelling alternative to conventional systems. Their ability to reduce volume and weight,

enhance control, and improve efficiency makes them particularly attractive for green hydrogen production. Nonetheless, the implementation of SSTs introduces new technical challenges, particularly in the design of medium-frequency transformers (MFTs) that operate at high power and voltage levels.

Designing such MFTs involves addressing several interrelated issues, including insulation coordination under non-sinusoidal stress, thermal management in compact configurations, and electromagnetic design compatible with high-current, high-frequency operation. Existing standards such as IEC 60076 are often inadequate or even irrelevant when applied to SST systems, necessitating new design practices and validation approaches.

This thesis is motivated by the need to bridge this gap through a systematic investigation of SST-based architectures for green hydrogen production. It aims to develop practical design methods for medium-voltage, medium-frequency transformers that are reliable, scalable, and aligned with the operational requirements of large-scale electrolyzer systems. The research is further driven by the opportunity to contribute to the global transition toward clean energy by enabling more effective and efficient use of renewable power in hydrogen production.

1.3. THESIS OBJECTIVE AND RESEARCH QUESTIONS

Designing a medium-frequency transformer for high-power applications is quite challenging for several reasons. Firstly, the existing transformer standards cannot be used directly for an HP-MV-MFT. For instance, considering IEC 60076 results in an oversized MFT because the insulation and clearance values in the aforementioned standard are for 50/60 Hz transformers. The presence of switching devices with fast switching capability changes the characteristics and behaviour of the transformer; therefore, it is essential to declare a reasonable set of test requirements for the MFT. MFT's overload, overcurrent, and short-circuit conditions are different since the power electronics section (inverter) cannot provide such a huge power for the MFT. Therefore, the existing standards cannot be used directly.

Secondly, the design of some parts of the HP-MV-MFT requires novel techniques. For example, the low-voltage busbar system of a high-power, medium-frequency system is a relatively unexplored area requiring careful design considerations. Based on the reviewed literature [2, 9–12] there are a number of studies about high-power busbars. On the other hand, there are studies about busbars in low-power and power electronics applications. However, the combination of both high-power and medium-frequency is relatively novel and needs to be more explored.

Finally, thermal management is challenging because by using higher

frequencies, the transformer becomes smaller. This compact transformer has smaller surfaces to dissipate the heat. In addition, using higher frequency results in high-frequency losses, and these extra losses make thermal management hard to design. Therefore, the research challenges are listed as follows:

- Insulation design (Medium-frequency stress + IEC 60076-3)
- Busbar design (Medium-frequency + high-current)
- Thermal management:
 - Higher power density → smaller surfaces for heat dissipation
 - Medium-frequency losses

The above-mentioned research challenges raise research questions. Overall, there are five main research questions as follows:

1. What are the design requirements for HP-MV-MFT in green hydrogen production applications?
2. Which set of test requirements are needed for the HP-MV-MFT based on the existing transformer standards?
 - What deviations of IEC60076-3 can be allowed under what conditions?
3. What type of high-frequency (dielectric) stress is expected for SST insulation?
 - How to develop and validate a model that assesses high-frequency (dielectric) stress?
4. What are the design requirements for a high-power, medium-frequency busbar?
5. Which cooling method is suitable for an HP-MV-MFT regarding its high-power density?

1.4. CONTRIBUTIONS

This thesis investigates the characteristics, design, and downscaled prototyping of an industrial-scale solid-state transformer (SST) for medium-voltage, medium-frequency applications. It considers the challenges specific to SST design at higher frequencies and voltages, with a focus on practical implementation aligned with industrial requirements. The study aims to provide a comprehensive understanding of the steps involved in designing and fabricating such transformers.

The work covers the development of a practical design procedure and optimization process for both full-scale and down-scaled prototypes.

It addresses challenges related to high-frequency operation, such as current unbalance in high-current windings and increased losses. The study includes methods to mitigate these issues while maintaining efficiency and reliability in transformer operation.

A novel approach introduced in this research is the use of semiconductive paint to encapsulate the electric field inside the windings, aiming to improve insulation performance by mitigating partial discharges. The thesis also documents the prototyping process, offering clear design and fabrication guidelines that can support future research and industrial developments in the field of medium-frequency, medium-voltage SSTs for green hydrogen and related applications.

1.5. THESIS OUTLINES

This thesis is structured to address the design, challenges, and prototyping of medium-voltage, medium-frequency solid-state transformers (SSTs) for green hydrogen production applications. The rest of this thesis is structured as follows:

Chapter 2 reviews the use of solid-state transformers in green hydrogen production, discussing relevant literature, SST topologies, and performance metrics. It compares three-phase and single-phase SST configurations and evaluates design considerations for various SST structures.

Chapter 3 focuses on the design and optimization of medium-voltage, medium-frequency transformers (MFTs) within SST systems. It outlines relevant standards, detailed design procedures for different SST topologies, and methods for performance evaluation and optimization.

Chapter 4 addresses insulation coordination for MFTs, covering dry-type and oil-type transformers, field grading techniques using semiconductive materials, and the use of 3D-printed parts as insulators. It also discusses partial discharge considerations and insulation requirements in power electronics.

Chapter 5 investigates parasitic parameters and current balancing issues in high-current, medium-frequency transformers. It describes the impact of parasitic elements, electromagnetic interference, and methods for mitigating current unbalancing, supported by experimental validation.

Chapter 6 examines losses, thermal management, and cooling system design. It presents analyses of no-load and load losses, employs data-driven methods for estimating core losses, and discusses strategies for effective thermal management and cooling in high-power density transformers.

Chapter 7 describes the prototyping and experimental setup of the down-scaled demonstrator. It covers design optimization, electric and magnetic field simulations, conductor design, and termination

considerations. The chapter also details test procedures, including partial discharge and breakdown tests, transformer testing, and SST performance evaluation.

Finally, Chapter 8 concludes the thesis, summarizing key findings and providing insights for future research directions in the design and application of SSTs for green hydrogen production.

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2

SOLID-STATE TRANSFORMER IN GREEN HYDROGEN PRODUCTION

*Do not go gentle into that good night,
Rage, rage against the dying of the light.*

Dylan Thomas

This chapter reviews the application of solid-state transformers (SSTs) in green hydrogen production, focusing on their role in improving power conversion efficiency and reducing system size for high-power electrolyzers. It explains the motivations for using SSTs instead of conventional line-frequency transformers, particularly for medium-voltage to low-voltage DC conversion. Key design considerations, such as topology selection, thermal management, and insulation requirements, are discussed, with comparisons among MMC, MMR, and ISOP topologies based on loss, size, and weight. The chapter provides practical insights and design methodologies for implementing SST-based systems in green hydrogen production.

Parts of this chapter have been published in:

Z. Li, **R. Mirzadarani**, M. Ghaffarian Niasar, M. Itraj, L. van Lieshout, P. Bauer, Z. Qin, "Medium-Voltage Solid-State Transformer Design for Large-Scale H₂ Electrolyzers," in IEEE Open Journal of Power Electronics, vol. 5, pp. 936-955, 2024 [1].

R. Mirzadarani, M. Ghaffarian Niasar, Zhengzhao Li, Z. Qin, P. Vaessen, P. Bauer, Lou van Lieshout, "Three-Phase Medium-Voltage Medium-Frequency Transformer for SST in Green Hydrogen Production," IECON 2023- 49th Annual Conference of the IEEE Industrial Electronics Society, Singapore, Singapore, 2023, pp. 1-6 [2]

The growing demand for large-scale green hydrogen production has introduced new challenges in power conversion, especially when electrolyzers are powered from medium-voltage grids. Conventional solutions using line-frequency transformers (50/60 Hz) and passive rectifiers are often large, heavy, and inefficient for high-power applications. To overcome these limitations, SSTs are being explored as a compact and efficient alternative. SSTs operate at higher frequencies and enable better control, improved power density, and reduced system size. This chapter introduces the concept of SSTs in the context of green hydrogen production and explains the design considerations and topological choices evaluated in this research.

The work presented in this chapter is based on two published studies. The first paper focuses on the detailed design and comparison of three SST topologies, MMC, MMR, and ISOP, for use in large-scale electrolyzer systems. It includes quantitative analyses of loss, size, and weight for each configuration. The second paper complements this by analyzing the design of the medium-frequency transformer (MFT) at the heart of the SST. It also compares single-phase and three-phase implementations and discusses the benefits of using three-phase MFTs in high-power applications.

Section 2.1 provides an overview of related work in SST and MFT design, highlighting key contributions from the literature and compares single-phase and three-phase SST architectures, explains why the three-phase configuration is preferred in this application. Section 2.3 introduces and explains three major SST topologies, MMC, MMR, and ISOP, along with a cascaded structure used for benchmarking. Finally, Section 2.4 presents performance metrics, including estimated losses, size, and weight, and discusses how these parameters vary across topologies and operating frequencies.

2.1. LITERATURE REVIEW

Solid-State Transformers (SSTs) have been proposed as a promising alternative to conventional line-frequency transformers (LFTs) for applications where volume, weight, and controllability are critical. In recent years, research has increasingly focused on SSTs for use in green hydrogen production, where electrolyzers require high-power, low-voltage DC inputs sourced from medium-voltage AC grids. Traditional power conversion methods, based on line-frequency transformers and passive rectifiers, struggle to meet the demands of compactness, efficiency, and dynamic control in large-scale electrolyzer systems.

Earlier studies on SSTs mostly focused on low-to-medium power levels, often employing single-phase transformers with ferrite or nanocrystalline cores. These configurations are widely reported in the literature due to their simplicity and suitability for lab-scale demonstrations. However,

for practical applications in the tens of megawatts range, such as hydrogen production powered by offshore wind farms, three-phase SSTs and oil-immersed medium-frequency transformers (MFTs) become necessary. Compared to single-phase SSTs, three-phase architectures offer reduced core size, better current handling, and improved harmonic performance, making them more suitable for high-power applications.

Despite the large body of work on SSTs, few studies provide a complete quantitative comparison of multiple topologies for megawatt-scale electrolyzers. Similarly, while various MFT designs have been proposed, detailed assessments of transformer insulation, thermal behavior, and their impact on overall system performance remain limited. This gap motivated the present work, which aims to evaluate and compare feasible SST topologies—namely Modular Multi-Level Converter (MMC), Resonant Modular Multi-Level Converter (MMR), and Input-Series Output-Parallel (ISOP), with practical transformer and cooling system considerations.

Table 2.1 provides a comprehensive survey of 52 medium-frequency transformer (MFT) prototypes reported in the literature between 2002 and 2025. Each row represents one prototype, while the columns summarize key design and performance attributes. These include the *rated power* (P , in kVA), *operating frequency* (f , in kHz), *core material*, *structural configuration* (Str.), *winding type*, *cooling method* (Cool.), and *insulation type* (Insul.).

Furthermore, the table lists the *primary and secondary voltage levels* (V_p and V_s , in kV), and reports available *insulation testing results*, such as partial discharge inception/extinction voltages, AC withstand, and lightning impulse levels. The data are collected from both academic and industrial sources and are annotated with references for traceability. This compilation aims to highlight the diversity of materials, topologies, and insulation strategies employed in state-of-the-art MFT development.

Table 2.1 highlights clear trends in the development of medium-frequency transformers over the last two decades. First, the applied frequency has gradually increased from a few hundred hertz in early prototypes to tens or even hundreds of kilohertz in recent designs. This trend is driven by the need for higher power density and smaller transformer size. Second, core materials have converged toward nanocrystalline and ferrite for high-frequency, lower-power designs, while amorphous and silicon-steel cores remain dominant for medium-voltage and high-power applications. Third, almost all high-power (>500 kVA) MFTs are oil-filled, demonstrating that oil immersion is still the most reliable solution for insulation strength, partial-discharge mitigation, and thermal management at high voltage levels. In contrast, dry-type insulation is predominantly used in lower-power or laboratory prototypes. Finally, the table shows a consistent increase in the rigor of insulation testing (AC withstand, impulse, and PD levels), underlining

the growing emphasis on insulation coordination as MFTs move toward multi-megawatt industrial applications.

Table 2.1: Survey of Medium-Frequency Transformers With Corresponding Electrical, Structural, and Insulation Parameters

No	Year	Title	P (kVA)	f (kHz)	Core Material	Str	Winding	Cool.	Insul.	V _p (kV)	V _s (kV)	Insulation Tests
1	2002	ABB [3]	350	10	Amorphous	Shell	Coaxial	Water	Solid	3.0	3.0	38 kV insulation test, 28 kV PD extinction, 95 kV impulse
2	2005	TU Delft [4]	50	50	Amorphous	Shell	Foil	Heat pipes	NA	0.75	0.6	NA
3	2007	ALSTOM [5]	1500	5	NA	NA	NA	Water	NA	15	1.65	NA
4	2007	ABB [6]	1200	0.4	Silicon steel	NA	Bar	Oil	Oil	15	1.8	NA
5	2007	Bombardier [7]	375	8	Nano-crystalline	Shell	Hollow	Water	Solid	2.8	2.8	33 kV PD free, 100 kV surge test
6	2008	West Bohemia [8]	12	0.4	NA	NA	NA	Natural convection	Solid	0.35	0.6	NA
7	2009	KTH [9]	3000	0.5	Amorphous	Shell	Litz	Natural convection	Solid	33	3	NA
8	2009	KTH [10]	170	4	Amorphous	Shell	Litz / Foil	Water	NA	15	1.8	NA
9	2011	FAU-EN [11]	450	5.6	Nano-crystalline	Core	Hollow	Water	Oil	25	NA	NA
10	2011	FREEDM-Gen. I [12]	20	3	Amorphous	NA	NA	Natural convection	Solid	7.2	0.24	NA
11	2012	IKERLAN-DAB [13]	400	1	Silicon steel	Shell	Foil	Natural convection	Oil	3	0.75	18 kV Insulation level
12	2012	IKERLAN-LLC [13]	400	5	Nano-crystalline	Core	Litz	Natural convection	Oil	3	0.75	18 kV Insulation level
13	2013	FREEDM-Gen. II [14]	20	20	Nano-crystalline	NA	NA	Natural convection	Solid	7.2	0.38	NA
14	2014	ABB [15]	1200	1.75	Nano-crystalline	Shell	Litz	Oil	Oil	15	1.5	34.5 kV rms dielectric test, 75 kV impulse
15	2014	ETH [16]	166	20	Nano-crystalline	Shell	Litz	Water	Solid	2	0.4	NA
16	2014	ETH [16]	50	40	Ferrite	Core	NA	NA	Solid	1	0.4	NA

No	Year	Title	P (kVA)	f (kHz)	Core Material	Str.	Winding	Cool.	Insul.	V _p (kV)	V _s (kV)	Insulation Tests
17	2016	FREEDM-Gen. III [17]	20	37	NA	NA	NA	Natural convection	Solid	7.2	0.28	NA
18	2016	ETH [18]	NA	100	Ferrite	Shell	Litz	Oil	Oil	14.4	NA	Partial discharge at 115 kV for 60 min, up to 136% overvoltage NA
19	2017	Chalmers [19]	50	5	Nano-crystalline	Shell	Litz	Conduction	Solid	1	3	NA
20	2017	Osaka [20]	5	1	Nano-crystalline	NA	NA	NA	Solid	3	NA	NA
21	2017	ABB [21]	240	10	Nano-crystalline	Shell	Litz	Forced air	Solid	0.6	0.9	AC test voltage 69 kV, PD test voltage 53 kV, LI test voltage 150 kV
22	2018	Virginia Tech [22]	15	500	Ferrite	Core	Litz	Natural convection	Solid	4.16	0.4	30 kV BIL test, 12 kV AC applied voltage, Partial discharge ≤ 3.8 pC
23	2019	EPFL [23]	100	10	Ferrite	Shell	Litz	Natural convection	Air	0.75	0.75	4 kV partial discharge NA
24	2020	ETH [24]	25	48	Ferrite	Shell	Litz	Natural convection	Solid	7	0.4	NA
25	2020	RWTH [25]	200	1.55	Nano-crystalline	Shell	Litz	NA	Solid	1	25	±25 kV dielectric strength test
26	2021	Chalmers [26]	50	5	Ferrite	Matrix	Litz	Oil	Oil	0.422	4.5	125 kV DC test voltage
27	2021	Texas [27]	200	15	Nano-crystalline	Core	Litz	Natural convection	Solid	NA	NA	Tested at 12 kV AC peak, PD-free up to 5.3 kV rms NA
28	2021	RWTH [28]	5000	1	Silicon steel	Core	Foil	Natural convection	Solid	5	NA	NA
29	2022	XJTU [29]	10	10	Ferrite	Shell	Litz	Natural convection	Solid	1	0.75	20 kV AC withstand voltage test
30	2022	Wisconsin [30]	100	50	Ferrite	Shell	Litz	Conduction	Solid	4.16	0.5	30 kV BIL test

No	Year	Title	P (kVA)	f (kHz)	Core Material	Str.	Winding	Cool.	Insul.	V _p (kV)	V _s (kV)	Insulation Tests
31	2022	Tsinghua [31]	60	10	Nano-crystalline	Core	NA	NA	Solid	10	NA	Partial discharge test at 10 kV, applied voltage test at 20 kV 55 kV BIL, 60 kV Hipot, 90 kV lightning impulse 60 kV Hipot test
32	2022	GIT [32]	50	16	Nano-crystalline	Shell	Coaxial	Forced air	Solid	7.2	0.35	
33	2022	ERICA [33]	125	10	Ferrite	Shell	Litz	Natural convection	Solid	1	0.75	
34	2022	Seoul [34]	2.4	59	Ferrite	NA	Litz	Natural convection	Solid	0.1	0.1	
35	2022	VNIT [35]	4	50	Sendust	Core	Litz	Natural convection	Solid	NA	NA	Insulation up to 120 kV peak (based on air gap design) NA
36	2022	SPEC [36]	3000	60	Amorphous	Core	Litz	Forced air	Solid	13.8	1.5	
37	2022	NCSU [37]	100	20	Nano-crystalline	Core	Foil	Oil	Oil	7.2	0.8	60 kV Lightning Impulse Test, 34 kV Applied Voltage Test, 23.5 kV Partial Discharge Test 10 kV DC Hipot Test, 20 kV Bushing Clearance Test
38	2023	Huawei [38]	28.4	40	Nano-crystalline	Shell	Litz	Forced air	Solid	12	0.75	
39	2023	Lodz University of Technology [39]	500	6	GOES Wound Core	Core	Foil	Forced air	Solid	0.8	0.533	34 kV RMS applied voltage, 75 kV lightning impulse test, Partial discharge ≤ 10 pC NA
40	2023	Virginia Tech [40]	15	200	Ferrite	Core	Litz	Forced air	Solid	4.16	0.8	
41	2024	Wuhan University [41]	1000	10	Ferrite	Core	NA	NA	Solid	10	1	

No	Year	Title	P (kVA)	f (kHz)	Core Material	Str.	Winding	Cool.	Insul.	V _p (kV)	V _s (kV)	Insulation Tests
42	2024	Zhejiang University [42]	20	50	Ferrite	Shell	Litz	NA	Solid	0.75	0.75	15 kV RMS line-frequency applied voltage NA
43	2024	University of Wollongong [43]	500	10	Ferrite	Core	NA	NA	Solid	4.16	0.48	
44	2024	Xi'an Jiao Tong University [44]	1000	10	Ferrite	Core	NA	NA	Solid	10	1	
45	2024	Hebei University of Technology [45]	50	20	Nano-crystalline	Matrix	Litz	Forced air	Solid	1	NA	Dielectric withstand test (electric field < 1.93 kV/mm)
46	2024	SPEC [46]	1000	10	Ferrite	Core	NA	NA	Solid	10	1	NA
47	2024	Virginia Tech [47]	15	180	Ferrite	Core	NA	NA	Solid	1.6	1	29 kV insulation capability
48	2025	IIT [48]	5.5	56	Ferrite	Core	NA	NA	Solid	0.11	0.22	NA
49	2025	IIT [49]	4.5	20	Ferrite	Core	NA	NA	Solid	0.9	0.587	NA
50	2025	CSU [50]	1	100	Ferrite	Matrix	NA	NA	Solid	0.22	0.22	NA
51	2025	FREEDM [51]	1000	10	Ferrite	Core	NA	NA	Solid	13.2	0.75	Partial Discharge (PD) tests per IEC 60076-11, max PD level <10 pC
52	2025	TU Munich [52]	3	20	Ferrite	Core	NA	NA	Solid	0.3	0.1	NA

2.2. THREE-PHASE VERSUS SINGLE-PHASE SST

In the context of large-scale green hydrogen production, the design of power conversion systems plays a crucial role in ensuring high efficiency, compactness, and reliability. Solid-State Transformers (SSTs), operating at medium frequencies, have emerged as a promising alternative to conventional line-frequency transformers, primarily due to their ability to reduce system size and weight while improving power quality. While many prior studies focus on single-phase SST configurations, these are often limited to lower power ratings. For applications requiring multi-megawatt power delivery, such as feeding industrial electrolyzers, a three-phase SST design presents several distinct advantages. This section outlines the key technical and practical reasons for favoring a three-phase SST over a single-phase alternative in high-power medium-voltage scenarios.

The main reason for selecting a three-phase SST is its suitability for high-power applications. In green hydrogen production, the electrolyzers typically operate at power levels of tens of megawatts. Most previous works on SSTs are based on single-phase designs and limited to a few hundred kVA. These configurations are not adequate for large-scale industrial use.

A three-phase SST enables more efficient use of the magnetic core. The return path of the magnetic flux is shared between phases, which allows for a smaller and lighter core for the same power rating. In the case studied, the three-phase medium-frequency transformer (MFT) showed a 69% reduction in active part weight compared to a conventional line-frequency transformer.

The Yd (star-delta) configuration used in the three-phase MFT also helps reduce voltage stress on the primary winding and current on the secondary winding. This leads to lower insulation requirements and smaller conductor sizes. As a result, both the transformer and the overall system become more compact and cost-effective.

From the power electronics point of view, a three-phase modular multilevel converter (MMC) results in lower harmonic distortion compared to a single-phase MMC. Although the number of switches is higher, the total switching device rating remains similar. Therefore, the system benefits from improved performance without a significant increase in cost.

Finally, the RMS current in the DC bus capacitors is lower in the three-phase configuration. This improves the thermal behavior of the capacitors and extends their lifetime. Together, these points support the decision to use a three-phase SST for medium-voltage, high-power hydrogen electrolyzer systems.

Choosing between single-phase and three-phase SSTs depends on the power electronics and transformer section; nevertheless, Table 2.2 compares them regarding the MFT. Furthermore, the comparison

between three-phase and single-phase transformers depends on the application and configuration of the MFT. Since the present study focuses on a step-down transformer required for green hydrogen production, the selected configuration is Yd (primary-star-secondary-delta) for two reasons. Firstly, in a star winding, the effective voltage of each phase is calculated as follows:

$$V_{\text{Phase}} = \frac{V_{\text{L-L}}}{\sqrt{3}} \quad (2.1)$$

where $V_{\text{L-L}}$ is the line-to-line voltage. Therefore, a lower voltage is connected to the primary winding. Secondly, in a delta winding, the line current is calculated as follows:

$$I_{\text{Line}} = \sqrt{3} \cdot I_{\text{Phase}} \quad (2.2)$$

where I_{Phase} is the phase current. Thus, the secondary winding deals with a lower current level. The cross-sectional area of the magnetic core is calculated as:

$$A_c = \frac{V_p}{k_w \cdot k_c \cdot B_m \cdot N_p \cdot f} \quad (2.3)$$

where B_m is the maximum magnetic flux density, N_p is the number of primary turns, and f is the frequency. k_w and k_c are the waveform and core stacking factors, respectively.

According to Equation 2.1 and Table 2.2, for the single-phase case, $V_{p1} = 2 \cdot V_{p3}$. Considering $N_{p1} = N_{p3}$ and all other parameters the same, this results in:

$$A_{c1} = 2 \cdot A_{c3} \quad (2.4)$$

where A_{c1} and A_{c3} are the core cross-sectional areas of the single-phase and three-phase transformers, respectively. The rating power of a single-phase transformer is calculated as:

$$S_1 = I_{p1} \cdot V_{p1} \quad (2.5)$$

where I_{p1} is the primary current and V_{p1} the primary voltage in the single-phase case. The rating power of a three-phase transformer is given by:

$$S_3 = 3 \cdot I_{p3} \cdot V_{p3} \quad (2.6)$$

where I_{p3} is the primary current in the three-phase case. To make the comparison, let $S_1 = S_3 = S$.

The area product of a single-phase transformer is calculated as:

$$AP_1 = A_{w1} \cdot A_{c1} \quad (2.7)$$

Table 2.2: Comparison of Parameters between Single-phase and Three-phase SSTs

Parameter	Single-phase	Three-phase
Primary voltage ($V_{p,rms}$)	$\frac{\sqrt{2}}{2} \cdot V_{DC}$	$\frac{\sqrt{2}}{4} \cdot V_{DC}$
Core cross-section area (A_c)	$\frac{\sqrt{2}/2 \cdot V_{DC}}{k_c k_w B_m N_{p1} f}$	$\frac{\sqrt{2}/4 \cdot V_{DC}}{k_c k_w B_m N_{p3} f}$
Primary current (I_p)	$\frac{\sqrt{2} \cdot S}{V_{DC}}$	$\frac{\sqrt{2}}{3} \cdot \frac{S}{V_{DC}}$
Primary area (A_p)	$\frac{\sqrt{2} N_{p1} S}{J \cdot V_{DC}}$	$\frac{\sqrt{2} N_{p3} S}{3J \cdot V_{DC}}$
Secondary voltage ($V_{s,rms}$)	$\frac{\pi \sqrt{2}}{4} \cdot V_{out}$	$\frac{\pi \sqrt{2}}{6} \cdot V_{out}$
Secondary current (I_s)	$\frac{2 \sqrt{2} \cdot S}{\pi \cdot V_{out}}$	$\frac{\sqrt{2} \cdot S}{\pi \cdot V_{out}}$
Secondary area (A_s)	$\frac{2 \sqrt{2} N_{s1} S}{\pi \cdot V_{out} \cdot J}$	$\frac{\sqrt{2} N_{s3} S}{\pi \cdot V_{out} \cdot J}$
Winding area (A_w)	$\frac{\sqrt{2} \cdot S}{J} \left[\frac{N_{p1}}{V_{DC}} + \frac{2N_{s1}}{\pi V_{out}} \right]$	$\frac{2 \sqrt{2} \cdot S}{J} \left[\frac{N_{p3}}{3V_{DC}} + \frac{N_{s3}}{\pi V_{out}} \right]$

while for a three-phase transformer, the area product is defined as [53]:

$$AP_3 = \frac{3}{2} \cdot A_{w3} \cdot A_{c3} \quad (2.8)$$

Referring again to Table 2.2, the comparison yields:

$$AP_1 = AP_3 \quad (2.9)$$

Although the area product of two cases are the same based on Equation 2.9, three-phase MFT saves size and weight in comparison to a single-phase solution when the weight of the active part is considered.

The three-phase transformer is smaller because the return path of the magnetic flux is completed through the legs of another phases. Therefore, the core can be selected smaller for the same power rating. On the other hand, the power electronics section (MMC) must be designed three-phase or single-phase according to the structure of the SST. The total rating of switching devices (number of switches $\times$ Rating voltage $\times$ Rating current) for single-phase and three-phase MMCs are the same; therefore, the total price of the switching devices are similar. Moreover, the three-phase MMC has better performance regarding the total harmonics distortion (THD). Based on the above discussion, it can be decided regarding the THD and MFT's weight and size. Therefore, the three-phase SST is more beneficial for green hydrogen production. The single-phase and three-phase cases are compared in Figure 2.1.

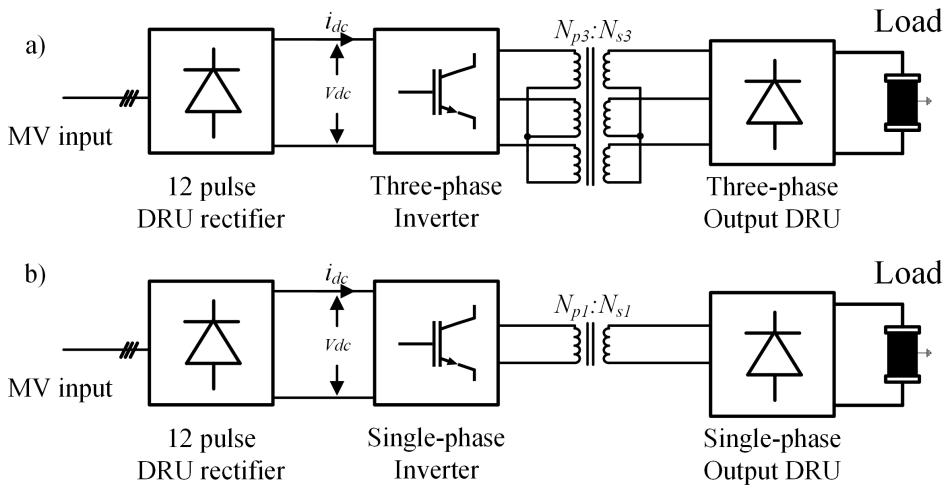


Figure 2.1: (a) Three-phase versus (b) single-phase SST structures. Both employ a 12-pulse DRU, but the three-phase design offers lower winding stress, improved harmonic behavior, and reduced transformer size for high-power applications.

2.3. TOPOLOGY DESIGN

Solid-state transformers (SSTs) can be realized through various topologies, each offering specific advantages and trade-offs in efficiency, modularity, control complexity, and insulation requirements. The choice of topology depends on application-specific constraints and the balance between system size, weight, and operational requirements, particularly for large-scale electrolyzer systems where unidirectional power flow is required.

2.3.1. DUAL ACTIVE BRIDGE (DAB)

The Dual Active Bridge (DAB) converter [54, 55] and resonant converters [56] are commonly used in medium-frequency transformer-based systems due to their high-frequency isolation and controllability. The DAB offers bidirectional power transfer capability, making it suitable for applications requiring power flow in both directions. However, in electrolyzer systems where the power flow is unidirectional, this capability is unnecessary, and the higher number of active switches increases system complexity and cost. Therefore, for unidirectional electrolyzer applications, resonant converters such as LLC are often preferred for their higher efficiency and lower component count [1].

2.3.2. CASCADED STRUCTURE SST

The cascaded structure converts line-frequency AC to medium or high frequency through modular power converters, enabling voltage conversion and galvanic isolation using medium-frequency transformers. Its modularity supports scalability across a range of voltage and current requirements. The Input-Series Output-Parallel (ISOP) configuration is particularly suitable for medium-voltage input to low-voltage output conversion. However, cascaded ISOP configurations require many low-power transformers that must be isolated to withstand high-level DC voltages, leading to increased system volume and cost due to the insulation complexity.

2.3.3. MODULAR MULTI-LEVEL CONVERTER (MMC)

The MMC-based SST topology utilizes central medium-frequency transformers and provides a compact solution with favorable semiconductor and capacitor ratings. In high-power applications, the three-phase MMC-SST reduces transformer size and the RMS current demand on DC bus capacitors, contributing to a compact design. However, MMC structures typically involve a high number of switches and complex control strategies, and they exhibit notable switching losses, especially in medium and high-frequency operations [57].

2.3.4. RESONANT MODULAR MULTI-LEVEL CONVERTER (MMR)

The MMR-SST topology builds on the MMC structure by integrating resonant circuits to enable soft switching, thereby reducing switching losses while maintaining the benefits of modularity and compactness. The MMR topology employs resonant modules in place of single active switches on the primary side and uses configurations that eliminate the need for a twelve-pulse setup due to sinusoidal resonant currents. This approach effectively addresses switching loss limitations in medium and high-frequency applications, making MMR-SST particularly suitable for

electrolyzer systems where unidirectional power flow, high efficiency, and reduced volume and weight are required [1].

2

2.3.5. INPUT-SERIES OUTPUT-PARALLEL TOPOLOGY (ISOP)

The ISOP topology connects DC-DC converter modules in series on the input and in parallel on the output, making it effective for stepping down medium-voltage AC to low-voltage DC with scalability and redundancy [58]. While efficient and modular, the ISOP approach requires numerous isolated transformers capable of withstanding full DC voltage, which significantly increases insulation complexity, system volume, and cost. For unidirectional electrolyzer systems, resonant converters within ISOP structures can improve efficiency, but the insulation and volume challenges remain a limitation for large-scale implementation [1].

2.3.6. SUMMARY AND TOPOLOGY SELECTION

Selecting the appropriate SST topology for large-scale electrolyzer applications involves balancing efficiency, size, weight, insulation requirements, system complexity, and cooling needs. While cascaded ISOP structures provide modularity, they may introduce challenges related to insulation coordination and increased system volume due to series-connected units. MMC-based SSTs can alleviate some of these challenges through multilevel voltage generation, which reduces voltage stress per submodule; however, this benefit depends strongly on modulation strategy and system design. On the other hand, MMR-based converters can reduce switching losses under resonant operation conditions, at the cost of increased control and design complexity.

In this work, these observations are used as qualitative guidelines rather than universally optimized conclusions, and the final topology selection is based on a combination of loss, size, and implementation considerations. This choice aligns with the requirements of unidirectional power flow in electrolyzer systems and the need for compact, efficient, and scalable designs suitable for high-power, medium-frequency applications. Additionally, for the front-end AC to DC conversion, a twelve-pulse diode rectifier unit (DRU) is employed, offering a cost-effective and reliable solution compatible with the SST system.

To address the lack of detailed quantitative comparisons in existing literature, this thesis includes a comprehensive evaluation of MMC, MMR, and ISOP topologies from the perspectives of efficiency, size, weight, and insulation requirements, considering practical factors such as insulation distance and the impact of the cooling system, including chiller size and weight, on overall system design. This detailed analysis supports informed topology selection and ensures that the design aligns with practical implementation considerations in green hydrogen production using SSTs. Three topologies are shown in Figure 2.2.

It should be noted that the above comparisons are used as qualitative design guidelines rather than universally optimized conclusions, as the relative performance of ISOP, MMC, and MMR topologies depends strongly on modulation strategy and operating conditions.

The selection of quasi-square wave operation in this work is primarily motivated by its lower switching losses and simpler implementation compared to fully multilevel waveform synthesis. However, this choice introduces higher harmonic content and potentially different insulation stress conditions. A comprehensive optimization of waveform shape considering efficiency, insulation stress, and device utilization is beyond the scope of this thesis, and is identified as an important direction for future work.

2.4. DESIGN PARAMETERS OF THE COMPLETE SYSTEM

The complete system, which includes two electrolyzers, has a total power rating of 45.4 MW. Based on agreement on standard IEEE519 [59] for this application, the input current ripple to the electrolyzer must be confined to within 3%. Exceeding this threshold can reduce the lifespan of the electrolyzer and adversely affect the rate of hydrogen production. This requirement will be addressed in the subsequent design phase.

2.4.1. 12-PULSE RECTIFIER

To satisfy the IEC 61000-3-6 current harmonics requirements of the Dutch Medium-Voltage grid, a 12-pulse rectifier circuit should be used. To generate the $\pm 15^\circ$ voltage phase shift required for this configuration, a three-winding isolated transformer is typically employed. However, due to the need for galvanic isolation and the use of 50 Hz as the operating frequency, such isolated transformers are bulky and heavy.

To address the size and weight reduction objectives of solid-state transformers, an autotransformer can be utilized in cases where isolation between the windings is not required. The use of an autotransformer enables the generation of the necessary phase shifts with reduced size and weight, aligning with the design goals of compact and efficient SST systems. The vector diagram and the configuration are shown in Figure 2.3.a and b, respectively. Based on the autotransformer vector diagram, the voltage transfer ratio of the 12-pulse autotransformer can be derived as $1/\cos(15^\circ) = 1.0353$.

Following the autotransformer, two 6-pulse diode rectifiers are connected to its output. The voltage transfer ratio of a 6-pulse diode rectifier from the input three-phase AC line-to-line peak voltage to the output DC voltage is given by: $\frac{3}{\pi}$. The diode selection procedure is as follows: first, the diode peak reverse voltage and average current are

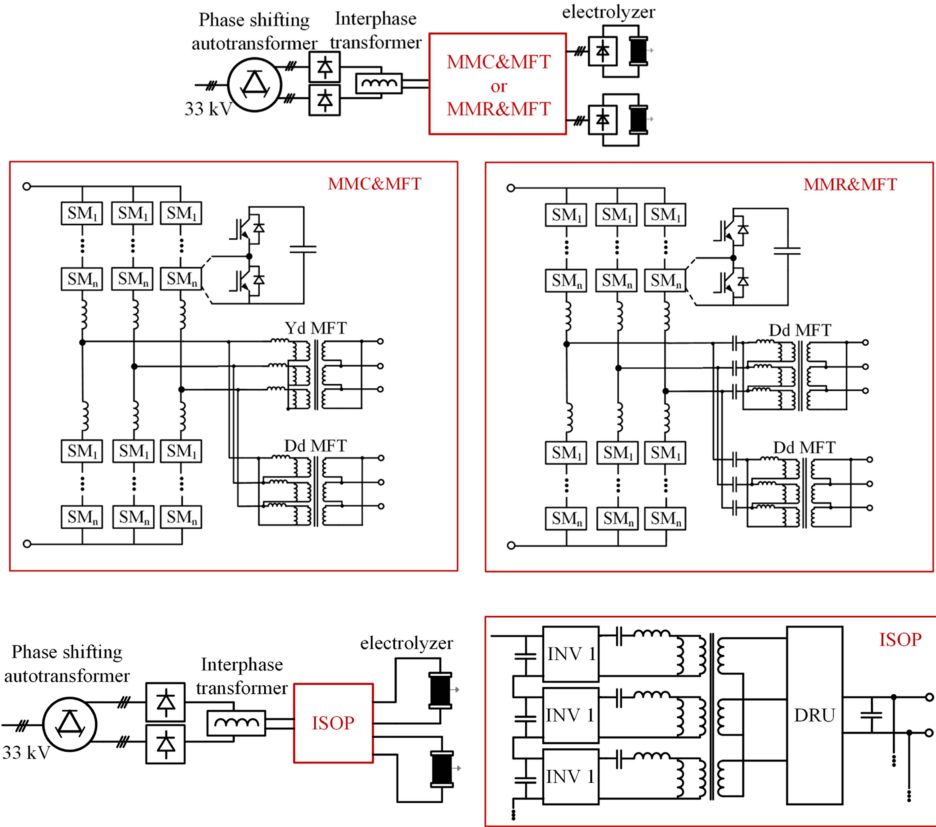


Figure 2.2: a) MMC-based and MMR-based SST configurations. Both systems utilize a DRU setup similar to the cascaded structure SST. The MMC-SST employs a Yd and Dd MFTs configuration, creating a twelve-pulse structure. b) The MMR working principle resembles the two level three phase LLC converter, utilizing two Dd MFTs. Due to the sinusoidal resonant currents of the MMR, which inherently have low THD, the necessity for a twelve-pulse configuration is eliminated. c) Cascaded structure SST for electrolyzer application. Instead of isolated transformer, autotransformer is used to generate the 30° phase shift angle for two DRUs, to save space and weight. The DC-DC cascaded structure uses three phase LLC modules with input series output parallel connection.

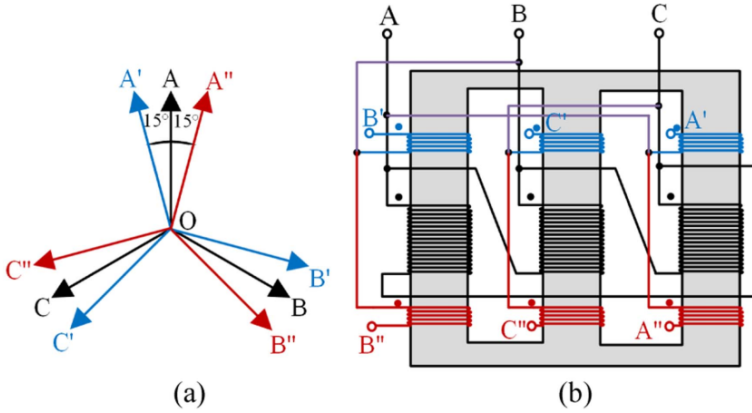


Figure 2.3: (a) Autotransformer vector diagram showing A' and A'' representing the generated phase A voltages with a 30° phase shift. (b) Windings of an autotransformer on a three-limb core.

evaluated through simulation, yielding values of 48.3 kV and 300 A, respectively. Then, a voltage safety factor of 2.25 and a current safety factor of 2 are applied, according to [60]. If the Infineon D471N diode, with a voltage rating of 9,000 V and a current rating of 600 A, is selected, then for each bridge, $\lceil 48.3 \text{ kV} \times 2.25 / 9 \text{ kV} \rceil = 13$ diodes should be connected in series.

The interphase transformer facilitates the parallel connection of two diode bridges, each fed by voltage sources with a phase difference [61]. Specifically, two interphase transformers are needed in our applications: one for each voltage polarity. The configuration of an interphase transformer and its associated connections are depicted in Figure 2.4. As a result, the output DC voltage of the interphase transformer can be calculated as $U_{dc} = 33 \times \sqrt{2} \times \frac{3}{\pi} \times 1.0353 = 46.2 \text{ kV}$.

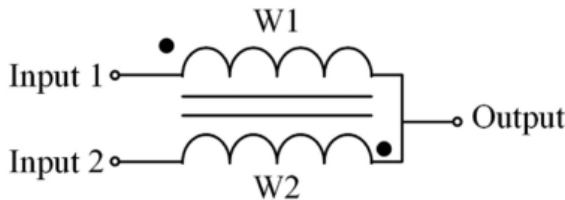


Figure 2.4: Interphase transformer diagram. W1 and W2 represents the two windings of the interphase transformer.

2.4.2. INVERTER AND MEDIUM FREQUENCY TRANSFORMER

The design and implementation of the inverter and medium-frequency transformer (MFT) depend on the selected SST topology. Different topologies require different inverter structures, switching strategies, and transformer designs to achieve the desired voltage conversion, galvanic isolation, and system efficiency.

In this thesis, the MMC-SST, MMR-SST, and ISOP-SST topologies are investigated. Each topology presents specific design considerations for the inverter and MFT, including operating frequency, voltage and current ratings, insulation requirements, and cooling needs. The following sections provide a detailed analysis of the design and performance of the inverter and MFT within these topologies, supporting the overall system requirements for large-scale electrolyzer applications.

The medium-frequency transformer (MFT) is designed based on the guidelines provided in IEC 60076 [62]. The detailed design procedure, including thermal, electrical, and insulation considerations for the MFT in the MMC-SST topology, is presented in Section 3 of this thesis.

MMC

The MMC design follows the design guidelines in [63, 64]. MMC's initially selected fundamental frequency is 400 Hz, but the design process could also be applied to other frequencies.

The maximum modulation index corresponding to the maximum operating point of the electrolyzer is set as 0.9 to leave a certain safety margin. The maximum arm current value according to simulation is 1050 A. With a current safety factor of 2, the 3.3 kV 1.2 kA ABB 5SNA 1200E330100 IGBT with a peak collector current of 2400 A is chosen, and the voltage rating is derated to 2 kV to have a safe margin for potential transient voltages. The number of submodules is calculated as 24. The submodule is half bridge submodule, since in this SST application, the MMC DC side short circuit fault is unlikely to happen compared with HVDC applications. The submodule capacitance is chosen as 1.45 mF to ensure that the submodule voltage variation is within the limit and the IGBT's power loss is minimal.

The arm inductance is calculated based on [64] as 0.9 mH. In this study, the losses attributed to arm inductance are neglected, as they contribute to only approximately 0.02% of the Modular Multi-level Converter's (MMC) rated power.

MMR

As mentioned in the introduction, the differences between MMR and MMC are inserted resonant capacitors and modulation methods. The Quasi Square Wave modulation method for the MMR is depicted in Figure 2.5.

The choice of quasi-square wave operation in this work is primarily motivated by its lower switching losses and simpler control implementation compared to fully multilevel waveform synthesis. However, this choice introduces higher harmonic content and potentially different insulation stress conditions. A comprehensive optimization of waveform shape considering efficiency, insulation stress, and device utilization is beyond the scope of this thesis. Therefore, the selected operating mode should be interpreted as a practical engineering trade-off rather than a globally optimal solution.

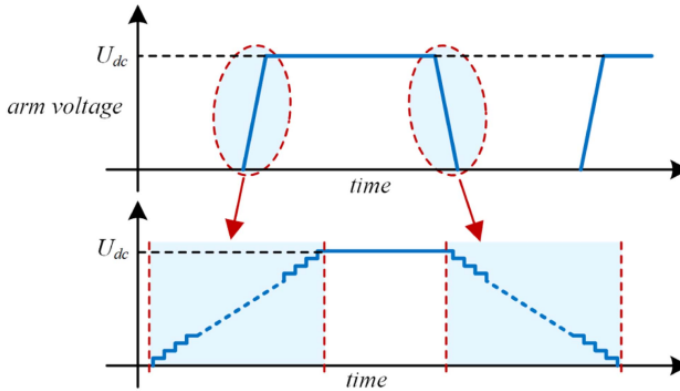


Figure 2.5: Quasi Square Wave modulation method. U_{dc} represents the input DC voltage. The generated staircase waveforms are designed to emulate those from two-level soft-switching converters, ensuring that the operating principles of the MMR align with those of the two-level, soft-switching LLC converters.

In theory, for MMR, primary side switches can achieve ZVS, and secondary side diodes can achieve ZCS, but because of the MMC arm current DC bias, the switching losses will exist. The overall configuration of the MMR with the secondary side DRUs is shown in Figure 2.2.a. In contrast to the MMC-SST, which employs a YdDd medium frequency transformer configuration to mitigate MMC output current harmonics and associated module losses, the MMR adopts a DdDd configuration. This shift is primarily due to the theoretical expectation that resonant and total MMC output currents in the MMR are sinusoidal, eliminating the necessity for a twelve-pulse configuration. Moreover, manufacturing simplicity is achieved as both medium-frequency transformers share the same configuration. Ideally, a YdYd transformer configuration would be preferable for high input voltage to low output voltage applications. However, according to [65], the resonant current in a three-phase Yd resonant converter is not sinusoidal and contains significant harmonics,

adversely affecting MMC loss reduction. According to simulations, the DdDd configuration ensures minimal THD in resonant currents, thus making it the preferred choice for reducing MMC losses. The resonant passive components design methods are referred to two-level LLC converters [66, 67]. The transformer designs will differ from those used in MMC-SST systems. The turn ratio n_1 of the Dd transformer is calculated as follows:

$$n_1 = \frac{U_{dc}}{U_{\text{electrolyzer}}} \quad (2.10)$$

where $U_{\text{electrolyzer}}$ represents the rated input voltage of the electrolyzer. The turn ratio is adjusted slightly to account for output voltage margin, resulting in a final turn ratio of 52 for the Dd transformer. The parameters of the MMR modules, such as submodule capacitance and the number of submodules, remain the same as those used in the MMC-SST configuration.

For the power converter loss estimation, the front-side twelve-pulse DRU losses remain the same as in the MMC-SST configuration. However, for the MMC within the MMR, the switching losses will differ due to the soft-switching mechanism, and the conduction losses will also vary because of the different current waveforms. The following section presents a theoretical calculation of the losses in the MMR. For the Dd transformer, the peak values of the secondary-side current and voltage can be calculated as follows:

$$U_{\text{sec-ph-peak}} = U_{\text{sec-LL-peak}} = \frac{\pi}{3} \times U_{\text{electrolyzer}} \quad (2.11)$$

$$I_{\text{sec-line-peak}} = \sqrt{3} \times I_{\text{sec-ph-peak}} = \frac{\sqrt{3} \times P_{\text{total}} \times 2}{U_{\text{sec-ph-peak}} \times 3} \quad (2.12)$$

The primary side voltage and current are calculated as follows:

$$U_{\text{pri-ph-peak}} = U_{\text{sec-ph-peak}} \times n_1 \quad (2.13)$$

$$I_{\text{pri-line-peak}} = \sqrt{3} \times I_{\text{pri-ph-peak}} = \frac{\sqrt{3} \times P_{\text{total}} \times 2}{U_{\text{pri-ph-peak}} \times 3} \quad (2.14)$$

As a result, the primary side line current peak value is calculated as 702 A. The total output current of MMC is:

$$I_{\text{total}} = 1404 \cos(\omega t + \phi_1) \quad (2.15)$$

Thus, the current is in phase with the MMC output square waves. The upper and lower arm currents of phase A are:

$$i_p = \frac{I_{dc}}{3} + \frac{I_{total}}{2} = 328 + 702 \cos(\omega t) \quad (2.16)$$

$$i_n = \frac{I_{dc}}{3} - \frac{I_{total}}{2} = 328 - 702 \cos(\omega t) \quad (2.17)$$

ISOP

As discussed in the introduction, LLC is more suitable than DAB for this unidirectional application scenario. At the initial stage of the design, the single-phase LLC was chosen as the submodule for the ISOP. Although this topology offers better scalability, it has notable disadvantages. Firstly, single-phase transformers are heavier and larger compared to three-phase transformers with the same voltage level and capacity. Secondly, the output capacitance is large due to the intrinsic second-order power ripple present in single-phase systems.

The ISOP topology is shown in Figure 2.2 [1, 58]. Full-bridge inverters are used due to the high-power requirements of the modules. The voltage rating of the switches remains the same as in the MMC-SST configuration, utilizing 3.3 kV IGBTs with a derated safety voltage of 2 kV. The number of cascaded full-bridge modules is calculated as $N = \lceil U_{dc}/U_{IGBT,derated} \rceil = 24$.

Instead of using the variable frequency modulation method to change the output voltage, the phase shift modulation (PSM) method is adopted in this study. The delay angle between the leading bridge and the following bridge is controlled to modify the output voltage [68]. In this way, the transformer design is less challenging than variable frequency modulation, because the transformer needs over-design for the latter to cover the whole operating frequency range.

Since the full bridges are grouped in sets of three, there are eight three-phase LLC converters whose inputs are connected in series. Four of these three-phase LLC converters are output parallel connected to one electrolyzer, and the remaining four are connected to another. The designed resonant inductance is 0.11 mH, the magnetizing inductance is 38 mH, and the resonant capacitance is 0.22 mF, resulting in a resonant frequency of 1000 Hz. The resonant inductance and magnetizing inductance are determined by the medium-frequency transformer. Following the same switch selection methodology used in the MMC-SST topology, the primary-side inverter switches are selected as 5SNA 1800E330400, with a voltage rating of 3.3 kV and a current rating of 1.8 kA, based on the measured current flowing through the switches. The secondary-side diode is selected as 5SDF28L4521, with three diodes connected in parallel to handle the high currents. The input DC capacitance for each full-bridge inverter is set to 2.4 mF to limit voltage ripple, while the output capacitance is set to 32 mF to meet the

electrolyzer's 3% current variation requirement. Detailed calculations are provided in [1].

2

2.4.3. MEDIUM FREQUENCY DRU RECTIFIER

Two 6-pulse medium-frequency diode rectifier units (DRUs) are connected to the output of the MFT. The calculated peak reverse voltage is 800 V, and the average current is 12 kA. Applying a voltage safety factor of 2.25 and a current safety factor of 2, the 5SDF28L4521 medium-frequency diode, with a voltage rating of 4,500 V and a current rating of 2,620 A, is selected. For each bridge, a total of 10 diodes should be connected in parallel. To ensure the electrolyzer input current ripple remains within the required 3%, a 50 μ H inductor is integrated prior to the connection with the electrolyzer.

For the MMR case, because of the working principle of the resonant converter, there is no reverse recovery of the diodes, thus, only the conduction losses of the diodes are considered. Because the primary side resonant current is pure sinusoidal at the resonant frequency, and the magnetizing inductance is much higher than the leakage inductance of the MFT, the secondary side current is also assumed to be sinusoidal. The conduction losses calculation method of the secondary side DRU is the same as that of MMC-SST. As a result, the conduction losses, switching losses of MMC, and the conduction losses of secondary side DRU are 191 kW, 68 kW, and 341 kW, respectively, for 400 Hz. For 1000 Hz, only the switching losses of the MMC will change to 170 kW, and others will remain unchanged [1].

The design of medium-frequency transformer will be detailed in Section 3.

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3

MEDIUM-VOLTAGE MEDIUM-FREQUENCY TRANSFORMER

*To see a World in a Grain of Sand
And a Heaven in a Wild Flower,
Hold Infinity in the palm of your hand
And Eternity in an hour.*

William Blake

This chapter presents the design, modeling, and performance evaluation of medium-voltage medium-frequency transformers (MV-MFTs) for solid-state transformer (SST) applications in large-scale green hydrogen production. It reviews relevant international standards for insulation, harmonics, and thermal performance, and details the transformer design process for three converter topologies: MMC-SST, MMR, and ISOP. For each case, magnetic core selection, winding configuration, FEM-based loss estimation, and thermal management are discussed. Comparative analysis across different operating frequencies is performed to identify the optimal frequency for each topology based on loss, size, and weight, with a focus on practical design trade-offs.

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R. Mirzadarani, M. Ghaffarian Niasar, Z. Li, Z. Qin, P. Vaessen, P. Bauer, L. Van Lieshout, "Three-Phase Medium-Voltage Medium-Frequency Transformer for SST in Green Hydrogen Production," IECON 2023- 49th Annual Conference of the IEEE Industrial Electronics Society, Singapore, Singapore, 2023, pp. 1-6 [2]

3.1. APPLIED STANDARDS

The design and analysis of medium-voltage medium-frequency transformers require compliance to relevant international standards to ensure safety, reliability, and performance. In this section, the standards used for the design, testing, and validation of the transformer are introduced. These standards provide guidelines for parameters such as insulation coordination, temperature rise limits, partial discharge measurement, and test procedures. Following these standards ensures that the transformer design aligns with industrial requirements and facilitates comparison with other designs reported in the literature. It is noteworthy that although the proper standards for SSTs do not exist, it is possible to get inspiration from existing standards for this application.

3.1.1. IEC 60076

IEC 60076 is a series of international standards that specifies requirements for the design, testing, and operation of 50/60 Hz transformers. This standard series provides general guidelines for power transformer construction, insulation coordination, temperature rise limits, and dielectric testing. In this study, IEC 60076 is applied to ensure that the design and operation of the medium-voltage medium-frequency transformers comply with established industry practices. The standard is used for both oil-immersed and dry-type transformers, supporting insulation coordination considerations and validating the thermal and electrical performance of the transformers under normal and specified conditions.

Here is the list of IEC 60076 standards relevant to the design, testing, and application of medium-voltage medium-frequency transformers:

- IEC 60076-1 [3]: General
- IEC 60076-2 [4]: Temperature rise for liquid-immersed transformers
- IEC 60076-3 [5]: Insulation levels, dielectric tests, and external clearances in air
- IEC 60076-4 [6]: Guide to the lightning impulse and switching impulse testing
- IEC 60076-5 [7]: Ability to withstand short circuit
- IEC 60076-7 [8]: Loading guide for oil-immersed power transformers
- IEC 60076-8 [9]: Application guide
- IEC 60076-11 [10]: Dry-type transformers
- IEC 60076-12 [11]: Loading guide for dry-type power transformers

3.1.2. IEC 60270

Partial discharge (PD) measurement is a critical diagnostic and quality assessment tool for high-voltage equipment, providing insight into insulation integrity and aging behaviour. IEC 60270 [12] defines the standard methods for measuring partial discharges under AC and DC voltages and establishes the terminology, test circuits, measurement procedures, and calibration requirements for consistent and reliable PD measurements. In this thesis, IEC 60270 is used to guide the measurement and interpretation of partial discharges in medium-voltage, medium-frequency transformers, ensuring that the insulation design and condition assessment meet internationally recognized procedures during design validation and performance testing.

3.1.3. IEEE 519

IEEE 519-2022 [13] is a standard that defines voltage and current harmonic distortion limits in electrical power systems, particularly at the point of common coupling (PCC) between users and system operators. This standard is used in this thesis to set and evaluate harmonic control objectives in the design and analysis of solid-state transformer systems and their power electronic interfaces. IEEE 519 is chosen over the IEC documents for harmonic control because it is widely adopted in power electronic system design and research communities, offering detailed, practical guidelines that align with the harmonic management needs in power converter and SST applications.

3.2. TRANSFORMER DESIGN

The transformer designs presented in this chapter follow the system design requirements detailed in Chapter 2. Three separate designs are considered to match the needs of the MMC-SST, MMR, and ISOP cases as described in the system architecture and operating conditions in Chapter 2. The designs are developed in accordance with the standards introduced in the previous section to ensure that insulation coordination, thermal management, and operational performance meet internationally recognized guidelines and testing procedures.

3.2.1. DESIGN OF THE MEDIUM-FREQUENCY TRANSFORMER FOR MMC-SST

In order to reduce the MMC harmonics according to IEEE 519 [13], one transformer is configured as Yd, and the other is configured as Dd. The delta primary gives a path for zero sequence triple (3rd, 9th, 15th, ...) harmonics currents to flow. These harmonics are always drawn because of the non-linear magnetizing (B-H) curve of the transformer

core material. Without a delta winding on the transformer these currents will flow through the lines, adding to their losses without contributing to power transfer. The third-harmonic components are in phase in all three phase voltages and therefore cancel in the line-to-line voltage. The values of the voltages (primary and secondary) and currents are first calculated manually and then verified according to Simulink model. The corresponding primary currents are 580 A for the Yd type and 338 A for the Dd type.

3

YD TRANSFORMER DESIGN

First, the voltage transfer ratio of the transformers is calculated. The MFT primary side line-to-line RMS voltage and the secondary side line-to-line RMS voltage are calculated according to:

$$U_{\text{pri-LL-RMS}} = \frac{U_{dc}}{2} \times m \times \sqrt{\frac{3}{2}} \quad (3.1)$$

$$U_{\text{sec-LL-RMS}} = U_{\text{electrolyzer}} \times \frac{\pi}{3} \times \frac{1}{\sqrt{2}} \quad (3.2)$$

where m is the MMC modulation index and $U_{\text{electrolyzer}}$ is the electrolyzer input voltage. The theoretical voltage transfer ratio of the MFT is calculated as 50. However, non-ideal commutation during switching reduces the effective transformer excitation voltage; therefore, a lower voltage transfer ratio of 44.5 is selected to meet the required electrolyzer voltage. The number of turns on the secondary side must be selected as low as possible. This is due to efficiency considerations, considering the high current on the secondary side. Practically, it is not feasible to select $N_s = 1$ as continuously transposed conductors (CTCs) require a minimum length to ensure all strands are appropriately positioned. Therefore,

$$N_s = 2 \rightarrow N_p = 52 \quad (3.3)$$

where N_s is the number of secondary turns, and N_p is the number of primary turns. The primary line-to-line voltage is $U_{\text{pri-LL-RMS}} = 25429$ V, and the phase voltage of the primary RMS value is $U_{\text{pri-ph-RMS}} = 14615$ V.

The following is a brief explanation of core material selection and design. The cross-section of the core can be calculated as follows:

$$A_c = \frac{U_{\text{pri-ph-RMS}}}{k_c \times k_w \times B_m \times f \times N_p} \quad (3.4)$$

where k_c is the core stacking factor, k_w is the form factor, B_m is the maximum flux density in the core, and f is the frequency.

Different core materials are feasible. For high-power applications, grain-oriented electrical steel and amorphous are more common choices [2]. The characteristics of electrical steel and amorphous are compared in Table 3.1 [14, 15]. According to Table 3.1, electrical steel is a good choice for 50/60 Hz applications. However, the higher the frequency, the higher the core loss. Therefore, alternative materials such as amorphous are more beneficial.

Table 3.1: Comparison of the Characteristics of Electrical Steel and Amorphous Alloys

Material	Layer thickness (μm)	Loss @50 Hz (W/kg)	Loss @400 Hz (W/kg)	B_{sat} (T)	μ_r	Density (kg/m^3)
Grain-oriented electrical steel	100	1.6	16	2.00	2000	7800
Amorphous	25	0.14	1.3	1.63	6500	7300

Another difference between a transformer made of electrical steel and the one made of amorphous is that electrical steel transformers are made of several plates for the limbs and yokes. However, amorphous transformers are made by a number of thin strips which are placed next to each other to form the entire core. From a manufacturing point of view, to insert the coils into an electrical steel core, the manufacturer removes the upper yoke, inserts the coils, and puts the upper yoke again [16, 17]. For an amorphous core, as shown in Figure 3.1, the manufacturer unlaces the strips, inserts the coils, and re-laces the strips again [18]. Although the manufacturing procedure is more complex for the amorphous-based transformers, amorphous is selected as the magnetic material of the core since the frequency in the present application is relatively high. The selected alloy is Metglas 2605HB1M, which is available in strips with various widths, such as 142.2 mm, 170.2 mm, and 213.4 mm [15]. The nominal thickness of the strips is 25 μm , which results in a good lamination.

Assuming $B_m = 1.2$ T, $k_c = 0.79$, $k_w = 4.44$, and $f = 400$ Hz in 3.4 results in $A_c = 0.167$ m². For MMC-SST topology, amorphous strips with a width of 213.4 mm are selected to maintain a safety margin and avoid any possibility of core saturation due to higher frequency phenomena. Therefore, the width of the core is calculated as 392 mm, as shown in Figure 3.2.a. The core is made of two cores stacked on top of each other to use the standard strips. The core dimensions are shown in Figure 3.2.b. The core loss of the transformer is calculated as follows [22]:

$$P_c = \text{Volume}_c \times \varrho_c \times P_{c0} \times \left(\frac{f}{f_0}\right)^\alpha \times \left(\frac{B_m}{B_{m0}}\right)^\beta \times \left(\frac{k_w}{k_{w0}}\right)^{2(\alpha-1)} \quad (3.5)$$

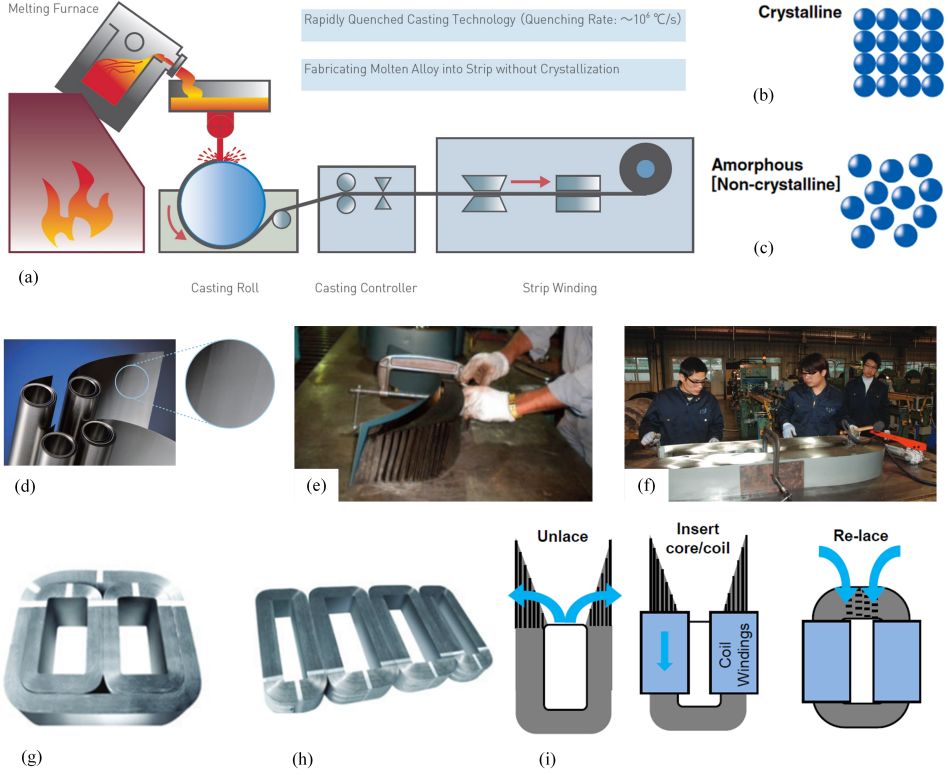


Figure 3.1: Overview of the amorphous core production and assembly process [19–21]: (a) melt spinning process for producing amorphous ribbons, (b) crystalline atomic structure, (c) amorphous atomic structure, (d) amorphous ribbon rolls, (e) core stacking process, (f) assembly line, (g)–(h) core shapes, (i) unlacing, inserting coil, and replacing process.

where Volume_c is the volume of the core, ρ_c is the density of the core, P_{c0} is the loss of the core per kilogram according to Table 3.1, f_0 is the reference frequency (400 Hz), B_{m0} is the reference maximum flux density (1.0 T), k_{w0} is the reference form factor (4.44), $\alpha = 1.47$ is the frequency dependency, and $\beta = 1.52$ is the flux density dependency.

Besides core losses, transformers are affected by “stray losses”, which stem from eddy currents induced in the transformer’s conductive components, such as clamps, flitch plates, tanks, and shunts [23]. The calculation of stray losses is a complex task, requiring the utilization of Finite Element Method (FEM) simulations or experimental estimation techniques to obtain accurate measurements [24, 25]. In this paper,

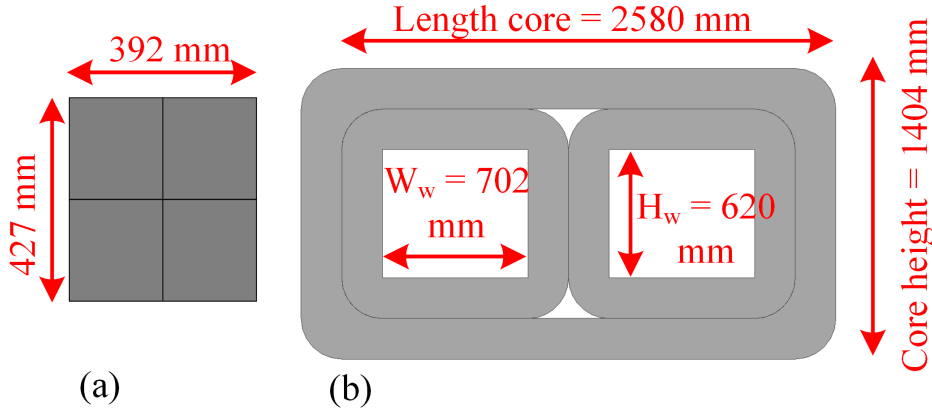


Figure 3.2: Details of the core design for the MMC-SST transformer: (a) the cross-section of the core, (b) the 3D view of the magnetic core.

stray loss is estimated based on the method described in [26].

After the core design is accomplished, the CTCs are selected for the conductors of the transformer. CTCs have two advantages. Firstly, CTCs can mitigate the skin effect due to the presence of several strands. Secondly, with the transposition of the strands, they all have similar lengths and experience similar background magnetic fields to avoid circulating current in the windings [27]. The primary CTC size CTC_{pri} and secondary CTC size CTC_{sec} can be designed as follows:

$$A_p = \frac{I_{pri-ph-RMS}}{J_p} = \frac{580}{2.5} = 232 \text{ mm}^2 \Rightarrow CTC_{pri} : 15 \times 8 \times 2 \text{ mm}^2 (n \times w \times t) \quad (3.6)$$

$$A_s = \frac{I_{sec-ph-RMS}}{J_s} = \frac{15200}{2} = 7600 \text{ mm}^2 \quad (3.7)$$

$$\Rightarrow CTC_{sec} : 13 \times 15 \times 19 \times 2 \text{ mm}^2 (N \times n \times w \times t)$$

where A_p and A_s are the required primary and secondary conductor sizes, respectively; $I_{pri-ph-RMS}$ and $I_{sec-ph-RMS}$ are the phase current RMS values of the primary and secondary, respectively; Based on typical design practice for oil-immersed MFT windings with forced cooling [2, 27], the maximum allowable current densities are chosen as $J_p = 2.5 \text{ A/mm}^2$ for the primary and $J_s = 2 \text{ A/mm}^2$ for the secondary; N is the number of parallel paths on the secondary side to handle large currents; n is the number of strands (which is an odd number); w is the width of each strand; and t is the thickness of each strand.

The primary and secondary conductors are shown in Figure 3.3.a and Figure 3.3.b, respectively. The transformer window is designed as shown in Figure 3.3.c. Note that because of high-current busbars, the secondary is built over the primary. The primary is made of 26 disks, and each has two which are connected in series. The secondary is made of 13 parallel disks to handle the high-current. While the distance between disks in the primary is for the oil ducts, the secondary has more space between disks to provide access to the busbar connection.

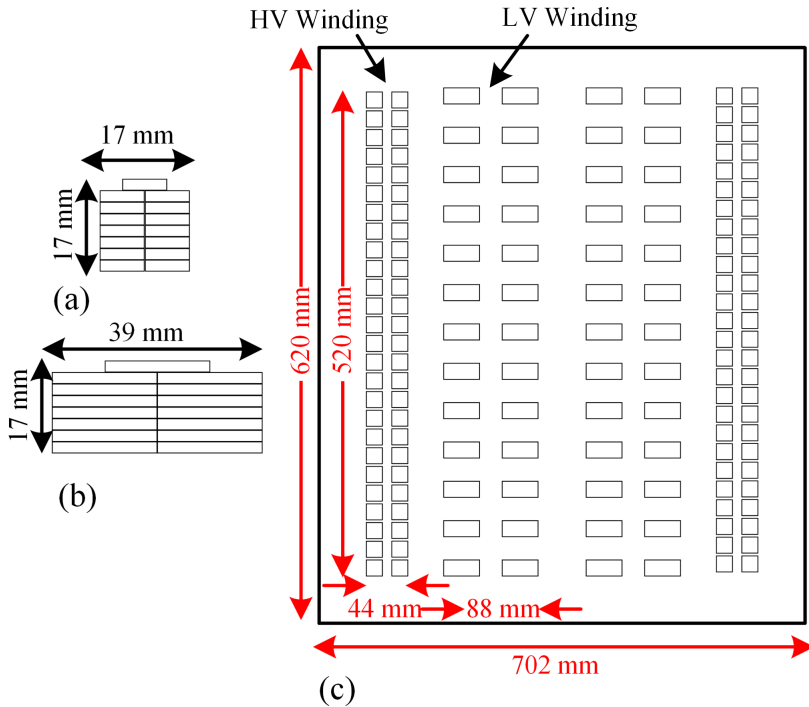


Figure 3.3: The details of the winding design of the MMC-SST transformer: (a) Yd CTC of primary, (b) Yd CTC of secondary, (c) Yd window.

A FEM simulation using a 3D ANSYS Maxwell model was performed to determine the leakage and magnetizing inductances, crucial for voltage impedance calculations and the resonant topology designs. Additionally, a 2D COMSOL model was used to compute the AC load losses, a key factor in assessing transformer efficiency and informing cooling system design. The results of the FEM simulations for MMC-SST MFT are summarized in Table 3.2. The meshing performance of the FEM simulation is shown in Figure 3.4. The 3D view of the MMC-SST MFT in ANSYS Maxwell is shown in Figure 3.5.

The simulation procedure and calculation of the parasitic parameters

are explained in the appendix. The calculations for power electronic parts of the SST are detailed in [1].

Table 3.2: FEM Results for MMC-SST

Configuration	Parameter	Value
Yd	L_{lk}	1.4 mH
	r_p	816.12 $\mu\Omega/\text{m}$
	r_s	50.36 $\mu\Omega/\text{m}$
Dd	L_{lk}	4.00 mH
	r_p	1549.80 $\mu\Omega/\text{m}$
	r_s	51.46 $\mu\Omega/\text{m}$

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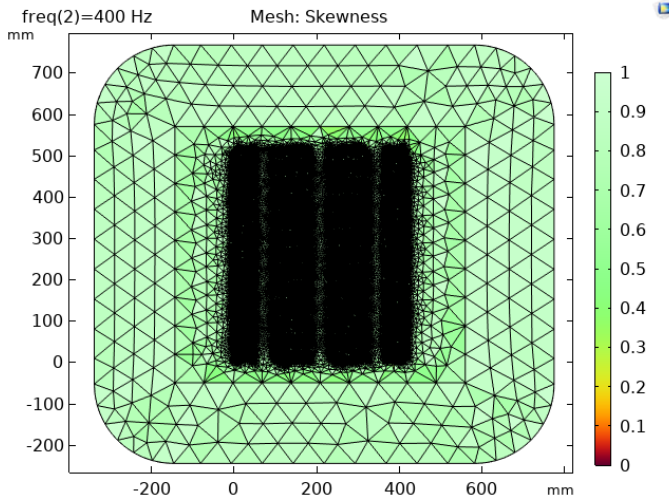


Figure 3.4: Meshing skewness of the COMSOL 2D simulation for Yd MMC-SST MFT, indicating the quality of meshing operation.

To estimate the copper loss of the transformer, first the Mean Length per Turn (MLT) is calculated as follows:

$$\text{MLT}_p = \pi \times (D_p + W_p) \quad (3.8)$$

where D_p is the diameter and W_p is the width of the primary winding. Then, the length of the primary can be calculated as follows:

$$l_p = N_p \times \text{MLT}_p \quad (3.9)$$

According to the FEM simulation result in Table 3.2, the average AC resistance r_p is 816.12 $\mu\Omega/\text{m}$. The load loss based on the AC resistance can be calculated as follows:

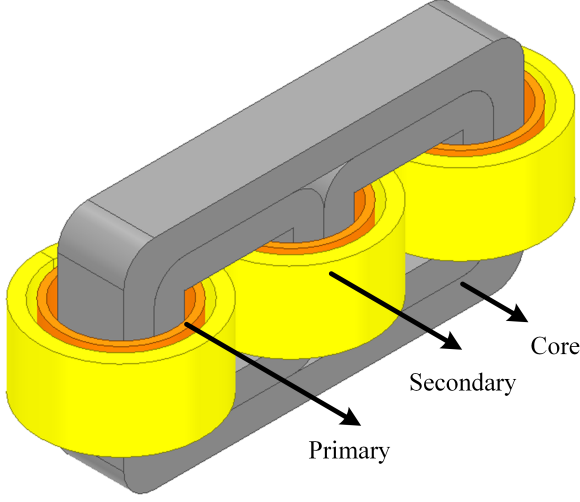


Figure 3.5: 3D view of the MMC-SST MFT.

$$R_p = l_p \times r_p \quad (3.10)$$

$$\text{Loss}_p = 3 \times R_p \times I_{\text{pri-ph-RMS}}^2 \quad (3.11)$$

The weight of the primary winding can be calculated as follows:

$$\text{Weight}_p = 3 \times l_p \times \text{CTC}_{\text{pri}} \times \rho_{\text{cu}} \quad (3.12)$$

where ρ_{cu} is the density of copper and CTC_{pri} is the area of primary conductor. The loss and weight of the secondary winding can be calculated using a similar method:

$$\text{Loss}_s = 3 \times R_s \times I_{\text{sec-ph-RMS}}^2 \quad (3.13)$$

$$\text{Weight}_s = 3 \times l_s \times \text{CTC}_{\text{sec}} \times \rho_{\text{cu}} \quad (3.14)$$

DD TRANSFORMER DESIGN

The design of the transformer for the Dd configuration is similar to the design of the Yd transformer. The difference is in the primary voltage, and in this case, it is $U_{\text{pri-LL-RMS}} = U_{\text{pri-ph-RMS}} = 25429 \text{ V}$. To maintain the same voltage on the secondary side, the number of turns on the primary side is higher (90 turns compared to 52 turns). Accordingly, the

primary current RMS value is 338 A. Therefore, the primary CTC can be designed as $CTC_{pri} = n \times w \times t = 13 \times 5 \times 2 = 130 \text{ mm}^2$. The primary consists of 30 disks, which have 3 turns each. The secondary structure is the same as the Yd transformer. The AC resistances per unit length for the Dd transformer are $r_p = 1549.8 \mu\Omega/\text{m}$ and $r_s = 51.544 \mu\Omega/\text{m}$, as summarized in 3.2.

INSULATION CONSIDERATIONS, THERMAL MANAGEMENT, AND TANK DESIGN

Given the power rating and associated losses, the MFT in this study is implemented as an oil-immersed transformer using mineral oil. The temperature rise limits according to IEC 60076-2 [28] are summarized in Table 3.3. Therefore, the oil temperature for the loss calculations in the COMSOL 2D simulation is considered as $T_{oil} = 78 + T_{amb}$, where T_{amb} is the ambient temperature. In accordance with IEC standard test conditions, the ambient temperature is taken as $T_{amb} = 20^\circ\text{C}$. The MFT for MMC-SST topology consists of two active parts: a Yd and a Dd transformer, in one tank.

Different cooling methods for liquid-immersed transformers are expressed using a four-letter code as defined in IEC 60076-2 [28]. Each letter describes the cooling media and the circulation mechanism.

The first letter specifies the internal cooling medium. *O* denotes mineral oil or a synthetic insulating liquid with a fire point $\leq 300^\circ\text{C}$. *K* denotes an insulating liquid with a fire point $> 300^\circ\text{C}$. *L* denotes an insulating liquid with no measurable fire point.

The second letter specifies the circulation mechanism of the internal cooling medium. *N* denotes natural thermosiphon flow through the cooling equipment and within the windings. *F* denotes forced circulation through the cooling equipment with thermosiphon flow within the windings. *D* denotes forced circulation through the cooling equipment, directed from the cooling equipment into at least the main windings.

The third letter specifies the external cooling medium. *A* denotes air, and *W* denotes water.

The fourth letter specifies the circulation mechanism of the external cooling medium. *N* denotes natural convection, and *F* denotes forced circulation using fans or pumps.

Accordingly, the selected cooling designation **OFWF** indicates oil as the internal cooling medium with forced circulation, and water as the external cooling medium with forced circulation, as shown in Figure 3.6. This cooling system is selected because the MFT exhibits relatively high loss density while maintaining a compact physical size, making OFWF cooling more suitable due to its high heat removal capability and effectiveness in controlling winding and oil temperatures.

To avoid complexity, it is assumed that the oil flow inside the tank has no eddy patterns. The oil flow between disks is assumed to be laminar. This assumption is considered reasonable for the present design

Table 3.3: Temperature Rise Limits According to IEC 60076-2 [28]

Components	Temperature rise limit (K)
Top insulation liquid	60
ON (Oil-Natural) and OF (Oil-Forced) cooling systems – average winding	65
Hot-spot winding	78

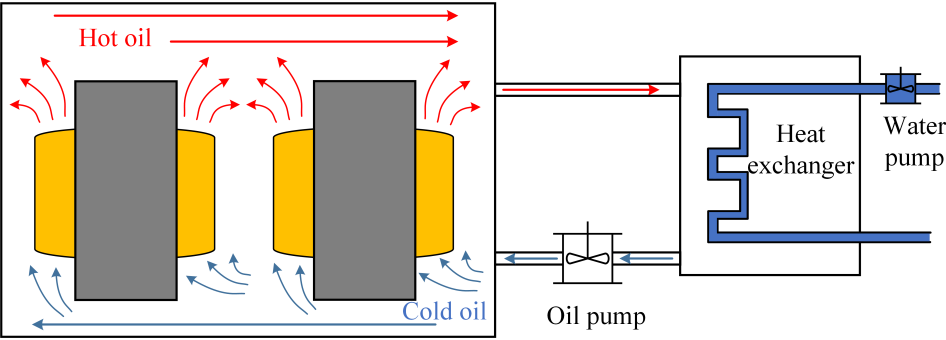


Figure 3.6: Schematic of an oil-forced, water-forced (OFWF) cooling system.

because the cooling is forced and guided by the OFWF system, and the oil flow in winding ducts is typically laminar or weakly transitional under nominal operating conditions. Possible local recirculation or secondary flow effects mainly influence local hot-spot temperatures, while their impact on the average temperature distribution and overall thermal resistance, which are the focus of this model, is limited. The thermal design follows the methods described in [29–31]. The active parts inside the tank are shown in Figure 3.7. As shown in Figure 3.7, The spatial layout of the magnetic components includes the following distances: the distance between each MFT and the side wall of the tank (d_1), the distance between the two MFTs ($2 \times d_2$), the vertical distance between the MFTs and the upper lid of the tank (d_3), and the distance between the MFTs and the bottom of the tank (d_4).

The CTC strands are enameled, and the CTC itself is insulated by oil-immersed paper. A pressboard makes the disk structure support the winding structure during normal operation and fault forces. The windings are made of disks, which are more compact and provide better voltage distribution, as shown in Figure 3.8.

The rated voltage of the transformer is calculated using Equation 3.1 which is 25.429 kV_{rms}. For insulation design, the input voltage is considered as $U_{pri-LL-RMS} = 25.5$ kV. Based on IEC 60076-3, the test voltage levels for transformers with the highest voltage for equipment winding between 24 kV and 36 kV are as follows: Full Wave Lightning

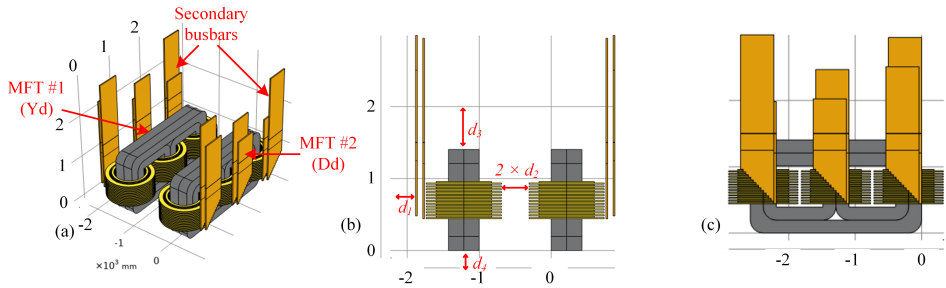


Figure 3.7: 3D view of the active parts inside the tank (a) overall view, (b) side view 1, and (c) side view 2. The high-current busbars on the secondary side are designed to facilitate the parallel connection of the secondary disks and provide sufficient space for the horizontal busbars outside the tank.

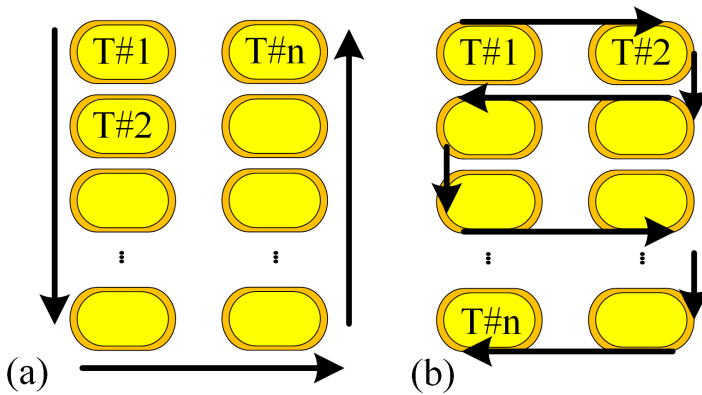


Figure 3.8: (a) A layer winding and (b) a disk winding. T#1, T#2, and T#n represent the 1st turn, 2nd turn, and nth turn, respectively.

Impulse (LI) = 170 kV, Chopped Wave Lightning Impulse (LIC) = 187 kV, and Line Terminal AC Withstand (LTAC) = 70 kV [5]. However, for the present design, the lightning Impulse test is not considered because there is no risk of lightning in this system for two reasons. Firstly, the MFT is not connected to an exposed transmission line. Secondly, there are front-end components like the autotransformer, DRUs, and MMC before the transformer, which are protected against lightning. Therefore, the MFT is designed to handle the Terminal AC Withstand (LTAC) test, which is 70 kV_{rms}. Consequently, the insulation design of the MFT is based on the line terminal AC withstand (LTAC) requirement of 70 kV_{rms}. The maximum transient overvoltage at the transformer terminals is limited by surge arresters to a level well below the standard lightning impulse test voltages, ensuring that the electric stress within the transformer insulation remains within the intended design margins. The cores and the tank are earthed, thus they have zero potential. Because of the presence of high-current busbars of the low-voltage winding, the primary winding, which is connected to medium voltage (MV), is placed near the core, and the secondary winding is placed over the primary. This means the highest electric field appears between the primary and the core, especially near the edges of the core.

OVERALL METRICS OF THE MMC-SST MFT

The transformer dimensions are 3232 × 1044 × 1404 mm (x × y × z). Thus, the volume of the active parts is 9.48 m³. The thickness of the tank plates is selected as $t_{\text{tank}} = 5$ mm. Thus, the tank size is calculated as:

$$X_{\text{tank}} = x_{\text{trans}} + (2 \times d_1) + (2 \times t_{\text{tank}}) \quad (3.15)$$

$$Y_{\text{tank}} = (2 \times d_1) + (2 \times d_2) + (2 \times y_{\text{trans}}) + (2 \times t_{\text{tank}}) \quad (3.16)$$

$$Z_{\text{tank}} = d_3 + d_4 + z_{\text{trans}} + (2 \times t_{\text{tank}}) \quad (3.17)$$

where x_{trans} , y_{trans} , and z_{trans} represent the dimensions of the transformer, and the clearances d_1 through d_4 are defined as shown in Fig. 3.7. Therefore, the volume of the tank is calculated as follows:

$$V_{\text{tank}} = X_{\text{tank}} \times Y_{\text{tank}} \times Z_{\text{tank}} = 3.442 \times 2.498 \times 1.814 = 15.6 \text{ m}^3 \quad (3.18)$$

The weight of the tank and the transformer oil are 1521 kg and 5447 kg, respectively. As a result, the detailed loss, total size, and weight breakdown of MFTs are shown in Table 3.4 and Table 3.5. The total size, weight, and loss of the transformers of all three topologies are summarized in Table 3.6. In addition, the loss, size, and weight of power

converter parts are also shown. Besides the size and weight information of power converter and MFTs, the chiller's information is also presented.

Table 3.4: MMC MFT Detailed Losses. The Copper Loss is Calculated Based on (3.11), and the Core Loss is Calculated Based on (3.5).

Copper loss (kW)	Core loss (kW)	Stray loss (kW)	Total losses (kW)
630.1	26.2	39.3	695.6

3

Table 3.5: MMC MFT Detailed Size and Weight

Transformer size (m ³)	Copper weight (ton)	Core weight (ton)	Tank weight (ton)	Oil weight (ton)
15.6	3.85	16.79	1.52	5.45

Table 3.6: Total Loss, Size, and Weight of SST. Power Converter Loss Includes Front Side DRU Loss, MMC/ISOP Loss, and Medium Frequency DRU Loss. Converter Size Represents the Converter Tower Size Considering Insulations. The Transformer Size Represents the Total Size of the Tank. The Transformer Weight Includes Active Parts Weight, Tank Weight, and Oil Weight.

Topologies	Conv. loss (kW)	Trans. loss (kW)	Conv. size (m ³)	Trans. size (m ³)	Chiller size (m ³)	Conv. weight (ton)	Trans. weight (ton)	Chiller weight (ton)
MMC-SST (400Hz)	888.5	695.6	43	15.6	78	7.2	27.6	14.6
MMR (1000Hz)	751.5	817	39.9	10.44	77.45	5.76	16.35	14.5
ISOP (1000Hz)	481.5	806.3	8.94	13.78	64.24	2.4	28.8	11.98

To cool down the cooling liquid used in the heatsink and the medium frequency transformer, a chiller is needed. The chiller size and weight are considerable and need to be added. According to the product data from DeltaTemp [32], it is possible to derive curve-fitting models that correlate size and weight with cooling capacity. Curve-fitting expressions can be formulated as follows:

$$chiller_size \text{ (m}^3\text{)} = 4.7 \times 10^{-5} \times P \text{ (W)} + 6.015 \quad (3.19)$$

$$chiller_weight \text{ (kg)} = 8.974 \times 10^{-3} \times P \text{ (W)} + 870.4 \quad (3.20)$$

The total power loss P requiring liquid cooling amounts to about 15345 kW, comprising 839 kW from converter loss (MMC and secondary DRU) and 695.6 kW from MFT loss. Thus, the required chiller cooling capacity is 1534.6 kW, and its corresponding size and weight are estimated to be 78 m³ and 14.6 tons, respectively.

3.2.2. DESIGN OF THE MEDIUM FREQUENCY TRANSFORMER FOR MMR

The MMR topology employs two Dd MFTs. The voltage and current of the transformers are measured in the Simulink model. For the Dd transformer used in the MMR topology, $\int U_{\text{pri-ph}} dt = 9.28$, with $N_p = 104$ and $N_s = 2$. The remaining design process follows the same procedure as the MFTs in the MMC-SST and is therefore not repeated here.

The physical structure of MMR MFT is similar to MMC-SST MFT, where there are also two active parts inside one tank, and rating power and voltages are similar. The FEM simulation results of MMR MFT are summarized in Table 3.7. The MFT detailed loss is summarized in Table 3.8, and the size and weight breakdown information is summarized in Table 3.9. The loss, size, and weight information for power converter, MFTs, and chiller for MMR at 1000 Hz are summarized in Table

Table 3.7: FEM Results for MMR MFTs

Frequency	Configuration	Parameter	Value
1000	Dd	L_{lk}	3.762 mH
		r_p	2970.4 $\mu\Omega/m$
		r_s	71.761 $\mu\Omega/m$

Table 3.8: MMR MFT Detailed Losses

Copper loss (kW)	Core loss (kW)	Stray loss (kW)	Total losses (kW)
733.5	33.5	50.25	817

Table 3.9: MMR MFT Detailed Size and Weight

Transformer size (m ³)	Copper weight (ton)	Core weight (ton)	Tank weight (ton)	Oil weight (ton)
10.44	3.04	8.6	1.15	4.04

3.2.3. DESIGN OF ISOP MFT

The transformer design for this case follows a method similar to that used in the MMR and MMC-SST. However, the number of transformers

here is eight, according to the structure. The converter produces a square-shaped voltage waveform; thus, the peak voltage must be considered ($U_{\text{pri-ph-peak}} = 3260$ V). The number of turns in the primary and secondary windings are 10 and 2, respectively. The ISOP is a resonant LLC topology, and it is advantageous to utilize the leakage inductance of the MFT as the resonant inductor. Therefore, the primary and secondary windings are positioned on top of each other to achieve the desired high leakage inductance, as shown in Figure 3.9.a. Another difference is that the ISOP MFT is a shell-type transformer because building the windings as similar as possible is essential, mainly regarding the leakage and magnetizing inductances. Using a shell-type transformer guarantees that the leakage inductances remain similar. The shell-type ISOP MFT is shown in Figure 3.9.b.

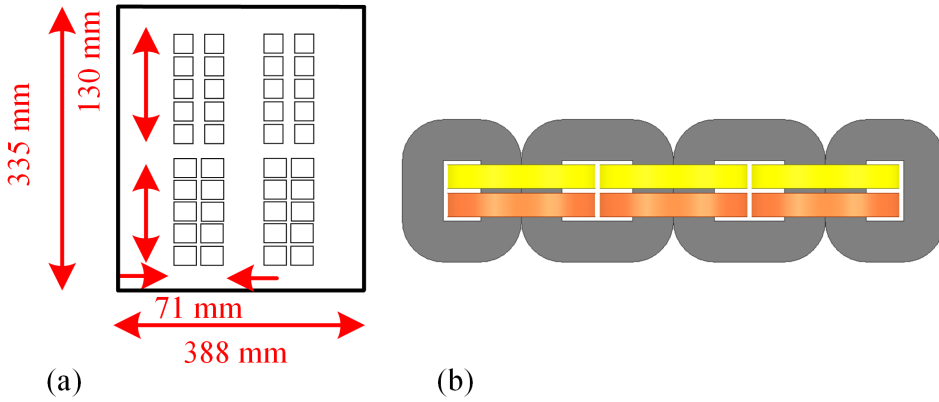


Figure 3.9: Details of the ISOP transformer design: (a) ISOP MFT window layout, (b) ISOP shell-type transformer with primary and secondary windings placed on top of each other to enhance leakage inductance.

The cross-section of the core for the MFT at 1000 Hz is:

$$A_c = \frac{U_{\text{pri-ph-peak}}}{k_c \times k_w \times B_m \times f \times N_p} = \frac{3260}{0.79 \times 4 \times 1.26 \times 1000 \times 10} = 0.082 \text{ m}^2 \quad (3.21)$$

Therefore, the core is selected as $2 \times 170.2 \times 240$ mm. The primary and secondary currents are $I_p = 1334$ A and $I_s = 6465$ A, respectively. Thus, the cross-sectional area of the CTC conductor on the primary side is $\text{CTC}_{\text{pri}} = 21 \times 13 \times 2 = 546 \text{ mm}^2$, and for the secondary side it is $\text{CTC}_{\text{sec}} = 5 \times 21 \times 16 \times 2 = 3360 \text{ mm}^2$ (five parallel paths are used on the secondary side). A 3D view of the ISOP MFT is shown in Figure 3.10.a.

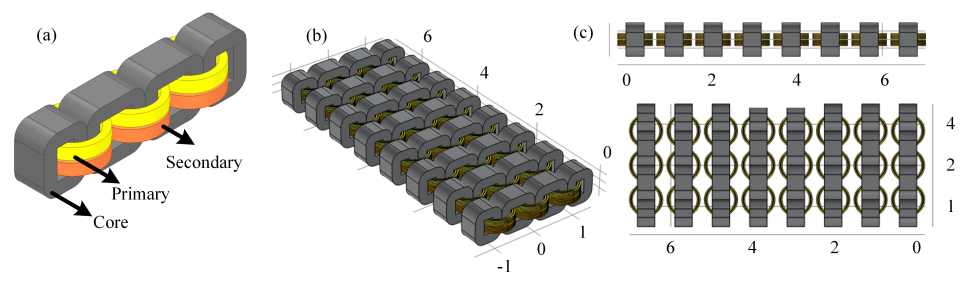


Figure 3.10: The ISOP MFT: (a) 3D view of a single MFT showing the core, primary, and secondary windings, (b) 3D view of eight MFTs arranged inside the tank, and (c) side views of the full assembly from different angles.

INSULATION AND THERMAL CONSIDERATIONS

The insulation and thermal considerations for ISOP MFT are similar to those for MMC-SST and MMR. To decrease the insulation stress, the cores are clamped to the middle point of the input DC Bus and insulated from the tank accordingly. There are eight active parts inside one tank, as shown in Figure 3.10.b and Figure 3.10.c.

OVERALL METRICS OF THE ISOP MFT

The transformers’ detailed loss, total size, and weight breakdown are shown in Table 3.10 and Table 3.11. The FEM simulation results for ISOP MFT are summarized in Table 3.12. The loss, size, and weight information for the power converter, MFTs, and chiller for ISOP at 1000 Hz are summarized in Table 3.6.

Table 3.10: ISOP MFT Detailed Losses

Copper loss (kW)	Core loss (kW)	Stray loss (kW)	Total losses (kW)
644.8	64.6	96.9	806.3

Table 3.11: ISOP MFT Detailed Size and Weight

Transformer size (m ³)	Copper weight (ton)	Core weight (ton)	Tank weight (ton)	Oil weight (ton)
13.78	4.4	15.8	1.74	6.86

Table 3.12: FEM Results for ISOP MFTs

Frequency	Parameter	Value
1000 Hz	L_{lk}	110.4 μH
	r_p	475.49 $\mu\Omega/\text{m}$
	r_s	88.903 $\mu\Omega/\text{m}$

3.2.4. PERFORMANCE METRICS AT DIFFERENT FREQUENCIES

Following a similar design process as shown above, the losses, sizes, and weights at different frequencies of the MMC-SST topologies are shown in Figure 3.11. In the comparative analysis of frequency response, note that for the transformer, only active parts (core and windings) are considered since this simplification is used to ensure manageable computations. For a similar reason, though the converter tower will be different for each analyzed frequency, most of the size changes come from the capacitor size differences between frequencies. Thus, only the capacitors are included to represent the dimensions of the converter parts.

As depicted in Figure 3.11.a, the optimal frequency for the MMC-SST varies with respect to loss, weight, and size. The primary objective of using SST is to reduce the size and weight compared to traditional line frequency transformer-based converter topologies. This reduction is particularly advantageous in high-power applications, as it minimizes the physical footprint of installations where land and space usage can be costly. Therefore, in our optimal design considerations, size and weight are prioritized over losses. Consequently, the most suitable operating frequency for the MMC-SST has been identified as approximately 400 Hz. At this frequency, the design achieves the smallest size and very low weight while maintaining relatively low losses.

As shown in Figure 3.11.b, the optimal fundamental frequency for the MMR topology is identified as approximately 1000 Hz, prioritizing size and weight reduction as key design targets. The frequency sweep range for MMR is different from MMC-SST as shown in Figure 3.11.a, because the analysis aims to identify the optimal frequency corresponding to the turning point. Due to the soft switching characteristics of MMR, it is necessary to consider a broader frequency range. This approach ensures that the optimal frequency is accurately determined. A similar rationale applies to the frequency sweep range analysis for the ISOP topology.

Following a similar analysis process, the losses, sizes, and weights of the ISOP topologies at different operating frequencies are shown in Figure 3.11.c. Due to the zero-voltage-switching (ZVS) turn-on characteristic, turn-on losses are effectively eliminated. However, turn-off losses and circulating-current-related losses are still present and increase with switching frequency, while conduction and magnetic

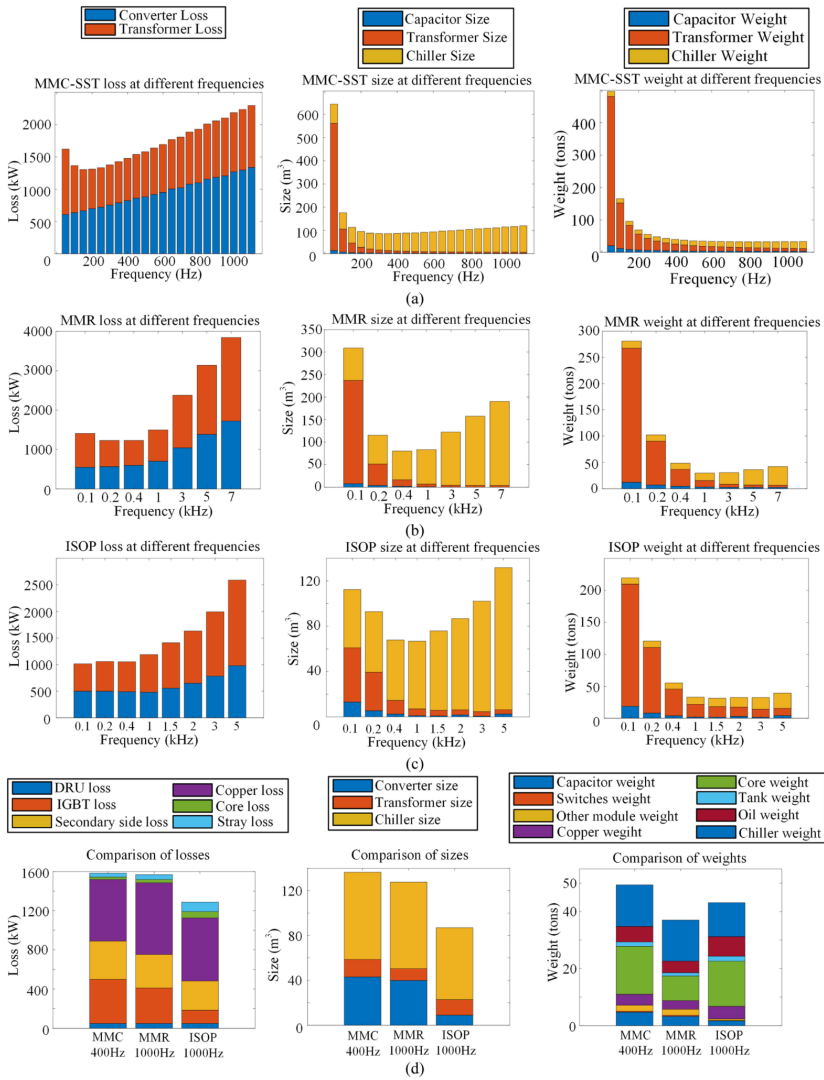


Figure 3.11: (a) MMC-SST, (b) MMR, and (c) ISOP loss, size, and weight at different frequencies. In these figures, for transformer loss, size, and weight estimations, only the active parts are considered for simplicity. For the converter section, the focus is on the changes in size and weight of the capacitor, which are influenced by variations in frequency. Changes in other components are considered minimal and thus not accounted for in this analysis. (d) Topologies comparison results. The optimal operating frequencies for the MMC-SST, MMR, and ISOP are 400 Hz, 1000 Hz, and 1000 Hz, respectively, according to (a), (b), and (c).

losses decrease as the frequency increases. As a result, the total converter power loss exhibits a non-monotonic behavior with frequency. The turning point observed in Figure 3.11.c corresponds to the frequency at which the increase in turn-off and circulating-current losses outweighs the reduction in conduction and magnetic losses, causing the overall loss trend to reverse.

As indicated in Figure 3.11.c, the optimal fundamental frequency for the ISOP topology is set at approximately 1000 Hz, with a primary focus on reducing size and weight.

3.3. CHAPTER CONCLUSIONS AND KEY FINDINGS

This chapter presented the design and performance evaluation of MV-MFTs for SST-based, large-scale green-hydrogen applications. The main conclusions and findings are as follows:

- **Standards framework:** Although dedicated standards for SST-based MV-MFTs are still evolving, the design and validation workflow can be grounded in established practice by combining IEC 60076 (insulation and thermal limits), IEC 60270 (partial-discharge measurement methodology), and IEEE 519 (harmonic performance requirements at the PCC).
- **MMC-SST transformer realization:** To support harmonic mitigation and practical system integration, the MMC-SST implementation uses two active parts in one tank (Yd and Dd). The adopted winding configurations provide a practical path for triplen harmonic currents and enable compliance-oriented system design.
- **Material and winding technology selection:** For the considered medium-frequency range, amorphous core material (Metglas 2605HB1M) was selected to reduce core loss compared with grain-oriented electrical steel. Continuously transposed conductors (CTCs) were adopted to manage high currents and mitigate skin/proximity effects and circulating currents.
- **Loss modeling approach:** A combined FEM-based workflow was used: 3D electromagnetic simulation for leakage/magnetizing inductances and 2D FEM/thermal modeling for load-loss estimation and thermal design inputs. This approach enables consistent estimation of copper, core, and stray losses for the considered geometries and operating frequencies.
- **Cooling concept:** Because the MV-MFTs operate with high loss density in a compact volume, an oil-forced water-forced (OFWF) cooling concept was selected to provide sufficient heat-removal capability and temperature control. For system-level comparison,

the required chiller size and weight were also quantified and shown to be non-negligible in the overall footprint.

- **Frequency trade-off and recommended operating points:**
A frequency sweep revealed a clear loss–size–weight trade-off. Prioritizing footprint reduction as a key SST objective, the recommended fundamental operating frequencies are:
 - MMC-SST: **400 Hz**
 - MMR: **1000 Hz**
 - ISOP: **1000 Hz**

These selections correspond to the turning points where further frequency increase yields diminishing size/weight benefits while losses and/or practical constraints become dominant.

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4

INSULATION COORDINATION

You shall not pass!

Gandalf The Grey

This chapter presents a comprehensive investigation into the insulation coordination of solid-state transformer (SST) systems used in green hydrogen production. It covers both dry-type and oil-immersed transformer insulation strategies, highlighting analytical, numerical, and experimental approaches to address electric field control, partial discharge mitigation, and thermal management. Special attention is given to novel field grading techniques using semiconductive screening and stress cone termination, as well as the use of 3D-printed components in high-voltage insulation systems. The chapter concludes with insulation guidelines for power electronics modules and summarizes key design recommendations to enhance compactness, reliability, and partial discharge immunity in medium-voltage, medium-frequency SST applications.

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R. Mirzadarani, F. Muñoz Munoz, Z. Li, Z. Qin, P. Vaessen, M. Ghaffarian Niasar, "Exploring 3D-Printed Solutions for Medium-Voltage Insulation Applications," in International Symposium on High Voltage Engineering (ISH), Karuizawa, Japan, 2025 [2]

Z. Li **R. Mirzadarani**, M. Ghaffarian Niasar, M. Itraj, L. van Lieshout, P. Bauer, Z. Qin, "Medium-Voltage Solid-State Transformer Design for Large-Scale H₂ Electrolyzers," in IEEE Open Journal of Power Electronics, vol. 5, pp. 936-955, 2024 [3].

This chapter focuses on insulation coordination in medium-voltage, medium-frequency transformers used in solid-state transformer systems for green hydrogen production. It begins with an investigation into dry-type transformer insulation, highlighting electric field behavior, partial discharge risks, and the role of structural gaps. Electric field analysis is supported by analytical derivations and FEM simulations to better understand local field enhancements.

The chapter then explores semiconductive screening and field grading techniques, including the use of stress cones, to improve insulation performance while maintaining compactness. Experimental validation is provided through partial discharge measurements. The use of 3D-printed components as insulating elements is examined, comparing their performance to bulk (non-printed) materials and identifying relevant design parameters. In addition, insulation requirements for power electronics modules within the solid-state transformer system are addressed. Finally, insulation considerations in oil-immersed transformers and power electronics modules are addressed, offering design guidance based on field simulations and standardized test requirements.

In this chapter, the insulation design is primarily investigated for a dry-type transformer configuration. Although oil-immersed transformers are commonly used in full-scale industrial systems, the experimental demonstrator developed in this work was intentionally designed as a dry-type unit. This choice was motivated by practical considerations, including easier maintenance, cleaner operation, and a reduced risk of fire and oil leakage in a laboratory environment. In addition, dry-type construction simplifies manufacturing, modification, and testing of insulation concepts, which is essential for experimental validation. For these reasons, the dry-type transformer serves as a suitable platform for exploring insulation coordination, partial discharge behavior, and field grading techniques, while the insights obtained remain relevant for full-scale oil-immersed implementations.

The choice between dry-type and oil-immersed transformer concepts is also closely related to the distribution of losses within the transformer. In conventional 50 Hz power transformers, the total losses are typically divided into no-load (core) losses and load losses. Core losses generally account for approximately 10–20% of the total losses, while load losses dominate at rated current. Load losses mainly consist of winding copper losses, with additional stray losses arising from leakage flux inducing eddy currents in windings, structural parts, and the metallic tank.

In medium- and high-frequency transformers, such as those used in SST applications, the relative importance of stray losses becomes more pronounced. The increase in operating frequency amplifies eddy-current losses in conductive structural components, particularly metallic tanks and enclosures, if not carefully mitigated. As a result,

the presence of a conductive tank can lead to disproportionately high stray losses compared to low-frequency designs. Dry-type transformers inherently avoid tank-related eddy current losses and therefore offer a practical platform for medium-frequency operation, especially at the demonstrator level. For the full-scale system, an oil-immersed transformer is inevitable due to thermal management requirements, whereas a dry-type configuration is selected for the demonstrator MFT, where insulation behavior and field control concepts can be investigated under controlled laboratory conditions.

4.1. DRY-TYPE TRANSFORMERS

Dry-type transformers offer several advantages over oil-immersed transformers, such as reduced fire hazards, minimal maintenance requirements due to the absence of oil, lower operational costs, and minimized environmental risks associated with oil leakage. Their robust design also provides high mechanical strength against short circuit forces. These features make dry-type transformers particularly suitable for densely populated or environmentally sensitive areas. However, the design of the insulation system of a dry-type transformer and the prevention of partial discharges (PDs) are challenging, particularly in high-voltage, medium-frequency applications [4, 5].

One of the key advantages of dry-type transformers in medium-frequency Solid-State Transformer (SST) applications lies in their structural design, which help reduce frequency-related losses. In oil-type transformers, the presence of a metallic tank becomes a critical factor affecting energy efficiency at higher frequencies. Increasing frequency raises stray losses, primarily due to eddy currents in the tank. These eddy currents, induced in the conductive material of the tank, cause extra heat, reducing overall efficiency and potentially leading to overheating issues. In contrast, dry-type transformers inherently avoid these frequency-induced stray losses by lacking an oil tank, a key advantage in medium-frequency SST applications. Without a conductive tank, the formation of eddy currents and stray losses are significantly reduced, resulting in more efficient operation in higher-frequency environments.

The insulation in a dry-type transformer consists of solid materials such as epoxy resins and several air gaps between primary and secondary, windings and core, and between phases. The primary purpose of these gaps is to provide electrical insulation and to prevent dielectric breakdown between the HV and LV windings and between the windings and the core. The gap size depends on the voltage level and the dielectric strength of the surrounding medium (air). Moreover, adequate gaps facilitate cooling and heat dissipation. Since dry-type transformers rely on natural or forced air cooling, appropriate spacing is essential to ensure a proper heat transfer. The window gaps must maintain

mechanical stability and withstand vibrations or movements during operation and transportation. Table 4.1 summarizes the dimensional information of several dry-type power 50 Hz transformers and shows thermal and mechanical factors influencing the gap between the low-voltage (LV) winding and the core. Nevertheless, thermal and insulation considerations are critical for the gap between high-voltage (HV) and low-voltage (LV) windings. The transformers summarized in Table 4.1 represent practical dry-type, line-frequency units widely used in industrial and renewable energy applications. These data points are included to provide realistic reference values for insulation dimensions, rather than experimental or downscaled prototypes.

4

Table 4.1: The Geometrical Information of Several Dry-Type Power 50 Hz Transformers

P^1 (kVA)	V_{HV}^2 (kV)	V_{LV}^2 (V)	r_{core}^3 (mm)	$r_{1,ins,HV}^4$ (mm)	$r_{1,ins,LV}^4$ (mm)	$r_{1,ins,out}^5$ (mm)	LV-Core Gap (mm)	HV-LV Gap (mm)
160	6	400	97.5	101	128.5	165	3.5	36.5
250	6	400	97.5	101	128.5	165	3.5	36.5
315	13.2	220	104.7	110	135.4	170	5.4	34.6
400	13.2	220	104.7	110	135.4	170	5.4	34.6
500	6	400	112.5	117	142.5	180	3.5	35.5
630	6	400	122.7	127	154.5	190	3.5	35.5
800	6.6	400	124.5	128	158.3	195	3.5	36.7
1000	6.6	400	124.5	128	158.3	195	3.5	36.7
1250	10	400	136.5	142	170.5	205	5.3	41.6
1600	10.8	400	145.8	149.2	182.7	225	5.3	42.9
2000	10.8	400	145.8	149.2	182.7	225	5.3	42.9
2500	6	400	162.5	166	217.1	260	3.5	42.9
3150	6	400	162.5	166	217.1	260	3.5	42.9

¹ P : Rating power.

² V_{HV} and V_{LV} : Rated voltages of high and low voltage windings, respectively.

³ r_{core} : Radius of the core.

⁴ $r_{1,ins,HV}$ and $r_{1,ins,LV}$: Inner radius of insulation for HV and LV windings, respectively.

⁵ $r_{1,ins,out}$: Outer radius of LV winding.

4.1.1. THE ELECTRIC FIELD IN A DRY-TYPE TRANSFORMER

The windings of a dry-type transformer can be assumed as cylinders to avoid complexity. This assumption is common in preliminary insulation analysis and enables closed-form expressions that offer intuitive insights. The electric field between the cylinders shown in Figure 4.1.a can be derived as follows:

$$E = \frac{V_1 - V_2}{r \ln\left(\frac{r_2}{r_1}\right)} \quad (4.1)$$

where V_1 and V_2 are the voltages of the inner and outer cylinders, respectively, and r_1 and r_2 are their corresponding radii. The maximum electric field, $E_{\max}$, can be calculated as follows:

$$E_{\max} = \frac{V_1 - V_2}{r_1 \ln\left(\frac{r_2}{r_1}\right)} \quad (4.2)$$

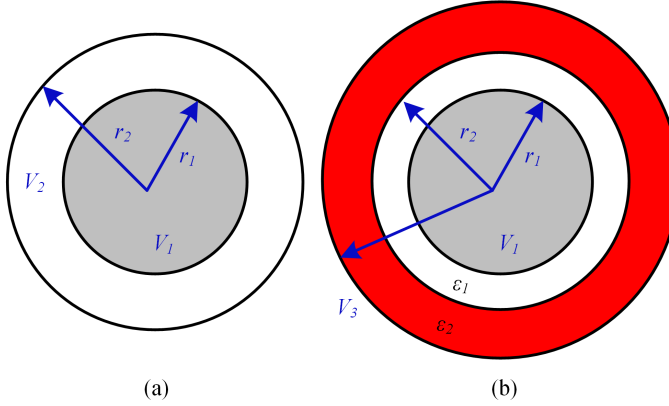


Figure 4.1: (a) Two concentric cylinders representing the winding around the core, (b) gaps filled with different materials.

However, the gap between windings is usually filled by different materials, such as epoxy resin and air, as shown in Figure 4.1.b. Thus 4.1 is rewritten as follows:

$$E_i = \frac{V_1 - V_3}{r\epsilon_i \left[\frac{1}{\epsilon_1} \ln\left(\frac{r_2}{r_1}\right) + \frac{1}{\epsilon_2} \ln\left(\frac{r_3}{r_2}\right) \right]}, \quad i = 1, 2 \quad (4.3)$$

where ϵ is the relative permittivity. In a typical transformer, the low-voltage (LV) winding is placed near the core, and the high-voltage (HV) winding is placed over the LV winding. Each winding is covered with epoxy resin from both the inner and outer sides. Therefore, the problem is to calculate the electric field in the gaps and inside the epoxy resins, as shown in Figure 4.2. Therefore, the field calculation problem for the LV-Core gap is as follows:

$$E_i = \frac{0 - V_{LV}}{r\epsilon_i \left[\frac{1}{\epsilon_0} \ln \frac{r_{LV_epoxy_in}}{r_{core}} + \frac{1}{\epsilon_{epoxy}} \ln \frac{r_{LV_in}}{r_{LV_epoxy_in}} \right]} \quad (4.4)$$

where V_{LV} is the voltage of the LV winding, $r_{LV_epoxy_in}$ is the inner radius of the LV epoxy, r_{core} is the radius of the core, r_{LV_in} is the inner radius of the LV winding, ϵ_0 is the permittivity of air, and ϵ_{epoxy} is the

permittivity of epoxy resin. Note that the core is grounded; thus, its voltage is zero. Similarly, the electric field in the HV-LV gap can be calculated as follows:

$$E_i = \frac{V_{LV} - V_{HV}}{r\epsilon_i \left[\frac{1}{\epsilon_{epoxy}} \ln \frac{r_{LVepoxy_out}}{r_{LV_out}} + \frac{1}{\epsilon_0} \ln \frac{r_{HVepoxy_in}}{r_{LVepoxy_out}} + \frac{1}{\epsilon_{epoxy}} \ln \frac{r_{HV_in}}{r_{HVepoxy_in}} \right]} \quad (4.5)$$

where V_{HV} is the voltage of the high-voltage (HV) winding, r_{LV_out} is the outer radius of the low-voltage (LV) winding, $r_{HVepoxy_in}$ is the inner radius of the HV epoxy, and r_{HV_in} is the inner radius of the HV winding. Therefore, the LV-Core and HV-LV gaps can be calculated from Equations (4.4) and (4.5), respectively.

The derived field expression provides an estimate of the maximum radial field stress between conductors is air, which is directly related to the onset of partial discharges in solid insulation. While the above equations provide insight into the radial electric field in idealized concentric geometries, practical transformer windings exhibit more complex field behavior. Local electric field enhancements often occur due to geometric discontinuities such as bolt protrusions, casting seams, and surface roughness. As illustrated in Figure 4.3, even simple features like a semi-cylindrical or semi-spherical bump can increase the local field intensity by a factor of 2 to 3. Micro-protrusions can produce even greater amplification, with field enhancement factors (β) exceeding 40 in extreme cases. These localized effects can push the electric field beyond 1.5 MV/mm, even when the average field remains below 40 kV/mm [6].

Sharp edges, common features in transformers particularly within their cores, can enhance the electric field considerably (Figure 4.4). The electric field intensity around these sharp edges can be determined using the method of conformal transformation. This approach provides a mathematical framework for analyzing electric field dynamics in areas with sharp edges [7]. However, with the advancements in numerical methods and Finite Element Method (FEM) simulations, it has become more advantageous to employ FEM for calculating non-uniform electric fields. In this paper, FEM simulations are utilized wherever necessary to enhance the precision and efficacy of electric field calculations.

Thermal considerations determine the LV-core gap when the LV voltage level is sufficiently low. In contrast, the HV-LV gap must be designed to satisfy both insulation coordination and thermal requirements. Increasing the thickness of the epoxy insulation is not a favorable solution, as it reduces the available air gap and degrades heat transfer. From an electrical perspective, the electric field in the air gap is higher than in solid insulation due to the lower relative permittivity of air.

The design of the HV-LV gap is primarily governed by partial discharge inception and surface flashover criteria under the applied AC voltage.

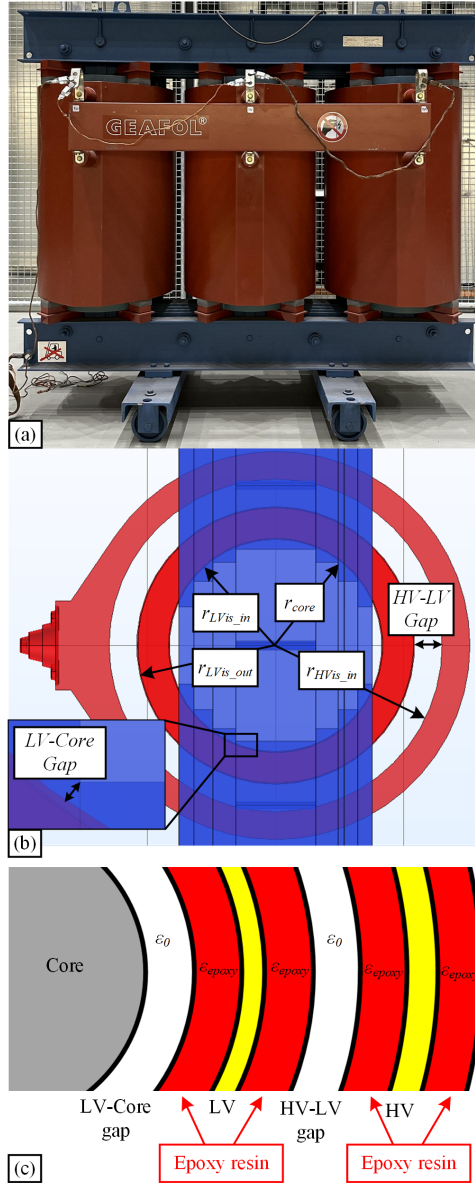


Figure 4.2: Structural overview and field configuration of a dry-type medium-voltage transformer. (a) Photograph of a three-phase cast-resin transformer, (b) Cross-sectional schematic of winding geometry and gap definitions, (c) Radial electric field configuration and material composition, including air gaps and epoxy-encapsulated winding sections.

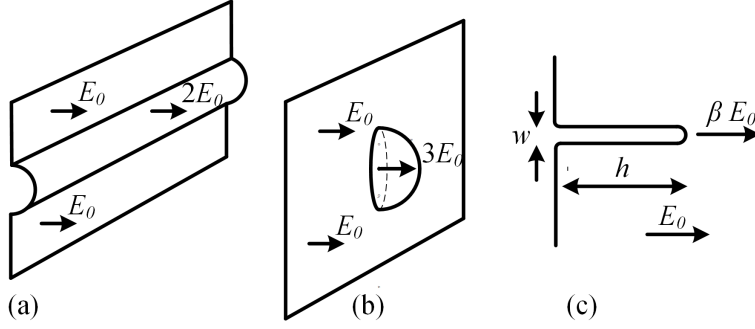


Figure 4.3: Electric field enhancement due to geometrical protrusions and surface irregularities: (a) A semi-cylindrical seam or welding joint doubles the local electric field to approximately $2E_0$, (b) A semi-spherical feature, such as a bolt head or droplet, can triple the field to $3E_0$, (c) Micro-scale field enhancement at sharp tips or protrusions, where the local field is amplified by a factor of β , defined by geometry as $\beta \approx 2 + h/(w/2)$. These effects demonstrate how small-scale features can significantly increase the risk of partial discharge and surface flashover, even under moderate background field conditions.

In the present work, insulation coordination is evaluated based on steady-state and power-frequency test voltages relevant to solid-state transformer operation. Lightning impulse stresses are not considered for the demonstrator design, as the transformer is not directly connected to an exposed transmission line and is operated within a power-electronics-dominated system. Consequently, the gap dimensions are selected to ensure PD-free operation and sufficient margin against surface flashover under the applied voltage stress. Let material 1 be the epoxy resin that is used to build the winding insulation, and material 2 is air. The homogeneous electric field is defined as follows:

$$E = \frac{V}{d} \quad (4.6)$$

where V and d are the applied voltage and the distance between electrodes, respectively. An adequate minimum air gap must be maintained to ensure the electric field remains below a certain threshold. Let us define E_{PD} as the maximum allowable electric field strength in air to avoid partial discharge (PD). Then, the minimum required air gap can be calculated as:

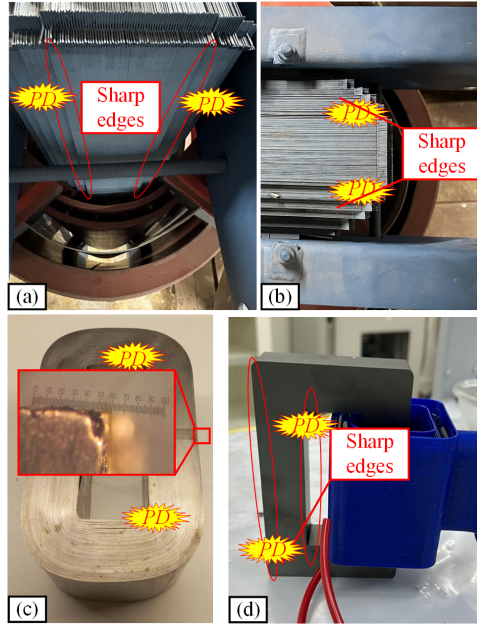


Figure 4.4: Sharp edges in different magnetic cores: (a) and (b) electrical steel 50 Hz core, (c) amorphous core, and (d) ferrite core.

$$d_{\text{air}} = \frac{V - \left(\frac{E_{\text{PD}} \cdot t_{\text{is}}}{\epsilon_r} \right)}{E_{\text{PD}}} \quad (4.7)$$

where V is the applied voltage, ϵ_r is the relative permittivity of the insulation (epoxy resin), E_{PD} is the PD inception voltage, and t_{is} is the thickness of the insulation.

While this analytical approach provides physical insight, the full electric field behavior is later validated through FEM simulations to account for 3D edge effects and mixed dielectric interfaces.

4.2. FIELD GRADING USING SEMI-CONDUCTIVE MATERIALS

Utilizing semiconductor screens to encapsulate conductors and the insulation system is a well-known practice in the field of high-voltage (HV) and extremely high-voltage (EHV) cable technologies [8]. This approach typically involves triple extrusion, where insulation and semiconductive layers are extruded around the conductor in a single operation. After extrusion, the conductor is covered with a

semiconductive layer, an insulation layer, and another semiconductive layer around the insulation. This configuration ensures a robustly bonded interface between the conductor and the dielectric, and addressing potential issues that might arise during thermal cycling, such as separating the insulation layer from the conductor or the metallic screen and forming gaps. These gaps can lead to electric field enhancement, partial discharge initiation, and eventually, cable failure. Additionally, this method prevents the intensification of electric field strength in the insulation region next to the conductor and metallic screen, which can occur due to the stranded structure of the conductor and the metallic screen and the presence of sharp edges on these metallic surfaces [9].

The semiconductive material, which shares a base composition similar to the insulation, plays a crucial role in this context. During the thermal cycling process, the compatibility of material properties between the semiconductive layer and the insulation minimizes the likelihood of gap formation between these layers. Any gap between the conductor and the semiconductive layer poses no danger, as both the semiconductive layer and the conductor maintain the same voltage, thus ensuring a close to zero electric field in these gaps, eliminating the possibility of partial discharge initiation.

This section investigates the use of semiconductive screening for the windings of a medium-voltage dry-type transformer. Unlike the approach in [10, 11], which involve separate XLPE cables for winding construction, the proposed method encapsulates the entire winding, including the conductors and cast resin insulation, within semiconductor screens. While encapsulating windings to reduce the air gaps can help reduce the size of any dry-type transformer, it is more significant when used to decrease the gap size in medium-frequency transformers (MFTs). MFTs, used in solid-state transformer (SST) applications, are designed to minimize the size of magnetic components. However, their design is constrained by two key factors: insulation requirements and thermal management [3]. Dry-type MFTs offer a more straightforward and less complex alternative compared to oil-insulated MFTs. Grid-connected MFTs must withstand voltage tests, such as acceptance criteria of IEC 60076-3 [12] and IEC 60076-11 [13], regarding their types. Considering that the larger gaps cancel the main advantage of a compact SST, this paper presents a novel technique, semiconductive screening, which helps reduce the dry-type MFTs' overall size. Encapsulating the windings of a dry-type transformer also offers additional benefits, such as enhancing safety and mitigating electromagnetic interference (EMI) issues.

To tackle stray losses in medium-frequency SST applications, semiconductive screening offers a practical and effective solution, particularly in dry-type transformer designs. In oil-immersed transformers, a metallic tank naturally contains the electric field and the conductors must

brought out of the tank via proper bushings. However, in dry-type transformers, the absence of such an enclosure allows the electric field to extend beyond the epoxy insulation, increasing the risk of surface discharge.

The proposed semiconductive screen serves a similar field-containment function, but with the added advantage of significantly reduced eddy current losses. Unlike metallic tanks, the semiconductive screen have limited conductivity, which minimizes the formation of eddy currents even at higher switching frequencies. Furthermore, the screen's minimal thickness further suppresses the eddy current generation. These combined properties, low conductivity and small thickness, ensure that any induced currents are weak and contribute negligibly to overall stray losses. As a result, semiconductive screening provides a compact and efficient means of controlling electric field boundaries while mitigating frequency-related losses in MFTs [14, 15].

Field grading and insulation design in SST applications are the subject of several studies [16–22]. Reference [16] details the development and testing of two MFT prototypes for use in a three-phase Dual-Active Bridge (DAB3) DC-DC converter system. These transformers are designed to connect offshore wind farms to medium-voltage grids. The first prototype (MFT 1) required additional inductors to achieve the needed series inductance. The second prototype (MFT 2) integrated this inductance within the same core using a stacked winding configuration, eliminating the need for extra inductors. MFT 2 also featured high-voltage secondary windings insulated with semiconductive and cast resin materials, ensuring compatibility with ± 25 kV medium-voltage applications.

Reference [17] focuses on the electrical shielding of medium-voltage, medium-frequency transformers subjected to high dv/dt PWM voltages. It investigates the distribution of electric fields in the insulation, at the surface, and in the air for a specific MV/MF transformer. The study emphasizes the design of an effective shield to reduce the electric field without increasing losses. The shielding strategies employ semiconductive and resistive materials to manage the insulation challenges posed by high voltage and frequency applications, ensuring better performance and reliability of the transformers. Reference [18] addresses the insulation design for high-frequency transformers in medium-voltage, air-cooled, silicon carbide (SiC)-enabled solid-state transformers. It proposes an insulation system integrating polymer and inorganic materials to balance high dielectric strength and thermal conductivity. This system, designed to be partial-discharge-free, demonstrates an insulation level of ± 95 kV and an effective thermal conductivity approximately 2-3 times higher than similar polymer materials, making it suitable for high-voltage and high-frequency applications. In [19], the authors present a study on the internal

electric field shielding structure for high-voltage windings in medium-voltage, high-frequency transformers. They introduce a shielding layer incorporating low resistivity semiconductive (LSC) material and copper foil to optimize electric field distribution and enhance insulation reliability. This approach aims to reduce the peak internal electric field, thereby lowering the risk of partial discharge, and is validated with a prototype transformer that successfully passes insulation tests.

The subject of [20] is an improved high-voltage winding internal shielding structure for medium-voltage high-frequency transformers, focusing on reducing eddy losses. It proposes a new structure using semiconductive materials and copper foil, designed to minimize eddy current losses while maintaining effective electric field shielding. Reference [21] focuses on the design of a high-frequency transformer with medium-voltage insulation for a resonant converter in solid-state transformers. It presents an innovative approach to insulation, using semiconductive materials to enhance electric field distribution, reduce peak internal electric fields, and lower partial discharge risks. Reference [22] presents a planar structure high-frequency transformer design for medium-voltage applications, focusing on insulation and shielding techniques. It utilizes modular PCB boards for high-voltage side windings, where a coated semiconductive shielding layer confines the high electric field within the PCB insulation.

While previous studies have explored various field grading techniques and shielding strategies in medium-frequency transformer designs, many have focused either on planar insulation layouts or cable-based windings. However, comprehensive methods that integrate semiconductive screening directly within the epoxy-based winding encapsulation for dry-type medium-voltage transformers and proper strategy for designing the termination are lacking. The present chapter builds upon established principles from HV/EHV cable technology and adapts them to transformer windings into medium-frequency solid-state applications, with the aim of improving insulation compactness and PD performance without compromising thermal dissipation or mechanical integrity.

The main contributions of the present chapter are as follows:

- A novel implementation of semiconductive screening for encapsulated windings in medium-voltage, medium-frequency dry-type transformers;
- Analytical and FEM-based modeling of electric field distribution considering mixed air-epoxy insulation media;
- A detailed design and fabrication method for stress cone-based field grading, compatible with the proposed screening technique;
- Experimental verification via partial discharge measurements,

demonstrating a significant PD reduction from 46 pC to 2.2 pC, validating the effectiveness of the proposed approach.

Figure 4.5 shows the impact of applying the proposed semiconductive paint on the dimensions and overall volume of a 50 kW, 1 kHz MFT demonstrator. Applying the proposed method results in smaller transformer and lesser environmental footprints and cost.

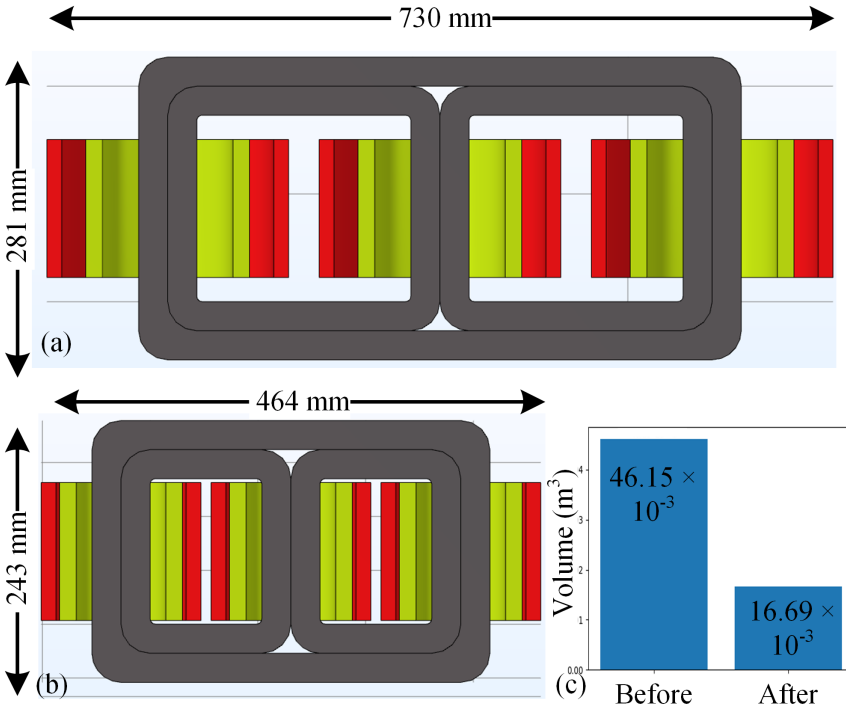


Figure 4.5: The impact of the proposed method: (a) dimensions of a 50 kW, 1 kHz MFT before applying the proposed method; (b) the same transformer after implementing the proposed method, showing a reduction in dimensions due to smaller required insulation clearances; (c) comparison of the total volume before and after applying the method.

4.2.1. LIMITATIONS OF LINE-FREQUENCY INSULATION TESTING FOR MEDIUM-FREQUENCY SSTs

The insulation design and validation presented in this work are primarily based on established standards such as IEC 60076-3 and IEC 60270, which are defined for line-frequency (50/60 Hz) sinusoidal voltage

excitation. However, the medium-frequency transformer (MFT) in a solid-state transformer (SST) is subjected to non-sinusoidal voltage waveforms with significantly higher fundamental frequencies (e.g., 400 Hz to 1 kHz) and fast voltage transitions.

This discrepancy introduces important limitations. Partial discharge (PD) behavior is known to depend not only on the peak electric field but also on the rate of voltage change (dv/dt), waveform shape, and repetition frequency. Compared to sinusoidal excitation, quasi-square or resonant waveforms may lead to different electric field distributions, charge accumulation dynamics, and PD inception conditions. In particular, faster voltage transitions can locally enhance electric stress, while higher repetition rates may increase cumulative insulation degradation.

As a result, PD inception voltages and insulation margins derived from 50/60 Hz testing cannot be assumed to be fully representative of actual SST operating conditions. In this work, line-frequency PD and AC withstand tests are therefore used as a baseline for insulation validation and manufacturing quality assessment, rather than as a complete characterization of insulation performance under converter-driven operation.

To mitigate this limitation, conservative design margins, field grading techniques (e.g., stress cones and semiconductive coatings), and FEM-based electric field analysis are employed. Nevertheless, further research is required to establish standardized testing procedures and design criteria for medium-frequency, non-sinusoidal excitation.

4.2.2. PARTIAL DISCHARGES AND SURFACE FLASHOVER IN TRANSFORMERS

Understanding and controlling partial discharges and surface flashover is essential, as these phenomena determine the electrical clearance requirements in transformer design. Reducing their occurrence enables more compact insulation design. Partial discharges (PDs) and surface flashover events are often the limiting factors in determining insulation clearances in dry-type transformers. The required distances between conductors, or between conductors and grounded components, are primarily dictated by the electric field intensity that initiates these effects. Therefore, mitigation of PDs and surface discharges enables more compact insulation design, which directly contributes to transformer miniaturization.

PARTIAL DISCHARGES

Partial Discharge (PD) in transformers is a critical phenomenon that refers to localized electrical discharges within the insulation system of the high-voltage electrical equipment [23–28]. These discharges can

happen when the insulation is weak or damaged, leading to gradual deterioration. The importance of mitigating PDs lies in their potential to cause insulation failure, resulting in costly downtime, hazardous working conditions, and reduced equipment lifespan. Early detection and management of PDs are crucial for ensuring the reliability and safety of power systems [29]. Analyzing the PD requires knowledge about PD patterns that can occur. Some patterns of PD are summarized in Figure 4.6 [23, 30, 31].

The primary cause of partial discharge (PD) in transformers is the presence of sharp edges, particularly around the core. These edges act as stress concentrators, significantly enhancing local electric fields and initiating PD. This phenomenon is most noticeable when geometric irregularities, like those of the core, disrupt the uniformity of the electric field distribution. In 50 Hz transformers, using electrical steel sheets often results in sharp edges, which can be a significant source of PD difficulties. On the other hand, in medium-frequency or high-frequency transformers, the ferrite cores and cores made from amorphous strips tend to have sharp edges, similarly giving rise to challenges associated with PDs (Figure 4.4).

An additional challenge in medium-frequency or high-frequency transformers arises from the necessity of using litz wires to mitigate copper loss at higher frequencies. However, due to the complex shape of the litz wire, the impregnating medium, such as the epoxy resin, may not stick very well to each other, and even during thermal cycles, different thermal expansion coefficients between the two might cause separation and formation of cavities which causes PDs and long term aging and failure. As a solution, [19] proposes using internal electric field shielding, which has effectively mitigated the possibility of the formation of PD.

Besides the factors mentioned above, other elements also influence the presence of PD in transformers. Among these, the presence of air bubbles in the epoxy resin during the casting process is particularly critical. A meticulous casting procedure is essential to guarantee the absence of air bubbles within the epoxy, involving precise step-by-step pre-heating, pre-degassing, mixing, casting, degassing, curing, and cooling, each of which must be conducted with precision to minimize the risk of PD.

SURFACE FLASHOVER

Surface flashover occurs along the improperly designed termination. Applying a coating to the conductors and increasing the distance between them does not necessarily mitigate all risks associated with the insulation system of a dry-type transformer. This typically occurs when a significant tangential (parallel) electric field component is present along the insulation surface, which drives charge transport and discharge

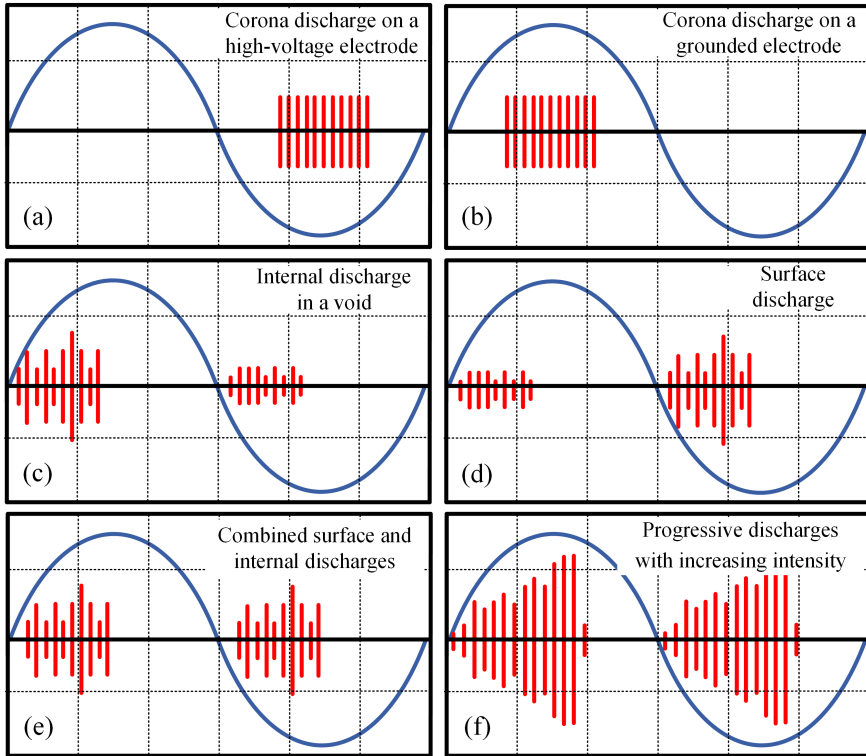


Figure 4.6: Oscilloscope response patterns of different types of partial discharge (PD). The blue waveform shows the phase of the applied AC voltage, while the red pulses indicate detected PD events. These phase-resolved patterns are commonly used to identify and classify the source of discharges: a) Corona discharge on a high-voltage electrode, b) Corona discharge on a grounded electrode, c) Internal discharge in a void, d) Surface discharge, e) Combined surface and internal discharges, f) Progressive discharges with increasing intensity. These patterns are used to interpret PD test results and assess insulation behavior under different conditions.

propagation.

Surface discharges occur when the applied voltage surpasses a certain threshold, leading to an intensified electric field caused by these initial discharges. This intensified field becomes adequate to initiate the propagation of discharge. A series of such discharges eventually triggers a surface flashover along the entire length of a winding, occurring at lower voltage levels compared to other types of electrical breakdowns [32, 33]. Therefore, proper termination is necessary to avoid flashover. Figure 4.7 illustrates a surface flashover incident, where, due to improper termination, a flashover occurred over a considerable distance from the high voltage point to the grounded core.

Understanding the conditions under which PDs and flashovers initiate helps define the motivation for implementing field control techniques, such as semiconductive screening and stress cone shaping, which are detailed in the subsequent sections.

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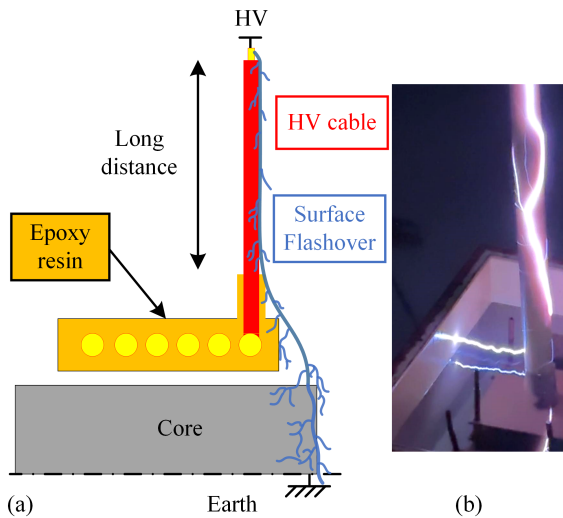


Figure 4.7: (a) Surface flashover across a long creepage distance between the HV cable and grounded core in a dry-type transformer, (b) photograph of a surface flashover captured during laboratory testing.

4.2.3. FIELD GRADING METHODS

Although electrode-based field grading slightly increases the volume at the winding terminations, it enables a significant overall reduction in insulation distances by mitigating electric field stress, which in turn supports transformer miniaturization.

Several methods are available to enhance electric field distribution, including electrode grading, epsilon grading, condenser grading, and resistive grading, as depicted in Figure 4.8 [34, 35]. Among these, electrode grading is often preferred due to its numerous advantages. This method enables precise control over electric field distribution by shaping the electrodes to achieve a more uniform field. This uniformity reduces peak electric field strength and lessens the risk of dielectric breakdown, particularly at points of natural field intensification, such as sharp corners or edges. Electrode grading's efficiency also lies in its minimal material requirements, contrasting with epsilon grading, which depends on dielectrics with specific permittivity, or resistive and condenser grading that requires additional components to act as resistance or capacitance. Simply reshaping existing electrode designs is material-efficient and versatile, applicable across various components and systems, from high-voltage transmission lines to transformers and capacitors. This versatility often translates into cost-effectiveness, avoiding the expense of adding extra components. In addition, the manufacturing costs are generally lower, and the likelihood of defects is reduced since the process involves modifying existing designs rather than introducing new elements. Graded electrodes, designed appropriately, are less likely to degrade compared to resistive or capacitive elements, leading to extended service life and reduced maintenance. Importantly, electrode grading focuses on design rather than incorporating dissipative components, contributing to more energy-efficient operations.

Bushings are crucial components that allow high-voltage conductors to pass safely through grounded barriers, such as transformer tanks or walls, while providing the required electrical insulation. However, due to design complexities and material properties, they face notable limitations in medium-frequency applications. A critical issue is capacitance and impedance matching. Transformer bushings inherently introduce parasitic capacitance, which at higher frequencies leads to significant reactive impedance, thereby disturbing the impedance matching of the connected circuits. This effect becomes more pronounced as the operating frequency increases, resulting in higher losses and waveform distortion, and thus necessitating careful bushing and termination design for effective medium-frequency performance.

Material limitations further limit the transformer bushings for higher frequencies. Standard materials like porcelain or resin-impregnated paper may not effectively withstand the stress of higher frequencies. Their electrical properties can change at such frequencies, leading to decreased reliability and a shorter lifespan. Moreover, bushings are subject to Frequency-Dependent Loss Factors, where insulation materials' loss factors increase at higher frequencies, causing inefficiencies and overheating risks. This necessitates choosing materials that can endure

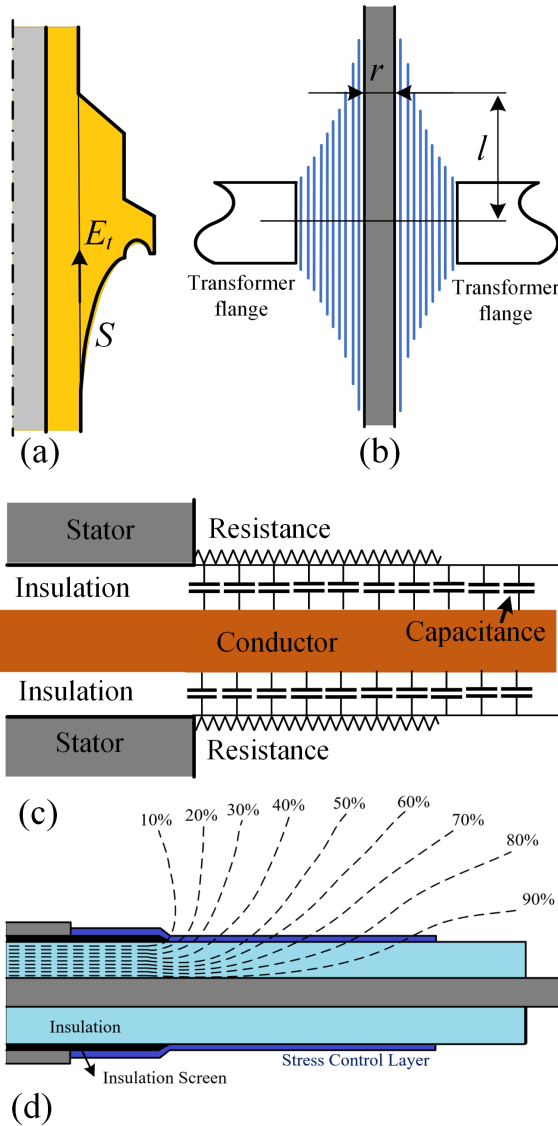


Figure 4.8: Field grading methods in practice: (a) Electrode grading using a stress cone, (b) Condenser bushing employing capacitive grading with floating foils, (c) Resistive grading in the stator of a high-voltage machine under AC stress, modeled with equivalent resistance and capacitance, (d) Epsilon grading using a stress control layer with varying permittivity to achieve a uniform electric field distribution along a cable termination.

the demands of medium-frequency applications.

Additionally, manufacturing tolerances become important at higher frequencies. The production process must achieve acceptable tolerances, as even minor imperfections significantly impact performance at these frequencies. This precision requirement heightens both the complexity and cost of producing bushings for medium-frequency applications, leading to a substantial challenge in design and manufacturing. This chapter employs electrode field grading as the preferred method for addressing the electric field distribution challenges in transformer bushings, particularly in medium-frequency environments, due to its efficiency, material economy, and adaptability across various applications.

The practical implementations of high-voltage insulation and termination systems are shown in Figure 4.9. The dry-type inner cone connector depicted in Figure 4.9.a is commonly used for medium-voltage cable terminations, offering a compact and shielded interface. In contrast, Figure 4.9.b presents a composite outdoor bushing, which integrates capacitive grading and is often equipped with a tap for partial discharge monitoring, suitable for outdoor and high-contamination environments.

4.2.4. ELECTRODE FIELD GRADING USING STRESS CONE

The primary purpose of field grading is to prevent excessive electric field strengths that can lead to breakdowns or failures in insulation materials. In high-voltage systems, sharp edges or points can create areas of high electric field strength, which can cause partial discharges or even complete insulation breakdown [23, 34].

Field grading is often achieved using materials with specific electrical properties or by shaping the conductors and insulating materials to distribute the electric field more evenly. For example, grading rings, made of conductive or semiconductive materials, are often placed around high-voltage equipment to smooth out the electric field. Proper design for field grading involves understanding the electric field distribution under various conditions, selecting appropriate materials, and shaping components to minimize areas of high electric field concentration. The primary challenge encountered in previous attempts to employ semiconductive screening for the encapsulation of MFT windings lies in the effective management of the electric field. When a semiconductive shield is positioned around a high-voltage conductor coated with epoxy resin, it neutralizes the electric field in the surrounding air. Nevertheless, a significant concern arises at the shield's end, where the electric field may intensify to hazardous levels, potentially leading to partial discharge (PD) or, in severe cases, electrical breakdown. However, as shown in Figure 4.10, proper termination can mitigate this challenge significantly.

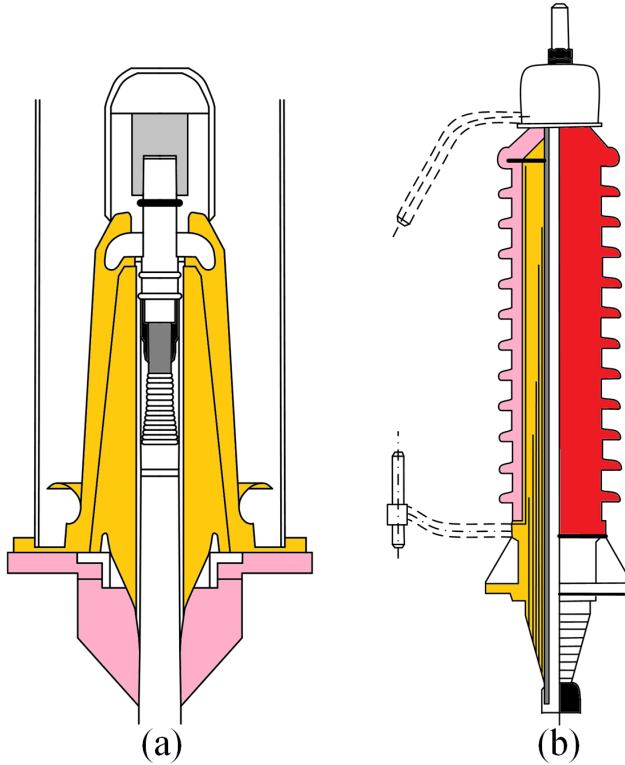


Figure 4.9: Cross-sectional views of high-voltage insulation applications: (a) A dry-type inner cone connector used in medium-voltage cable terminations, showing the compact arrangement of conductive and insulating components, (b) A composite outdoor bushing with integrated capacitive tap for partial discharge monitoring, illustrating the layered internal structure and weather-resistant housing. These designs highlight the importance of geometrical grading and proper interface control in high-voltage applications [23].

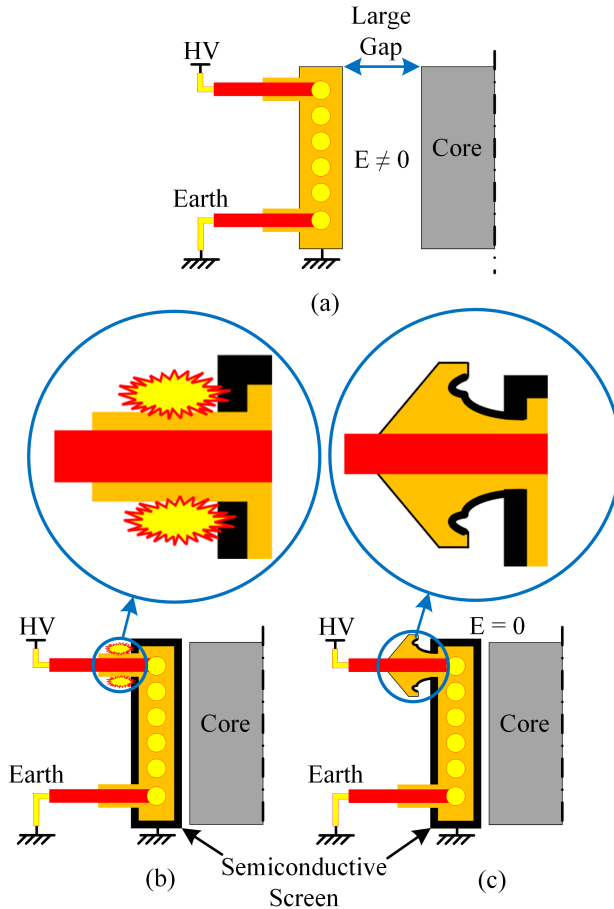


Figure 4.10: Conceptual comparison of electric field behavior at transformer winding terminations: (a) No screening — large gap and uncontrolled electric field ($E \neq 0$) between winding and core, (b) The high electric field concentration at the edge of the semiconductive screen termination without proper grading, leading to discharge risk, (c) Electric field mitigation using a stress cone structure. The field is smoothly graded, reducing the risk of breakdown and enabling more compact insulation design. Although the air gap increases the local electric field strength due to its lower permittivity, it is intentionally included to enable cooling and heat dissipation between the HV and LV windings. The field grading design ensures that PD-free operation is still maintained.

A stress cone in a cable terminal is an excellent example of electrode grading, as shown in Figure 4.11. The cable field is gradually reduced to prevent field concentrations and to create an even potential distribution over the external insulator. The shape of the stress cone is optimized by keeping the potential gradient along the interface, $\Delta P/\Delta x$, constant. The curve can be described as follows [23]:

$$x = \frac{V}{E_t} \left(1 - \frac{\ln(R/r)}{\ln(y/r)} \right) \quad (4.8)$$

where E_t is the tangential electric field, and R and r are the outer and inner radii, respectively, as illustrated in Figure 4.11. E_t is selected to maintain a maximum of 0.5 kV/mm at nominal voltage [23]. Mechanically, R must remain within the boundaries of the epoxy casting and available space. An initial approximation of $2r < R < 3r$ is used in practice, and then refined using FEM simulations to verify field distribution and ensure safe operation under medium-frequency stress conditions. The stress cone for field grading is applied for the field treatment at the end of the semiconductive screen in this chapter.

The stress cone geometry was selected to ensure a smooth transition of the electric field from the high-voltage winding to the grounded surrounding structure. The cone angle and electrode extension length were chosen based on iterative FEM simulations to reduce field peaks and maintain electric field strength below 0.5 kV/mm in the surrounding air and interface regions. The dimensions also satisfy minimum clearance requirements outlined in IEC 60076-11 [13]. This configuration ensures effective field grading while maintaining compactness suitable for dry-type MFT designs.

The combination of the upper stress cone and the surrounding field control ring increases the effective creepage distance and provides smoother field gradients. This integrated approach enables compliance with IEC 60076-11 [13] insulation requirements within a compact geometry. Although stress cones are traditionally used in MV cable terminations, their application in MFTs addresses similar field enhancement issues at winding terminations, particularly in compact dry-type designs where sharp edges and limited air clearances increase the risk of partial discharges.

It is important to note that the effectiveness of stress cone geometries and semiconductive field grading is influenced not only by the voltage amplitude but also by the waveform shape and operating frequency. Under non-sinusoidal excitation, such as quasi-square waveforms, the presence of higher harmonic components and fast voltage transitions can modify the local electric field distribution and surface charge behavior. As a result, field grading solutions designed based on line-frequency assumptions may not fully capture the stress conditions present in medium-frequency SST operation. This aspect is considered

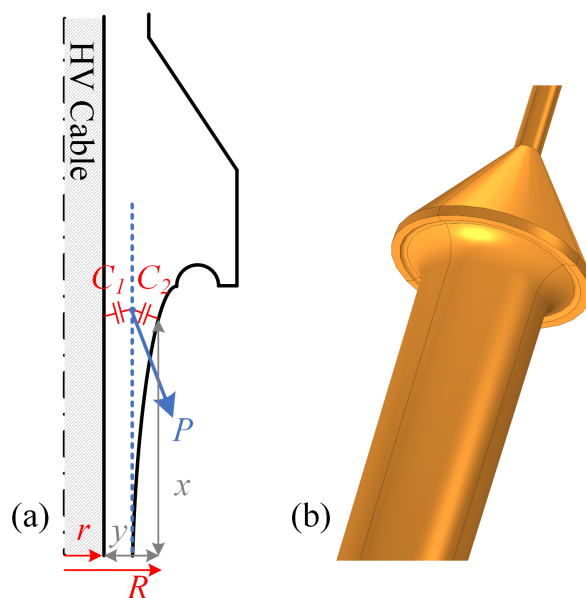


Figure 4.11: (a) The cross-sectional schematic of a stress cone showing the voltage grading path and equivalent capacitive network; (b) 3D view of the stress cone structure.

qualitatively in the present work and identified as an important direction for future investigation.

4.2.5. SEMICONDUCTIVE SCREENING

To reduce the electric field inside the gaps, semiconductive screening is performed to provide a semiconductive surface around the windings. This surface encapsulates the electric field; therefore, the electric field in the gap is zero ($E_{\text{gap}} = 0$ kV/mm). Semiconductive materials are preferred over conductive shields to enhance the system's efficiency. This choice is crucial to mitigate additional losses of the conductive shield. The presence of eddy currents induced by the magnetic field can lead to unwanted extra loss and heat generation, which can be effectively addressed using semiconductive materials. The stickiness of the semiconductive screen to the epoxy resin is essential, as any looseness or the presence of air bubbles can result in localized areas of high electric field strength. Strict adherence ensures uniform distribution of the electric field and prevents potential complications from inconsistencies in the bonding.

The composition of the semiconductive compound includes two parts. Part one is a blend of Butyl Acetate ($\text{CH}_3\text{CO}_2(\text{CH}_2)_3\text{CH}_3$), Ethyl Acetate ($\text{C}_4\text{H}_8\text{O}_2$), and Neopentyl Glycol ($\text{C}_5\text{H}_{12}\text{O}_2$), which is commercially available and used for different purposes such as adhesives and coatings. Part two is graphite powder to provide conductivity. The mixture is 85:15 weight-wise. This balanced mixture ensures optimal conductivity and stability. The semiconductive compound exhibits excellent adhesion to various epoxy resins and has robust conductivity. This versatility and high performance make it an ideal choice for applications requiring reliable electrical conductivity and strong bonding properties. Figure 4.12 shows the semiconductive screen, and Table 4.2 summarizes its characteristics. The compound is commercially available in adhesive and nail polish formulations, and in this chapter, it was applied manually using a small brush to ensure uniform coverage and strong adhesion to the epoxy resin.

Table 4.2: Characteristics of Semiconductive Compound

Base compound	Conductive filler	Ratio	Conductivity
Butyl Acetate, Ethyl Acetate, Neopentyl Glycol	Graphite powder	85:15	5.23×10^{-6} S/m (at 20°C, for typical thickness = 100 μm)*

* The semiconductive layer was manually applied using a fine brush to achieve a minimum target thickness of ~ 100 μm . This thickness was found to be sufficient for PD suppression while avoiding excessive local losses.

Accurately solving complex electric field calculations is important,

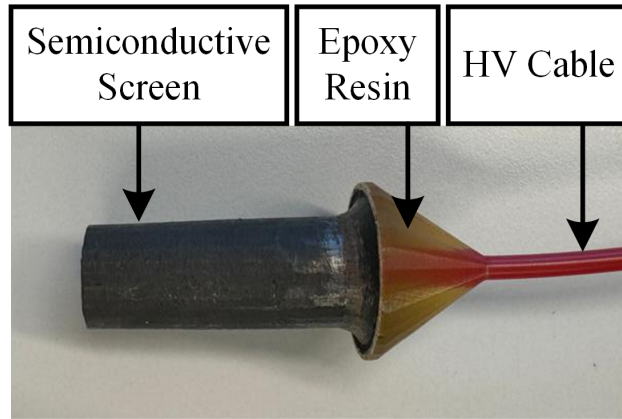


Figure 4.12: Semiconductive screen sticks to the epoxy resin.

however it is analytically complex. The Finite Element Method (FEM), COMSOL Multiphysics software in this chapter, is employed to examine the impact of the semiconductive screening and proper termination of the transformer. A COMSOL 2D axis-symmetrical simulation is performed in this chapter to analyze the effect of semiconductive screening on field grading. The setup includes an electrode, part of a high-voltage cable encapsulated in epoxy resin. A grounded plate electrode is placed to represent the influence of a core adjacent to the high-voltage electrode. Without the semiconductive screen, the simulation reveals an alarmingly high electric field strength in the air gap between the epoxy resin and the grounded electrode.

In contrast, introducing a semiconductive screen on the cone reduces the electric field strength in this gap to nearly zero, as illustrated in Figure 4.13.a and Figure 4.13.b. However, without proper termination, the electric field can be significantly higher on top of the epoxy resin. Proper termination can mitigate the intensity of the electric field, as depicted in Figure 4.13.b and Figure 4.13.c. These results clearly demonstrate the optimization effect of the semiconductive shield in reducing local field enhancements at the stress cone termination.

The stress cone and screening configuration were designed to replicate real dry-type MFT insulation geometries and meet IEC 60076-11 [13] Voltage test requirements were applied in the simulation. The inner electrode was energized at $22.75 \text{ kV}_{\text{rms}}$, while the grounded plate represented the transformer core potential. The outer surface of the semiconductive screen was assigned zero potential to simulate electrical grounding. Material properties were defined as follows: the epoxy resin was modeled with a relative permittivity of 3.5, and the surrounding air with a relative permittivity of 1. These conditions replicate the actual

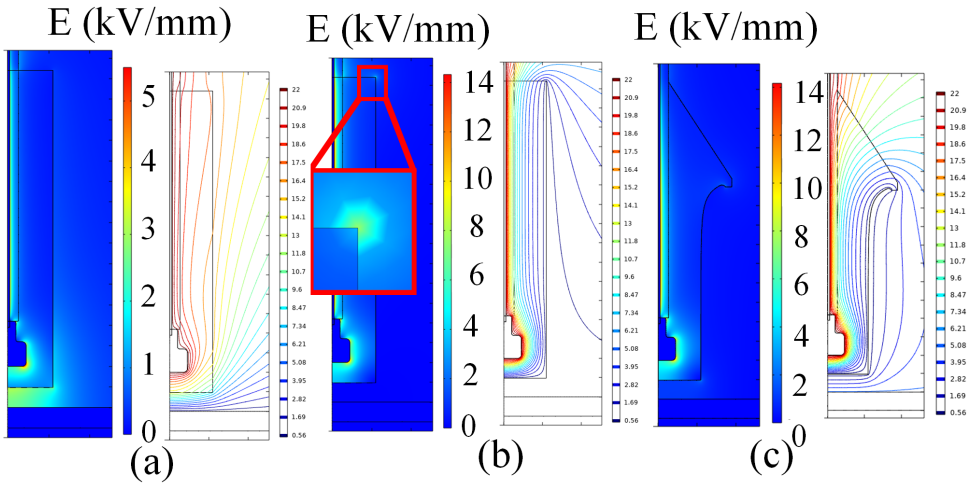


Figure 4.13: Electric field simulation results at the stress cone termination (a) without semiconductive shield, (b) with semiconductive shield but no proper termination, and (c) with both semiconductive shield and optimized stress cone termination. The semiconductive layer significantly reduces the electric field concentration in the air gap and on the epoxy surface.

insulation environment observed in the laboratory tests.

The setup ensures a representative field distribution while remaining suitable for laboratory PD measurement. Future developments of the test campaign will include thermal cycling, long-duration PD endurance testing, and evaluation on a complete transformer system. While the present prototype focuses on epoxy cast insulation around the windings and terminations.

STRESS CONE PREPARATION

The epoxy casting process was performed in a vacuum oven under carefully controlled thermal and pressure conditions to eliminate air voids. This procedure replicates industrial casting methods and was refined in the lab over several years to ensure void-free insulation. The highest voltage for equipment for the transformer is 17.5 kVrms. Therefore, the stress cone for the field grading is designed to handle an applied test voltage of 22.75 kVrms regarding IEC 60076-11 during the partial discharge measurement. The test is fulfilled if the maximum level of PD is 10 pC or less [13]. This voltage level is consistent with IEC 60076-11 requirements for dry-type transformers and corresponds

to typical voltage stress conditions in medium-voltage SST applications. The minimum air clearance from the HV terminal to the grounded screen edge was 28 mm, and the creepage distance along the epoxy surface was over 50 mm.

The cone is made of a two-part epoxy resin called CY225-HY225, which is commonly used in high voltage applications. The manufacturing process for the stress cone is meticulously designed and executed in the following steps:

4

1. **3D Printing of the Cone:** Initially, the cone is crafted using a 3D printer, utilizing Polylactic Acid (PLA) for its construction. This step ensures high accuracy in the cone's dimensions and shape.
2. **Creation of a Silicone Rubber Mold:** Utilizing the 3D-printed cone, a mold is cast from silicone rubber. This mold is vital for accurately replicating the shape and dimensions of the cone in subsequent stages. The silicone mold used to shape the epoxy stress cone was fabricated using a commercially available silicone rubber compound, chosen for its mechanical flexibility and compatibility with the epoxy curing temperature.
3. **Preparation of Cable and Electrode Assembly:** The cable and electrode are carefully positioned within the silicone rubber mold. This setup is then prepared for the next step—the casting of the epoxy resin.
4. **Epoxy Resin Preparation:** The epoxy resin is accurately mixed, pre-heated, and pre-degassed using a vacuum oven. This preparation ensures the resin's optimal flow and adhesion properties.
5. **Casting and Degassing of Epoxy Resin:** The prepared epoxy resin is then cast into the mold. Following the casting, a degassing process is employed to eliminate any trapped air bubbles, ensuring the integrity and electrical properties of the molded component.
6. **Curing of Epoxy Resin:** Once cast, the epoxy resin is heated to its curing temperature, maintained at 140°C. This step is vital for achieving the desired mechanical strength and electrical insulation properties.
7. **Mold Removal and Semiconductive Screening:** The mold is carefully removed after the resin has cured. Subsequently, the semiconductive screen is applied. This final step is pivotal in enhancing the final product's electric field grading capabilities.

Each step in this procedure is essential for achieving the desired characteristics and performance of the stress cone in its application.

Figure 4.14 shows details of the stress cone preparation process. To ensure consistency and eliminate internal voids, the epoxy was mixed and degassed under vacuum before casting and was subjected to additional degassing after pouring. All samples were visually inspected and tested for partial discharge activity to confirm the absence of air bubbles or void-related insulation defects. The semiconductive screen was applied after curing using a brush-coating technique, ensuring uniform coverage across all samples.

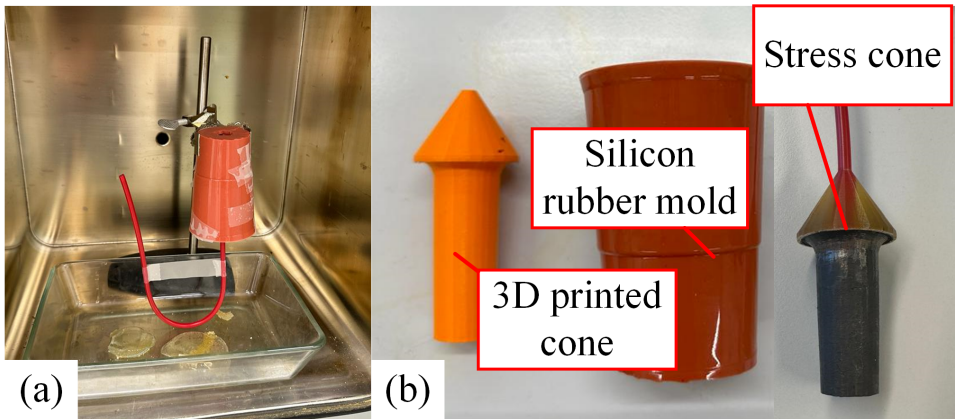


Figure 4.14: (a) The degassing procedure inside the vacuum oven, (b) the 3D printed cone, the silicon rubber mold, and the casted stress cone.

PARTIAL DISCHARGE MEASUREMENT FOR THE STRESS CONE

The partial discharge (PD) measurement in this chapter is conducted using the criteria of IEC 60270 [36] and IEC 60076-11 [13]. The experimental setup, illustrated in Figure 4.15, consisted of six key components: an HV step-up transformer, the test object, a filter, a coupling capacitor, a coupling device, and a measuring device. The setup components are listed in Table 4.3. The background noise of the test setup was measured to be less than 1 pC, ensuring the accuracy of the PD measurements. The results of the PD measurements are presented in Figure 4.16. These figures compare the PD behavior of the stress cone under two different conditions: (a) without semiconductive screening and (b) with semiconductive screening applied. The schematic diagrams represent the test setup for both cases, highlighting the addition of the semiconductive screen in the configuration (Figure 4.16.b). The upper plots present the Partial Discharge (PD) activity over time, showing a significant reduction in PD magnitude and occurrence after applying the semiconductive screen.

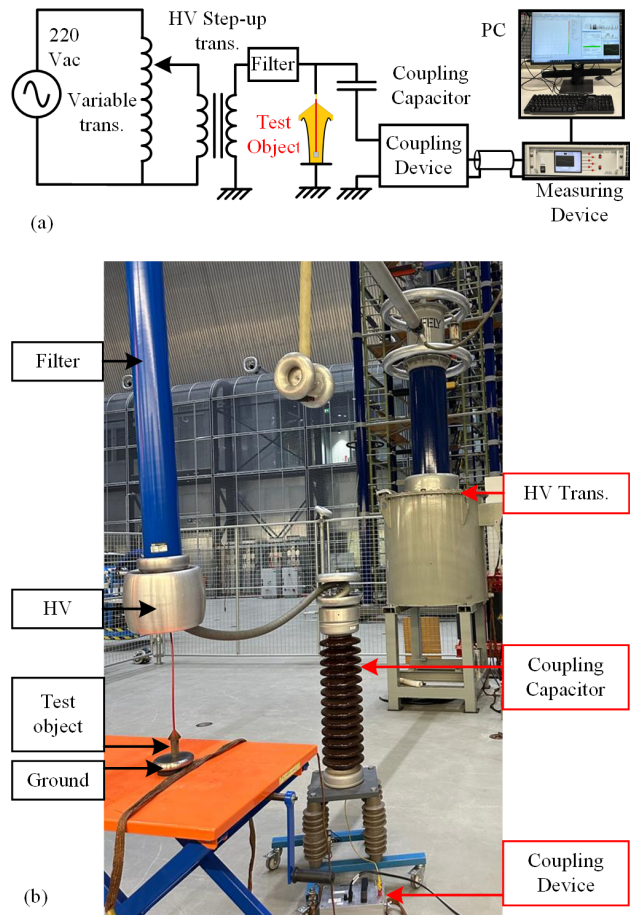


Figure 4.15: The PD measurement setup: (a) the block diagram, (b) the laboratory setup.

Table 4.3: PD Measurement Setup Characteristics

Component	Manufacturer	Type
HV step-up transformer	Haefely	TEOS 0.5/200 kV 100 kVA
Coupling capacitor	Haefely	1062 pF 100 kV
Measuring device	Hubbell	DDX-9101

The Phase Resolved Partial Discharge (PRPD) plots display the phase angle of the applied voltage (0–360°) on the horizontal axis, the PD magnitude in pico-Coulombs (pC) on the vertical axis, and the density of discharge events represented by a color bar.

In the absence of semiconductive screening, as shown in Figure 4.16.a,

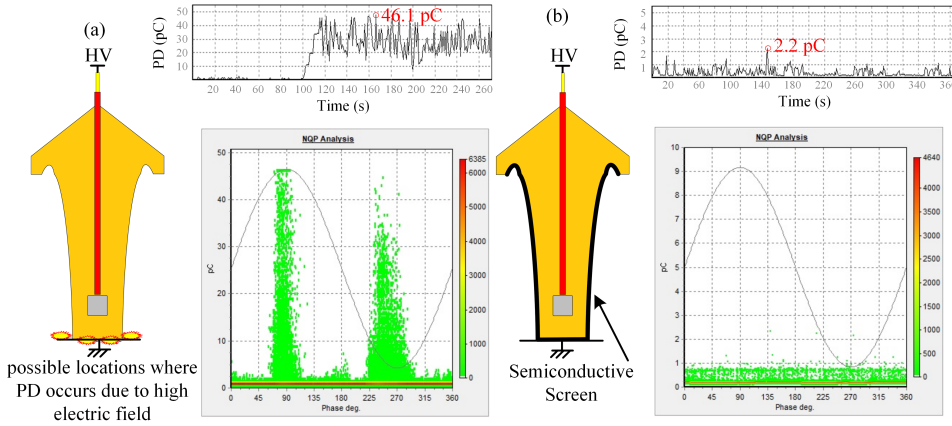


Figure 4.16: Partial discharge (PD) measurement results for the stress cone: (a) without semiconductive screening, showing significant PD activity and localized field enhancements, (b) with semiconductive screening, demonstrating reduced PD activity and improved electric field uniformity. The upper plots illustrate PD over time and the lower plots show the phase-resolved partial discharge (PRPD) patterns.

significant PD activity is observed, with discharge events concentrated around the high-field regions near the voltage peaks. The test object acts as a floating electrode, resulting in localized field enhancements and higher PD activity, with a maximum PD magnitude exceeding 46 pC. As shown in Figure 4.16, PD activities begin at around 110 s, while a considerable number of events fall within the range of 10–40 pC. In contrast, Figure 4.16.b demonstrates the PD performance after applying semiconductive screening to the stress cone. The screening effectively reduces the localized field concentrations, significantly lowering the overall PD activity. No discharge events exceeding 2.2 pC, considering the background noise level of around 1 pC, were recorded during the tests, reflecting a considerable improvement in insulation performance. The results indicate that the semiconductive screening ensures a more uniform electric field distribution, thus mitigating PDs and enhancing the reliability of the insulation system. The observed reduction in PD levels aligns with the design goals for medium- and high-voltage insulation systems, ensuring long-term operational reliability.

The PD measurement results are summarized in Table 4.4. The applied voltage and test duration differ between the two cases due to the presence or absence of semiconductive screening. For the configuration without semiconductive paint, the applied voltage was set considerably lower than the recommended value of 25 kV_{rms}, as significant PD activity

caused saturation of the PD detector amplifier. In contrast, after the application of the semiconductive paint, the PD activity was significantly suppressed. This allowed the test voltage to reach the recommended level (25 kV_{rms}) and the test duration to be extended to 360 s. The maximum PD level and the number of PD events were significantly reduced, demonstrating the effectiveness of semiconductive screening in improving insulation performance. This direct comparison highlights the effectiveness of semiconductive screening in reducing partial discharge activity and supports the practical necessity of the proposed method.

Table 4.4: PD Measurement Results

	Max Applied Voltage (kV _{rms})	Max PD Level (pC)	Test Time (s)	Noise Level (pC)
No Semi. Con. Paint	17.5 kV	46	250	1
Semi. Con. Paint	25 kV	2.2	360	

4.3. USING 3D-PRINTED PARTS AS INSULATORS

Using 3D printers to build the insulation parts of the transformers is an interesting idea mainly because of their fast production rate, precision, and competitive prices. 3D printers allows for rapid prototyping.

While employing 3D printers for low-voltage, low-power applications is promising, using 3D-printed parts in medium-voltage and high-power applications is challenging due to their thermal performance and unknown insulation behaviors. One of the most essential insulation behaviors to study for 3D-printed insulators is the Partial Discharge (PD) inside them. PD is the development of an ionized path that fails to bridge the whole gap between the electrodes and, therefore, can dissipate only limited energy [37–39]. On the one hand, PDs are quality indicator for an insulation system. On the other hand, PD itself, can damage the insulator and cause extra losses and extra heat. Therefore, a thorough investigation into the partial discharge behavior of 3D-printed materials is crucial to ensure reliability and safety in medium and high-voltage applications.

There are several studies on this topic. Reference [40] gives development and selection of suitable materials, process optimization, and integration of 3D-printed components in medium voltage power electronics. Authors in [41] investigate the dielectric and electrical properties of various 3D-printed thermoplastic polymers, including PLA, polyethylene terephthalate (PET), ABS, polycarbonate (PC), and

polyamide (PA), produced using Fused Filament Fabrication (FFF) for potential use in energy technologies. The study examines dielectric breakdown strength, permittivity, and dissipation factor, finding that while 3D-printed materials generally have lower dielectric strength than conventionally manufactured counterparts due to small cavities from the printing process, they may still be suitable for applications where geometric complexity is more important than high field strength or compactness.

Reference [42] studies the potential of using 3D printing technology with a micro-cellulose-doped polymer material for manufacturing high-voltage transformer insulation components. While the basic dielectric properties of the printed material are satisfactory, the study finds that the current 3D printing process results in internal voids that negatively impact electrical breakdown strength compared to traditional injection molding techniques.

Authors in [43] examine the dielectric properties of 3D-printed PETG, focusing on the effects of flame retardant additives. The study found that while the additives increased volume resistivity, they also led to higher space charge accumulation and reduced electrical breakdown strength, potentially impacting the material's effectiveness as insulation in high-voltage applications. These results highlight the need for careful consideration of additive effects when using 3D-printed materials in electrical insulation.

While the previous works significantly contribute to using 3D-printed materials in electrical insulation, the PD behavior of such parts is not investigated thoroughly. Hence, this chapter explores the partial discharge in 3D printed materials compared to the same trend in bulk materials.

4.3.1. PARTIAL DISCHARGE IN THE 3D PRINTED INSULATOR

In a 3D-printed part, there is a high probability of having cavities between the layers of the 3D-printed material. Thus, investigating the PDs caused by cavities is essential. The electric field, E , inside an insulation system is given by:

$$E = \frac{V}{d} \quad (4.9)$$

where V is the applied voltage, and d is the distance between the electrodes. In the presence of a defect, such as a void, the local electric field can become significantly higher than the average field, leading to the initiation of PD. The initiation of partial discharge occurs when the electric field within a void exceeds the breakdown strength of the gas inside it. This can be expressed as:

$$E_{void} = \frac{V_{bd}}{d_{void}} \quad (4.10)$$

where E_{void} is the electric field within the void, V_{bd} is the breakdown voltage of the gas inside the void, and d_{void} is the size of the void.

Partial discharge (PD) can be modeled as a series of short-duration current pulses $i(t)$. The current pulse $i(t)$ associated with a single PD event can be represented by:

$$i(t) = q \cdot \delta(t - t_0) \quad (4.11)$$

where q is the charge released during the discharge, and $\delta(t - t_0)$ is the Dirac delta function centered at time t_0 , indicating the occurrence of the discharge. The total current associated with multiple PD events is the sum of these individual pulses:

$$I(t) = \sum_n q_n \cdot \delta(t - t_n) \quad (4.12)$$

where q_n and t_n represent the charge and time of the n -th PD event, respectively. The PD detector measures the current pulses corresponding to each partial discharge and reports the associated charge in pico-Coulombs (pC). Every dot in the scatter plot represents the occurrence of one PD event, and the intensity of occurrences is visualized using a color map. The background noise can result in a high number of low-magnitude PD events. However, these must be filtered out to avoid misleading interpretations. For the test setup used in this chapter, a threshold of 1 pC is considered the background noise level, and events below this threshold have been filtered accordingly.

The voltage at which partial discharge initiates, known as the inception voltage V_{inc} , depends on the size and shape of voids in the insulation as well as on the gas pressure inside the void, as described by Paschen's law. Assuming a homogeneous electric field, PD initiates when the electric field inside the void exceeds the breakdown strength of the gas. This condition can be expressed as

$$V_{inc} = \frac{V_b d}{D} F \geq \frac{B p d}{\ln(p d) + k} \quad (4.13)$$

where V_b is the applied voltage between the electrodes, d is the void thickness, D is the total insulation thickness between the electrodes, p is the gas pressure inside the void, and B and k are gas-dependent Paschen constants. The factor F accounts for the shape of the void and the resulting local field enhancement. This relation shows that PD can initiate at applied voltages significantly lower than the bulk breakdown voltage of the insulation system.

As shown in Figure 4.17, a non-conventional PD measurement system consisting of a High-Frequency Current Transformer (HFCT), high-frequency amplifier, and oscilloscope was used to capture the PD activity in 3D-printed samples. The test setup is inside a Faraday cage to minimize electromagnetic interference (EMI) and mitigate the background noise. The test setup and the test object are shown in Figure 4.17. The measuring device records the number of PDs, their intensity, and the number of occurrences over the sampling time. The materials used for the test can be divided into two categories: bulk materials and 3D-printed ones. These materials and their relative permittivities are listed in Table 4.5 [44–46].

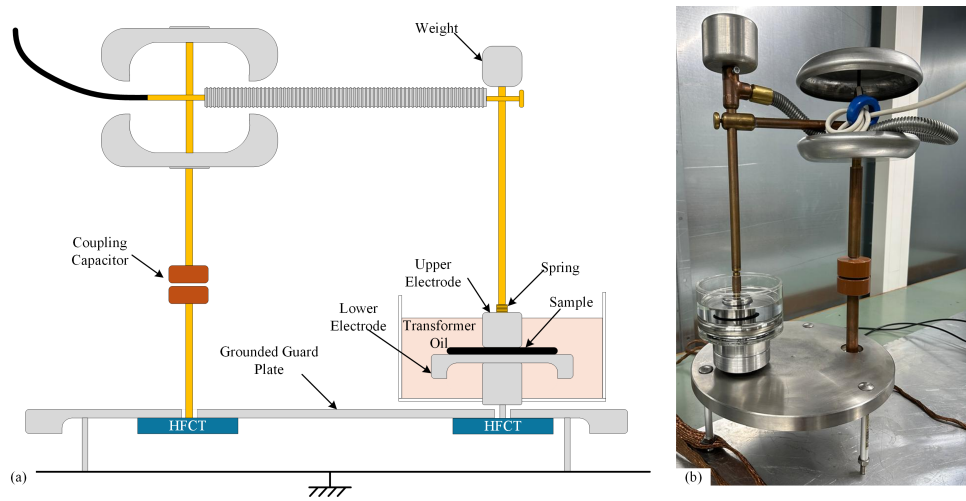


Figure 4.17: (a) The schematic of the test setup, and (b) the test setup inside the Faraday cage.

Table 4.5: The characteristics of tested materials.

Name	ϵ_r	Melting Temp. (°C)	Deflection Temp. (°C)	Type
PLA	3.76	160	57	3D
PETG	3.98	225	69	3D
PC	3.55	228	117	3D
PTFE	2.10	327	121	bulk
PMMA	3.79	160	105	bulk

In order to improve the modeling accuracy and veracity of the magnetic and electric FEM simulations, a spectrography study is performed on 3D printed disk samples of *polycarbonate* plastic. The polycarbonate is manufactured by *BambuLab*. Circular samples of 4 cm diameter are printed using additive technology at an infill of 100%, attempting to minimize the amount of trapped air cavities within the

sample. Fifteen samples are prepared with increasing thickness, as listed in Table 4.6.

Table 4.6: PC Sample thicknesses

Sample	1-3	4-6	7-9	10-12	13-15
Thickness (mm)	0.4	1	1.4	2	2.4

The goal of the study is to inspect the permittivity and loss profile of *polycarbonate (PC)* at different thicknesses. To do so, 3 V_{rms} are applied on the sample and capacitance is measured across the electrodes. The measurement is performed in a frequency range of quasi-DC (0.01 Hz) to 10 MHz. Every measurement is performed at a constant temperature of 22°C. The permittivity and dielectric loss are plotted in Figure 4.18 and Figure 4.19.

4

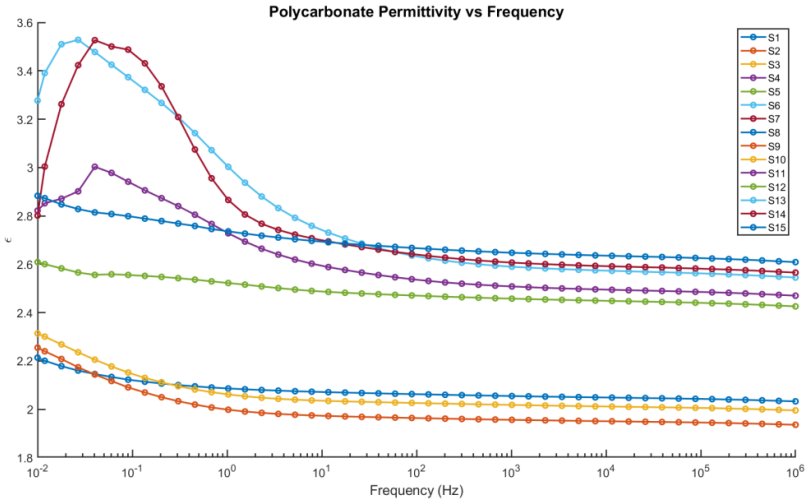


Figure 4.18: Permittivity of PC samples.

From the measurements, it can be seen that large deviations in permittivity and loss occur across the frequency spectrum, and only weak correlation can be found between permittivity/loss and sample thickness. This is particularly evident at low frequencies below 1 Hz. At the operating frequency of the MFT (1 kHz), permittivity varies between 2 and 2.8, with a larger correlation with sample thickness. For the purpose of electric field simulations performed in this thesis project, a value of 2.6 is assumed.

The PD measurement platform can be found in [47]. The samples were made of PC, with 100% and 50% infill, and a thickness of 1 mm. As shown in Figure 4.17, the samples were submerged in

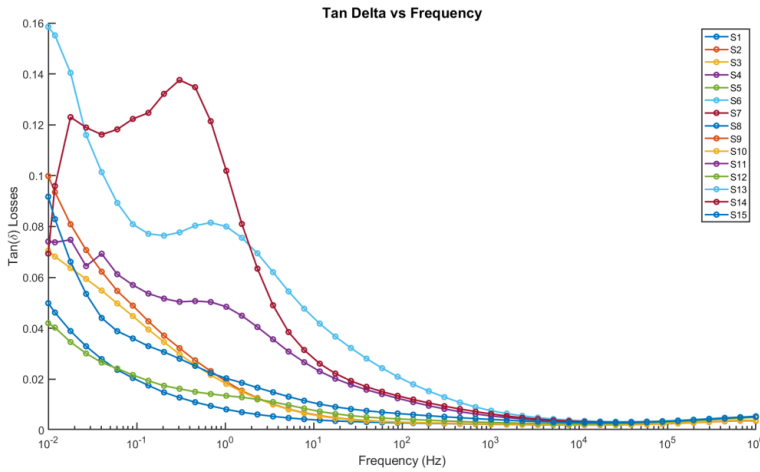


Figure 4.19: Dielectric loss of PC samples.

oil. Initially, the applied voltage to the sample was raised from zero to the PD inception voltage. Then, the inception voltage was fixed, and PD activity was constantly recorded until a full breakdown was achieved. The PD measurement results indicated the presence of two dominant discharge sources: one originating in the surrounding oil and another inside the 3D-printed material. These two mechanisms can be distinguished by their temporal pulse characteristics. Discharges occurring in oil typically appear as bursts or trains of damped oscillatory pulses, caused by streamer activity that generates multiple closely spaced micro-discharges [48]. In contrast, discharges associated with voids or defects inside solid insulation generally appear as isolated single pulses, corresponding to localized cavity or tree-like discharge events. Figure 4.20 shows an example of an individual PD pulse attributed to solid-insulation discharge, whereas Figure 4.21 illustrates a pulse train characteristic of PD activity in oil.

The PD pulses containing only single discharge events (i.e., no burst-type activity) were selected for plotting the phase-resolved partial discharge (PRPD) patterns corresponding to different stages of PD activity. Since partial discharges were continuously measured from inception until breakdown, the entire dataset was divided into batches containing thousands of pulses. For the 3D-printed PC material with 100% infill and subjected to 24-hour degassing, the PRPD patterns for different time stages are presented in Figure 4.22, where a tree-like behavior can be identified. For this sample, the inception voltage was 20 kV. Similar to a tree in epoxy resin [49]. From the results, it is possible to conclude that the PD activity induces a breakdown in the 3D

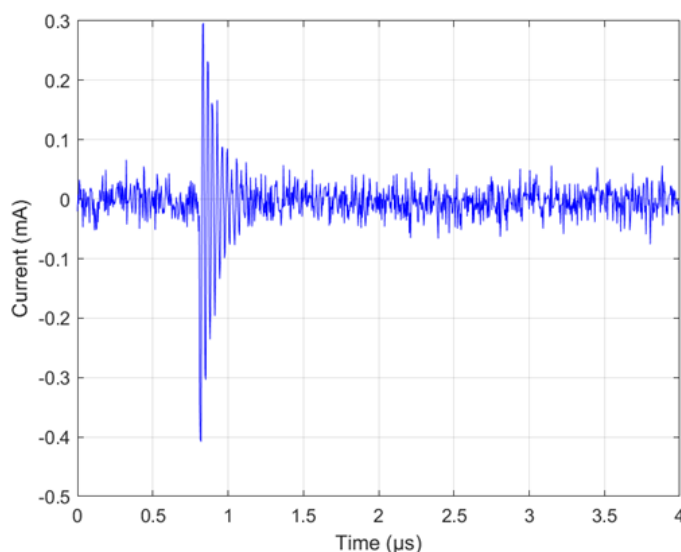


Figure 4.20: Current waveform of an individual partial discharge (PD) pulse captured by the HFCT. The damped oscillatory pulse peaks at approximately 0.3 mA and lasts about 1 μ s.

printed material because the PRPD patterns indicate a tree propagation inside the material.

The same analysis was performed for a PC sample with 50% infill and 1 mm thickness. The inception voltage was 12.1 kV, and the PRPD patterns for different stages are presented in Figure 4.23. In this case, fewer PD pulses were captured since breakdown was quickly reached. The PRPD patterns look very similar to the ones with 100% infill. It is also assumed that there is a tree-like propagation of the PD activity.

4.3.2. PARAMETERS OF 3D PRINTED INSULATION SYSTEM

To keep the consistency, the time frame for the following experiments is set to 1 minute.

3D-PRINTED SAMPLE VS BULK MATERIAL

To examine the impact of 3D printing the insulator parts vs. other fabrication methods, 3D-printed materials are compared to bulk materials with the same thickness, as shown in Figure 4.24. This figure proves that the difference between the number of PDs is not because of having small cavities inside the 3D-printed samples, it is mainly because of the edging effect of the HV electrode since higher the relative permittivity, higher the PDs.

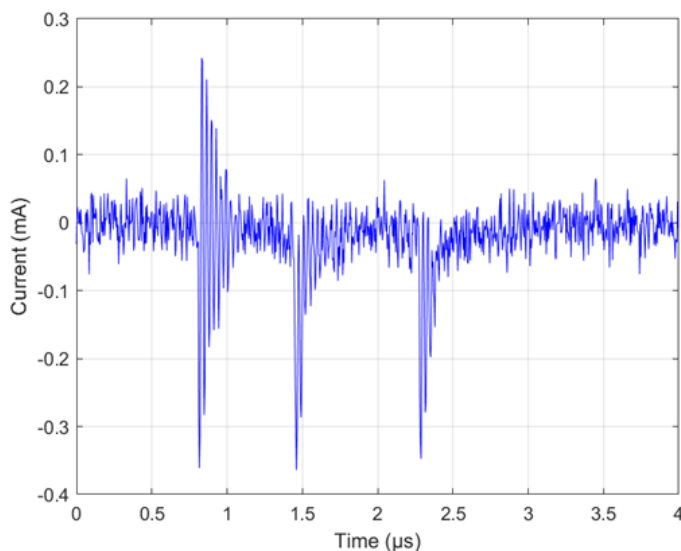


Figure 4.21: Train of partial discharge (PD) pulses captured by the HFCT. Multiple damped oscillatory pulses can be observed, each corresponding to an individual discharge event within a short time window.

THE IMPACT OF THICKNESS

The thickness of the test sample affects the insulation performance of the dielectric due to changing the electric field strength inside the test sample. The effect of thickness is investigated for two materials: 1) 3D-printed PLA and 2) bulk PMMA, as shown in Figure 4.25 and Figure 4.26, respectively.

THE IMPACT OF SPARSE INFILL DENSITY

The printing parameters impact the insulation performance of the 3D-printed samples. One of the most critical parameters is the material's strength, often described as "sparse infill density" in the slicer environment. This parameter defines the quality of the print inside the sample. The higher the density, the smaller the gaps inside the sample, resulting in better insulation performance, as shown in Figure 4.27. Note that the size of the nuzzle also plays a role in the size of the gaps; however, the impact of the sparse infill density is dominant. This parameter is shown in Figure 4.28.

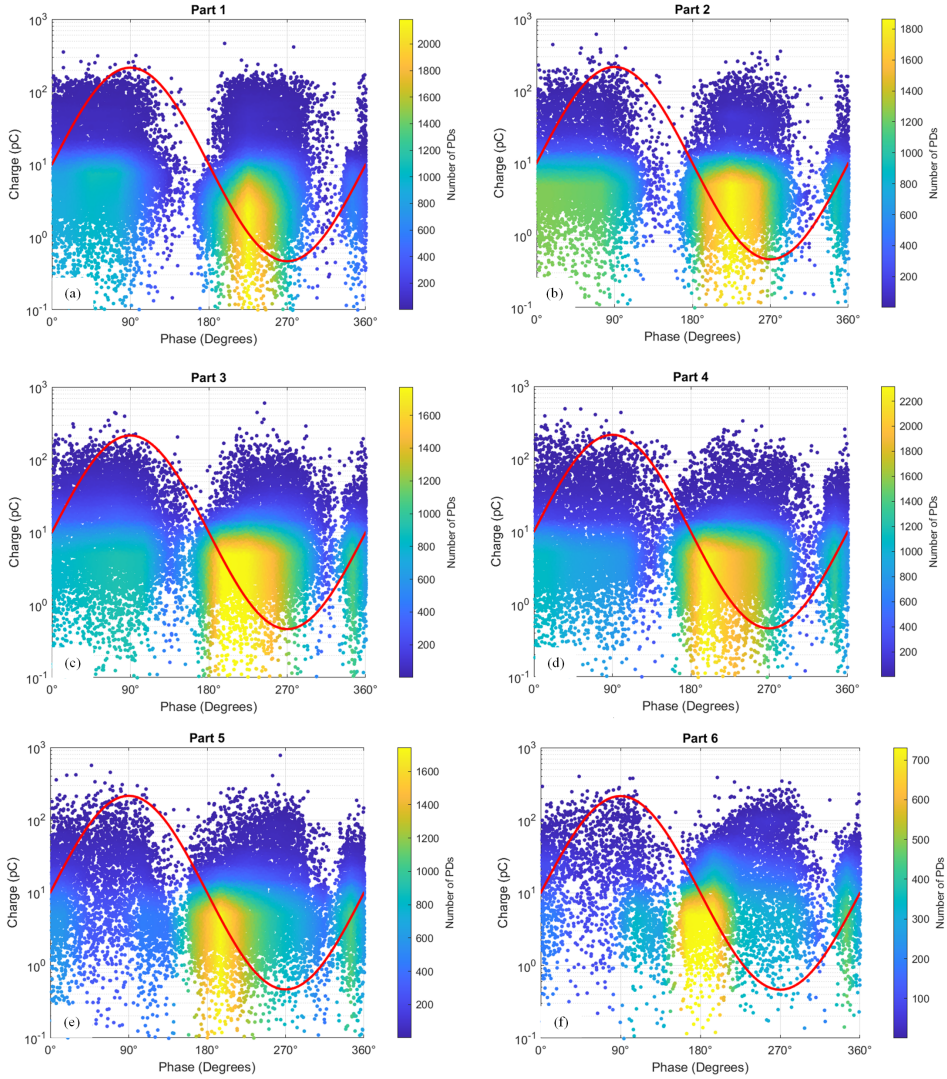


Figure 4.22: PRPD patterns: (a) stage 1, corresponding to the first 50,000 pulses measured right after inception, (b) stage 2; 50,000 pulses, (c) stage 3; 50,000 pulses, (d) stage 4; 50,000 pulses, (e) stage 5; 50,000 pulses, and (f) stage 6, corresponding to the pulses recorded before the breakdown.

THE EFFECT OF DIFFERENT MATERIALS

Different 3D-printed materials have different PD behaviors, as shown in Figure 4.29. This can be due to their different ϵ_r or different printing

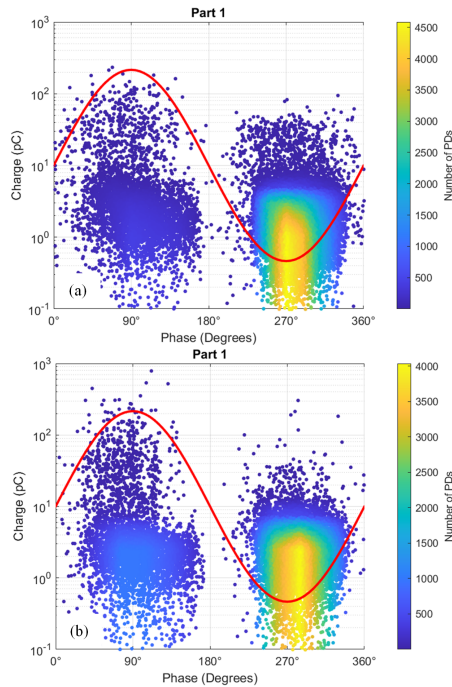


Figure 4.23: a) Initial PRPD pattern after inception, and b) PRPD pattern before breakdown.

because of distinct thermal performance.

It is noteworthy that deflection temperature is critical for an insulator. On the one hand, the 3D-printed parts are often placed near the heat sources (windings and core of the transformer); thus, they must endure high temperatures. On the other hand, the 3D-printed parts might be filled with epoxy resins. Some of the most commonly used epoxy resins in the transformer industries, such as [50], require high temperatures for degassing and curing procedures to ensure reliable insulation behavior. For instance, the epoxy resin CY 225 and the hardener HY 225 must be cured for a minimum of 4 hours at 130°C , according to the datasheet [50].

Before the curing process, degassing must be maintained for several hours based on the mixing method and shape of the mold. The 3D-printed parts, therefore, must withstand during this procedure without melting or deflecting. Regarding Figure 4.29 and thermal performance, PC (polycarbonate) shows promising performance in both insulation and thermal considerations.

Another possibility to enhance the thermal performance of 3D-printed parts is adding some carbon fibers into them. Adding carbon fiber

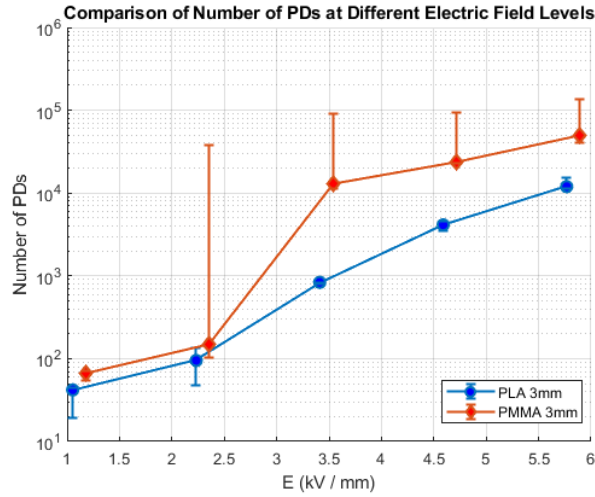


Figure 4.24: Number of PDs over different electric fields for PLA vs PMMA.

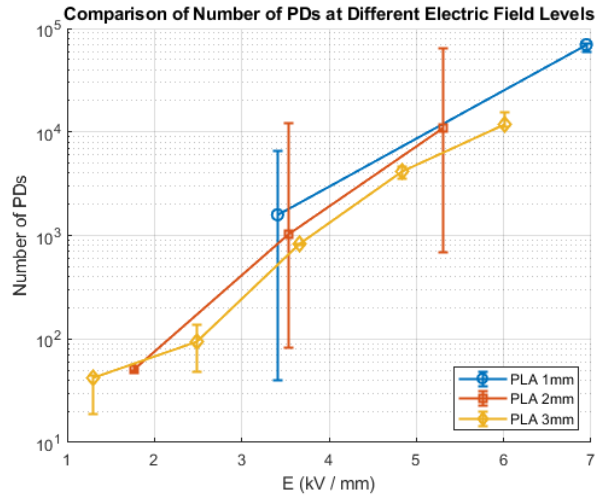


Figure 4.25: The impact of thickness on the number of PDs in 3D-printed PLA samples with 1 mm, 2 mm, and 3 mm thickness.

enhances the thermal performance of the 3D parts; nevertheless, it decreases the insulation characteristics significantly since carbon particles are electrically conductive. Thus, CF filaments are not recommended for high- or medium-voltage applications. The thermal performance of the three materials is compared in Figure 4.30.

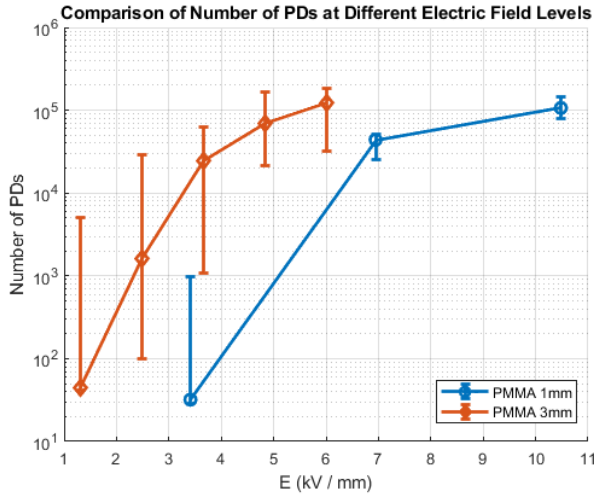


Figure 4.26: The impact of thickness on the number of PDs in PMMA samples with 1 mm and 2 mm thickness.

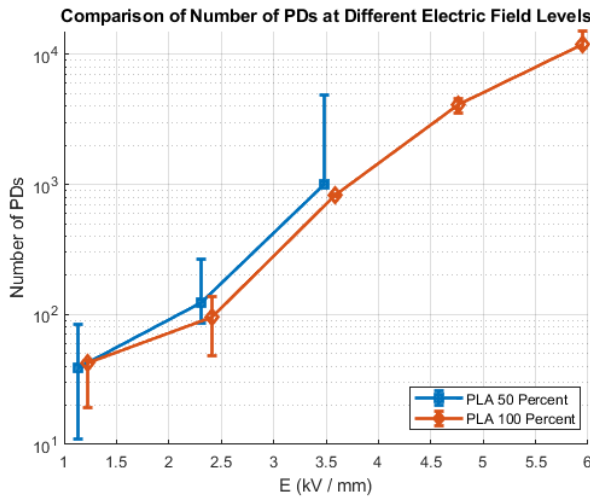


Figure 4.27: The impact of sparse infill density on the number of PDs in PLA samples with 50% and 100% density.

4.3.3. DESIGN GUIDELINES

In order to use the 3D-printed parts in the high- or -medium-voltage transformers, especially when they are considered as a part of an insulation system, the following factors and guidelines can be helpful:

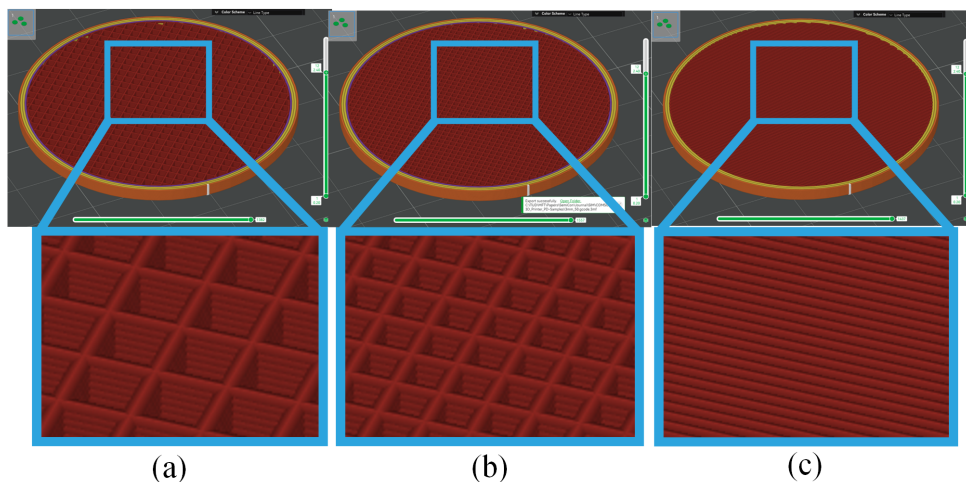


Figure 4.28: Sparse infill density: a) 50%, b) 75%, and c) 100%.

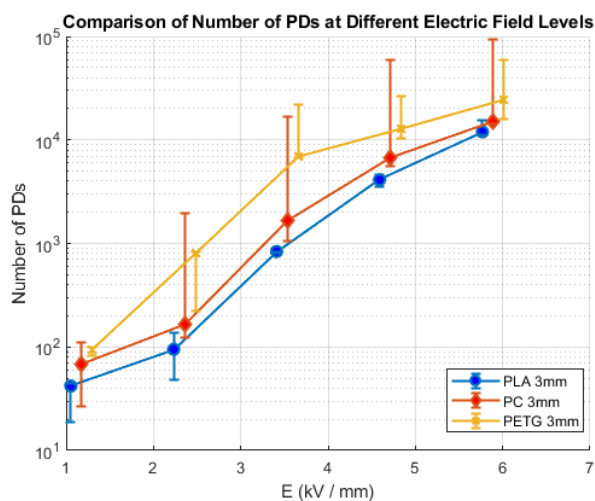


Figure 4.29: The comparison of PDs in three materials: PLA, PETG, and PC.

1. Use the smallest possible nozzle.
2. Set the sparse infill density to 100%.
3. select a thickness based on the (peak)voltage, keep the field strength below 3 kV/mm .
4. Avoid having sharp edges, gaps, and other shapes resulting in field

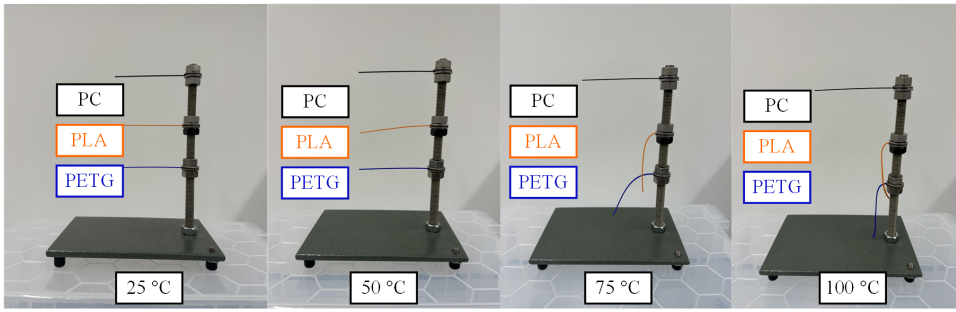


Figure 4.30: Thermal performance of 3D-printed filaments PC, PLA, and PETG under different temperatures: 25 °C, 50 °C, 75 °C, and 100 °C.

enhancement near or inside the insulator.

5. Consider the thermal performance of the selected material during the fabrication and operation. PC (polycarbonate) can be a good option.

4.4. OIL-TYPE TRANSFORMERS

The oil-type transformers differ from dry-type transformers in the cooling and insulation principles. While their cooling principles are explained in Chapter 3, particularly in Figure 3.6, the insulation considerations need more clarification.

The design of an oil-type transformer must be done according to IEC standards, in particular, IEC 60076-1 [51] and IEC 60076-3 [12].

4.4.1. MFT INSULATION FEM SIMULATIONS

To satisfy the insulation distances d_1 to d_4 , in Figure 3.7, these distances between the active parts (MFTs) and the walls of the transformer tank are determined by considering three main factors:

1. **Electrical Clearance:** Ensuring adequate spacing to prevent dielectric breakdown in transformer oil, in accordance with IEC 60156 standards [52].
2. **Thermal Management:** Proper spacing is essential for effective heat dissipation. This is particularly important due to the different thermal characteristics of transformer oil compared to air.
3. **Mechanical Stability and Safety:** Sufficient space must be provided to accommodate the thermal expansion of the oil, core, and winding components.

Among these considerations, thermal management is the most critical. The permissible electric field values at critical locations within the oil-filled transformer are carefully evaluated to ensure reliable long-term operation, as summarized in Table 4.7.

Table 4.7: Permissible Field Strength at the Critical Points of an Oil-Type Transformer [34]

No	Critical point		Electric field (kV/mm)
1	Oil, between barriers		10
2	Oil, along interfaces		<5
3, 4	Oil-impregnated paper	Perpendicular to layers	30 – 60
		Parallel to layers	6 – 10

Given the inherent uncertainties in these figures, a safety margin is crucial. Considering the maximum allowable electric field within the transformer oil to be 5 kV/mm and implementing a safety margin, the maximum electric field is 2.5 kV/mm. This conservative approach is further reinforced by the calculation for the minimum required gap between the windings and the tank, based on the peak of the maximum applied test voltage ($\sqrt{2} \times 70 \text{ kV} \approx 100 \text{ kV}$), resulting in:

$$G_{\min} = \frac{V_{\text{test_max}}}{E_{\text{oil_max}}} = \frac{100}{2.5} = 40 \text{ mm} \tag{4.14}$$

where $G_{\min}$ is the minimum required gap, $V_{\text{test_max}}$ is the peak of the maximum applied test voltage, and $E_{\text{oil_max}}$ is the maximum permissible electric field inside the oil. Then d_1 – d_4 must be greater than $G_{\min}$.

However, to ensure sufficient spacing for the effective laminar flow of oil, which is critical for both cooling and electrical insulation purposes, in this design d_1 , d_2 , and d_4 are selected as 100 mm—significantly exceeding the minimum requirement. Also, d_3 is selected as 300 mm to guarantee enough spacing on top of the transformer for convection heat dissipation. This extra spacing helps mitigate any potential risk associated with the alignment and positioning of the internal components during operation and maintenance, thereby enhancing the overall system reliability and robustness.

The FEM simulation of the electric filed is shown in Figure 4.31, which indicates the electric filed between the primary winding and core, and between disks of the primary winding, and between turns of a disk regarding the values of Table 4.7.

4.5. POWER ELECTRONICS INSULATION REQUIREMENTS

In the 3D design shown in Figure 4.32, Four insulation distances have been analyzed:

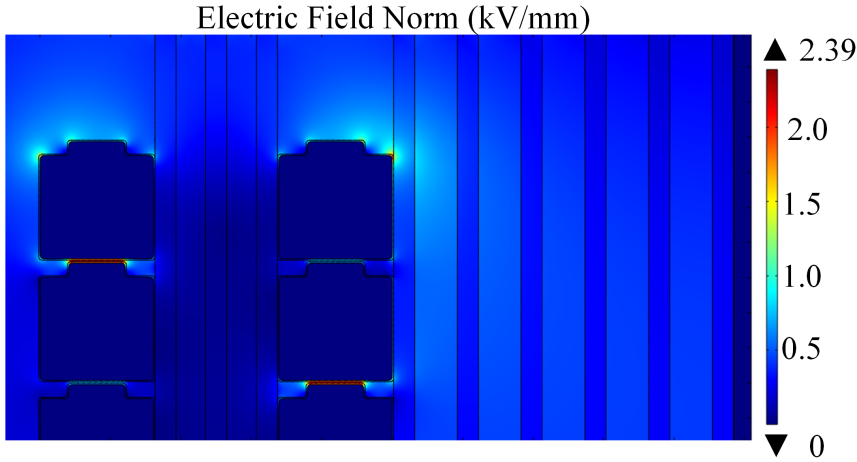


Figure 4.31: The FEM simulation of electric field in the primary disks. The press-board barriers and the CTC conductors are visible. The maximum electric field is 2.39 kV/mm.

1. **The distance between modules d_1 :**

The standard rated lightning impulse withstand voltage for the voltage difference between modules is

$$\frac{U_{dc}}{2N_{sm}} = \frac{23.1 \text{ kV}}{24} = 0.96 \text{ kV} \quad (4.15)$$

The corresponding standard lightning impulse withstand voltage is 40 kV, and the minimum clearance required for this voltage is 60 mm, which is set as d_1 .

2. **The distance between the rows of the converter towers d_2 :**

The voltage difference between rows is

$$\frac{23.1 \text{ kV}}{24} \times 16 = 15.4 \text{ kV} \quad (4.16)$$

requiring a lightning impulse withstand voltage of 95 kV. Based on the manufacturer datasheet [53], the insulator height required for 95 kV is 255 mm, which is set as d_2 .

3. **The distance between the lower row and the ground d_3 :**

The maximum output phase voltage is

$$\frac{U_{dc}}{2} = 23.1 \text{ kV} \quad (4.17)$$

and the required lightning impulse withstand voltage is 95 kV. The same insulator height of 255 mm as in the previous case is used, hence d_3 is set as 255 mm.

4. The distance between the converter towers d_4 :

The voltage difference between converter towers of the same phase is 46.2 kV DC, which is equivalent to an AC RMS voltage of 32.7 kV. The required lightning impulse withstand voltage is 170 kV peak, for which the required clearance is 32 cm. However, for ease of maintenance in case of module failure, this distance is increased to 1 m, which is set as d_4 .

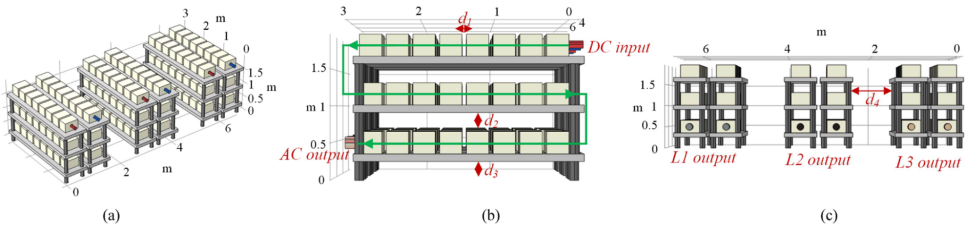


Figure 4.32: MMC converter 3D modelling: (a) overall structure, (b) side view 1, (c) side view 2. There are six converter towers in total, corresponding to six arms. Each tower has three rows, and each row contains eight submodules. L_1 , L_2 , and L_3 represent the three-phase MMC AC outputs.

4.6. CHAPTER SUMMARY AND KEY FINDINGS

This chapter investigated insulation coordination aspects relevant to medium-voltage, medium-frequency solid-state transformers (SSTs), with emphasis on dry-type demonstrator design and transferable insights for full-scale oil-immersed implementations. The main findings are summarized as follows:

- **Dry-type MFT insulation is primarily limited by PD and surface flashover, not bulk breakdown.** Analytical expressions for concentric geometries provide physical insight into the radial field distribution, but practical windings exhibit strong local field enhancements at discontinuities (e.g., seams, bolt heads, sharp edges). Therefore, insulation design must be based on peak field stress and interface behavior rather than average field values.
- **Field enhancement at micro-geometry can dominate PD inception.** Geometrical irregularities can increase the local field by factors of 2–3, while micro-protrusions may yield much larger

enhancement factors. Such localized peak fields can trigger ionization mechanisms at gas-solid interfaces even when the macroscopic average field remains moderate.

- **Semiconductive screening effectively confines the electric field and enables compact insulation.** Inspired by HV cable technology, semiconductive screening was implemented around epoxy-encapsulated windings to suppress the electric field in surrounding air gaps. FEM results confirm that screening reduces the air-gap field to nearly zero; however, improper termination may create hazardous field peaks at the screen edge.
- **Stress-cone termination is a practical solution for controlling the end-field of the screen.** A stress-cone-based electrode grading structure was designed and optimized to mitigate end-field intensification. Combined screening and stress-cone grading produced a smoother field distribution and reduced local maxima at the termination region.
- **Experimental validation demonstrates substantial PD suppression.** PD measurements performed according to IEC 60270 and IEC 60076-11 show that semiconductive screening reduced the maximum PD magnitude from approximately 46 pC (unscreened case) to below 2.2 pC at 25 kV_{rms}, approaching the noise floor of the measurement setup.
- **3D-printed insulation parts exhibit PD behavior strongly influenced by printing parameters.** HFCT-based measurements on 3D-printed PC samples indicate that internal defects and interfaces can produce PD activity evolving toward breakdown. Printing parameters such as thickness and sparse infill density significantly affect PD inception and progression. In particular, higher infill density and avoidance of sharp edges reduce PD activity, while material selection must also satisfy thermal constraints imposed by epoxy casting and operation.
- **Oil-immersed insulation coordination is primarily driven by thermal and clearance constraints.** For the oil-type concept, electrical clearances were conservatively selected based on permissible field values and safety margins, and then further increased to support reliable oil flow and heat dissipation. FEM simulations confirm that the resulting peak electric field remains within the selected conservative limits.
- **Power electronics clearances were derived from standardized withstand levels.** Insulation distances in the MMC submodule arrangement were determined based on the relevant voltage differences and corresponding lightning impulse withstand

requirements, with additional margin where required for practical maintenance and accessibility.

Overall, the chapter shows that compact insulation design for MV/MF SST transformers requires explicit control of local electric field peaks and termination fields. Semiconductive screening combined with stress-cone grading provides a practical and experimentally validated approach to achieve PD suppression while maintaining compactness.

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5

PARASITIC PARAMETERS AND CURRENT BALANCING

Opposites are not contradictory but complementary.

Niels Bohr

This chapter presents a comprehensive analysis of the parasitic parameters and current balancing challenges in medium-frequency transformers (MFTs) used in solid-state transformer (SST) systems for green hydrogen production. It begins with a detailed examination of the parasitic resistances, inductances, and capacitances inherent to transformer windings, highlighting their frequency-dependent behavior and impact on high-current, high-frequency operation. Analytical and numerical techniques, including FEM and energy-based methods, are used to evaluate parasitic effects. Particular attention is given to the issue of current unbalancing in parallel-connected winding disks, which is critical in high-current secondary windings. A novel mitigation technique is introduced, based on series-connected compensation inductors using compact toroidal cores. The method is validated through simulation and experimental measurements, demonstrating improved current sharing, reduced localized losses, and enhanced thermal performance. The findings contribute to the design of more efficient and reliable SSTs in large-scale renewable energy applications.

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This chapter discusses the modeling and impact of parasitic parameters in medium-frequency transformers (MFTs), with a focus on their influence on current distribution in high-power applications. It begins by describing the origin and behavior of parasitic resistances, inductances, and capacitances in transformer windings. Analytical models and numerical simulations are used to evaluate these effects, particularly under high-frequency operation. The chapter then examines the current unbalancing issue in parallel-connected winding disks, explaining its root causes and consequences.

A mitigation approach based on series compensation inductors is introduced and analyzed both through simulation and experimental validation. The final sections present the design, testing, and results of a prototype, demonstrating the effectiveness of the proposed method in achieving current balance and reducing localized losses.

5

5.1. PARASITIC PARAMETERS OF AN MFT

Any transformer can be modeled as an ideal component that transfers voltage and current between its primary and secondary windings according to the turn ratio. The voltage transformation follows:

$$\frac{V_s}{V_p} = \frac{N_s}{N_p}, \quad (5.1)$$

and the current transformation is:

$$\frac{I_s}{I_p} = \frac{N_p}{N_s}, \quad (5.2)$$

where V_p and V_s are the primary and secondary voltages, I_p and I_s are the primary and secondary currents, and N_p and N_s are the number of turns in the primary and secondary windings, respectively.

However, the ideal transformer model does not account for the non-ideal behaviors present in a real transformer. These non-ideal behaviors include parasitic resistances, parasitic inductances, and parasitic capacitances, which affect the performance of the transformer, especially at high frequencies.

The parasitic resistances represent the ohmic losses in the windings and can be modeled as series resistances with the windings. The core loss of the transformer, which includes hysteresis and eddy current losses, is typically modeled as a shunt resistance in parallel with the magnetizing inductance.

The parasitic inductances are primarily due to leakage flux, which does not contribute to energy transfer between the windings and can be modeled as leakage inductances in series with the windings. The magnetizing inductance accounts for the magnetizing current required

to establish the magnetic flux in the core and is represented as a shunt inductance across the primary winding.

The parasitic capacitances in a transformer arise from the potential differences between the winding layers and between the windings and the core or the tank. Although the value of these capacitances is not frequency-dependent, the impact of them becomes significant at high frequencies and can affect the voltage distribution across the windings and the high-frequency behavior of the transformer.

Two commonly used equivalent circuits to represent the non-ideal transformer are the T-model and the π -model. In the T-model, the leakage inductances of the primary and secondary windings are represented as series inductances on each side of the ideal transformer, and the magnetizing inductance is placed as a shunt branch across the primary side. In the π -model, the leakage inductances are divided equally and placed on both sides of the primary and secondary windings, while the magnetizing inductance remains as a shunt branch.

These equivalent circuits provide a practical method to analyze the transformer under non-ideal conditions, allowing the designer to evaluate the impact of parasitic parameters on the voltage regulation, efficiency, and dynamic performance of the transformer. In the following section, the T-model is considered for analyzing the transformer under non-ideal conditions. The T-model is selected in the following section because it provides a straightforward representation of the transformer parasitics while maintaining sufficient accuracy for analyzing the medium-frequency transformer behavior. The models described are shown in Figure. 5.1.

5.1.1. PARASITIC RESISTANCES

Parasitic resistances represent the ohmic losses in the transformer windings. These values vary with temperature and frequency, although the impact of frequency is dominant. At direct current (DC) or low frequencies, the resistance can be calculated using:

$$R_{dc} = \frac{\rho \cdot l}{A}, \quad (5.3)$$

where ρ is the electrical resistivity of the conductor material, l is the length of the conductor, and A is the cross-sectional area of the conductor.

As the frequency increases, the effective parasitic resistance increases due to the skin effect and proximity effect in the windings. These frequency-dependent effects lead to higher AC resistance compared to the DC value, resulting in increased power losses in the transformer at higher operating frequencies.

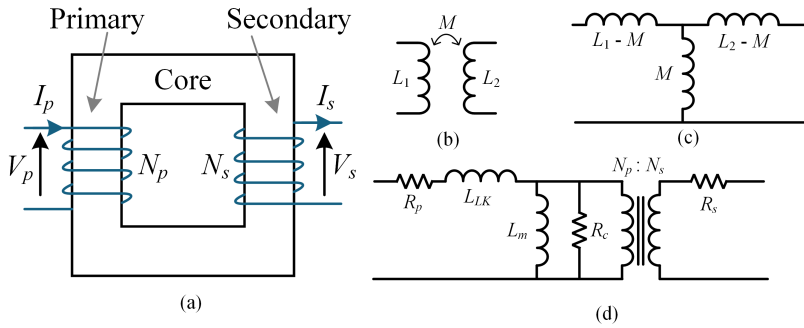


Figure 5.1: Illustration of transformer modeling approaches. (a) Physical structure of a transformer showing primary and secondary windings around the core. (b) Representation of coupled inductors with mutual inductance M . (c) Equivalent circuit of coupled inductors using the T-model. (d) T-model equivalent circuit of a transformer including parasitic elements: primary and secondary winding resistances (R_p , R_s), leakage inductance (L_{LK}), magnetizing inductance (L_m), and core loss resistance (R_c) in parallel with the magnetizing branch, together with the ideal transformer of turns ratio $N_p : N_s$.

The skin effect refers to the tendency of alternating current to concentrate near the surface of the conductor as the frequency increases, effectively reducing the current-carrying cross-sectional area and increasing the AC resistance.

The skin depth, δ , which indicates the depth at which the current density falls to $1/e$ of its value at the surface, is given by:

$$\delta = \sqrt{\frac{2\rho}{\mu\omega}}, \quad (5.4)$$

where ρ is the electrical resistivity of the conductor, μ is the magnetic permeability, and $\omega = 2\pi f$ is the angular frequency.

As the frequency increases, the skin depth decreases, leading to a higher effective resistance in the conductor compared to the DC resistance. This effect needs to be considered when calculating the parasitic resistances of the transformer windings operating at medium frequencies.

The proximity effect refers to the redistribution of current within a conductor caused by the magnetic fields of nearby conductors carrying alternating currents. This effect further increases the AC resistance of the windings, especially in transformer windings with closely spaced conductors. Unlike the skin effect, the proximity effect depends on the

geometry and arrangement of the conductors, as well as the phase and magnitude of the currents in adjacent conductors, making its explicit analytical derivation more complex. For this reason, proximity effect losses are often evaluated using empirical formulas, numerical methods, or finite element analysis in practical transformer design.

Dowell [2] developed a method to calculate the frequency-dependent AC resistance and leakage inductance in transformer windings, introducing correction factors to DC values by considering eddy current effects, including proximity effects, in multilayer and sectionalized windings. Ferreira [3] extended the analytical modeling by demonstrating the orthogonality between skin and proximity effects, enabling more accurate AC resistance calculations in multilayer windings, including round conductors. Kaymak et al. [4] compared various analytical methods for calculating AC resistance and leakage inductance in medium-frequency transformers against 2D FEM simulations and measurements, showing that modified Dowell-based methods provide reliable accuracy for common transformer geometries.

To improve the estimation of proximity effect losses in round conductors, Luo et al. [5] proposed a revised analytical method that refines Ferreira's formula by considering conductor interactions under uniform external fields, validated against 2D FEM results with good accuracy. Luo et al. [6–9] further enhanced the modeling by incorporating Bessel function-based formulations and the method of images, accounting for 2D and 3D edge effects and enabling accurate winding loss estimation across different porosity factors. Taher et al. [10] introduced the method of fundamental solutions (MFS) as a flexible numerical approach to analyze skin and proximity effects in conductors with arbitrary geometries, reducing computational complexity compared to volume-based FEM while maintaining accuracy. Kulkarni and Khaparde [11] provided a comprehensive review of stray loss evaluation methods in power transformers, emphasizing the need for accurate eddy current and proximity loss estimation in thermal and loss optimization. Wojda and Kazimierczuk [12] developed an approximate model for Litz-wire and multi-strand windings, adapting Dowell's method to include proximity effects within bundles, and derived optimal strand diameter and count for minimizing AC resistance in high-frequency designs.

These studies form a foundation for accurate parasitic resistance estimation in medium-frequency transformer design, enabling informed winding design choices under high-frequency operation, which is essential for thermal management and system efficiency in solid-state transformer applications.

5.1.2. PARASITIC CAPACITANCES

Under impulse conditions, transformers experience high-frequency voltage components, leading to capacitive effects not present under normal operation. Accurate capacitance calculations are necessary for circuit and traveling wave models to simulate this behavior [13].

The parasitic capacitances also influence the voltage distribution within the transformer during lightning and impulse events, whether occurring under test conditions or in actual operation. However, in the context of this study, the transformer is part of a solid-state transformer (SST) system, where such events are unlikely due to the presence of front-end protection and a modular multilevel converter (MMC). As a result, the medium-frequency transformer (MFT) will not be subjected to standardized lightning or impulse tests. This might suggest that the accurate modeling and design of parasitic capacitances is less critical. Nevertheless, parasitic capacitances remain important, as they affect the high-frequency behavior of the transformer, influence voltage distribution across the windings, and impact insulation coordination within the SST system.

Various models have been proposed for representing the parasitic capacitances in transformers [13–21]. The most common approach includes four main components: (1) the self-capacitance of each winding, (2) the inter-winding capacitance between primary and secondary windings, (3) the capacitance between the windings and the core, and (4) the capacitance between the windings and the transformer tank. Since both the core and the tank are conductive and typically grounded, the winding-to-core and winding-to-tank capacitances are modeled as capacitances to ground. The model is shown in Figure 5.2.

In most designs, the primary and secondary windings are arranged concentrically around the core. As a result, one winding—usually the inner winding—has a significant capacitance to the core, while the outer winding primarily exhibits capacitance to the tank. The presence of the inner winding, being conductive, acts as an electrostatic shield, reducing the direct capacitance between the outer winding and the core.

The self-capacitance of a winding consists of several components, depending on the winding type. One common element is the turn-to-turn capacitance (C_{tt}), which represents the capacitance between adjacent turns. For disk windings, such as those used in the transformer studied here, there is also disk-to-disk capacitance (C_{dd}), which arises between adjacent winding disks. Calculating these capacitances can be challenging due to the complexity of the winding geometry and the interactions between different parts of the structure.

A widely used approach for estimating these capacitances is the *energy method*, which is frequently applied in the literature. This method involves computing the electrostatic energy stored in the winding configuration and relating it to the equivalent capacitance

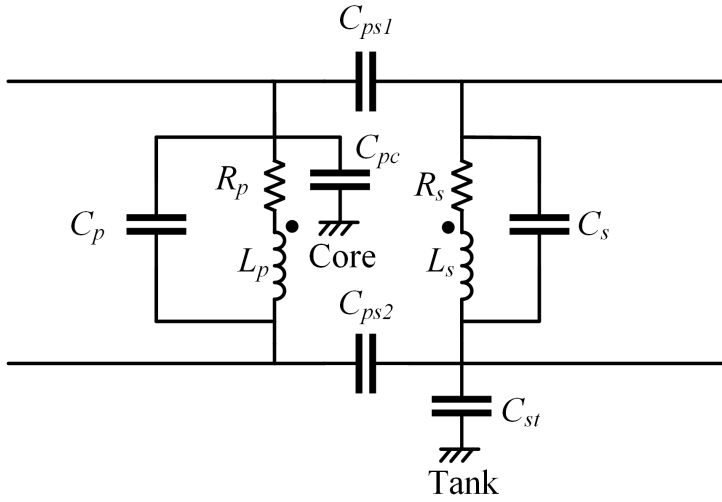


Figure 5.2: Equivalent circuit model of a transformer including parasitic capacitances and winding resistances.

5

using fundamental electromagnetic relationships. It provides a practical means of evaluating complex capacitive networks in transformer windings, especially when analytical expressions are not easily derived.

The energy method is commonly used for calculating parasitic capacitances in transformer windings, particularly when dealing with complex geometries such as disk windings. In this method, a linear voltage distribution is assumed across the winding section. The electrostatic energy stored in the system is calculated, and the capacitance is derived using the following relation:

$$C = \frac{2W_e}{V^2} \quad (5.5)$$

where C is the capacitance, W_e is the electrostatic energy, and V is the applied voltage across the winding segment. This method is effective in capturing the capacitive coupling between winding sections, winding to core, and winding to tank, especially when used in combination with finite element simulations [20]. The total self-capacitance of the winding can be expressed as follows [13, 21, 22]:

$$C_{\text{stray}} = \frac{4}{3}C_{\text{dd}} + \frac{N_t - 1}{N_t^2}C_{\text{tt}} \quad (5.6)$$

where C_{dd} is the disk-to-disk capacitance, C_{tt} is the turn-to-turn capacitance, and N_t is the total number of turns. These capacitances are computed using the following analytical expressions:

$$C_{tt} = \frac{\varepsilon_0 \varepsilon_r 2\pi R_{ave}(h + d)}{d_{tt}} \quad (5.7)$$

$$C_{dd} = \frac{\varepsilon_0 \varepsilon_r \pi (R_{out}^2 - R_{in}^2)}{d_{dd}} \quad (5.8)$$

where R_{ave} is the average radius of the turn, R_{out} is the radius of outer turn, R_{in} is the radius of inner turn, d_{tt} is the distance between two adjacent turns, and d_{dd} is the distance between two adjacent disks. The relative permittivity ε_r of the insulation system is calculated using a weighted combination of the constituent layers:

$$\frac{1}{\varepsilon_r} = \frac{1}{\varepsilon_1} \frac{d_1}{d_{tot}} + \frac{1}{\varepsilon_2} \frac{d_2}{d_{tot}} + \frac{1}{\varepsilon_3} \frac{l_1}{l_{tot}} + \frac{1}{\varepsilon_4} \frac{l_2}{l_{tot}} \quad (5.9)$$

These equations allow analytical estimation of parasitic capacitances in disk-wound transformer windings, which can be further refined through FEM-based methods.

The analytical estimation of turn-to-turn and disk-to-disk capacitances can be performed using the simplified models and geometric approximations illustrated in Figure 5.3.

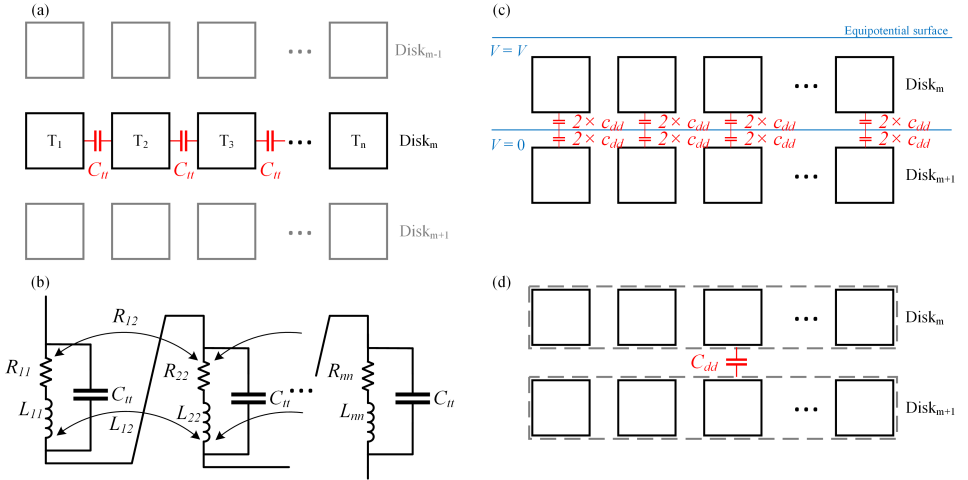


Figure 5.3: Modeling approach for parasitic capacitances in disk-type transformer windings: (a) turn-to-turn capacitance (C_{tt}) structure within a disk, (b) equivalent RLC network incorporating C_{tt} , (c) distribution of disk-to-disk capacitances (C_{dd}) between adjacent disks considering equipotential layers, and (d) FEM-based approximation of C_{dd} between disk layers.

5.2. UNBALANCING ISSUE IN A HIGH-CURRENT WINDING AND MITIGATION

Solid-State Transformers (SSTs) are increasingly employed in high-power applications, they are commonly connected to medium-voltage grids and tasked with supplying relatively low-voltage loads such as green-hydrogen electrolyzers [23, 24]. This operational scenario necessitates handling high currents on the secondary side. Meanwhile, SSTs typically contain medium frequency transformers. Handling high currents at such a frequency requires can be complex.

To manage high current in transformer windings, a number of current-handling techniques have been employed in both low- and medium-frequency applications. These include the use of continuously transposed conductors (CTCs) and parallel winding paths, which help distribute current and reduce skin and proximity effects. However, these methods show limitations at medium frequencies, where proximity effects become dominant and winding geometry plays a larger role [25–27].

CTCs can mitigate the skin effect significantly; nevertheless, the impact of the core and the proximity effect become dominant when the applied frequency is higher. Implementing a number of parallel paths addresses practical issues when big conductors are needed to handle the high current. However, the proximity effect is again problematic.

While Continuously Transposed Conductors (CTCs) are effective in low-frequency transformer applications, their practicality reduces at medium frequencies. As frequency increases, skin depth decreases, requiring smaller strands to minimize AC resistance. However, high-current applications, simultaneously, demand a large conductor cross-section, resulting in a complex, multi-strand structure that is difficult to implement. Moreover, the use of multiple parallel conductors, whether CTCs or Litz wires, intensifies the proximity effect, particularly in the presence of a magnetic core. This leads to significant current imbalance and localized heating, making CTCs and similar techniques less suitable for medium-frequency, high-power transformer designs.

In low-frequency, high-current transformers, such as arc furnace transformers, additional techniques such as geometric modifications (e.g., varying disk diameters, winding height, or spacing) are employed to improve current sharing. While effective at 50/60 Hz, these geometrical methods become impractical at higher frequencies due to the need for significant variations and increased construction complexity. In addition, SSTs face additional complexities since the operation under high- or medium-frequency excitation leads to enhanced proximity and skin effects, exacerbating high current issues [17]. While litz wires are a viable solution for low-voltage power electronics applications to mitigate these effects, they are impractical for medium-voltage,

medium-frequency SSTs [28].

The literature on SSTs and high-current handling techniques is extensive. Prior research has explored various aspects of SST design, including their efficiency, thermal management, and electromagnetic performance [23, 29–36]. Studies on arc furnace transformers have detailed the benefits of using CTCs and parallel paths to manage high currents. However, there is a gap in the literature concerning the direct application of these techniques to medium-voltage, medium-frequency SSTs.

The present study introduces a novel method to mitigate imbalance current sharing issues by employing a number of series connected compensation inductors for high-current windings including parallel disks. The proposed method introduces a modular, scalable compensation scheme using toroidal cores, enabling disk-level current balancing without altering the transformer geometry. The approach further includes tunable inductance control via auxiliary windings or permanent magnets, which allows real-time or design-phase optimization. These innovations offer a unique combination of frequency adaptability, practical integration, and experimental validation, distinguishing the method from existing techniques.

The proposed method is compared to the existing approaches in Table 5.1. While prior studies have explored current balancing through geometric variations, transposed conductors, or winding arrangements, these approaches often face limitations in medium-frequency transformer applications due to increased losses, constructional complexity, and low tunability. Notably, existing compensation techniques do not address current unbalancing at the disk-level with precision or flexibility suitable for MV SSTs.

Table 5.1: Method Effectiveness at Medium-Frequency for MV SST

Method		Effectiveness at MF	Complexity	Tunability	unbalancing reduction	Remarks
Parallel without compensation	Disks Compensation [37]	Low to Moderate	Low	-	-	Simple but leads to uneven current due to inductance differences. End-coil current and mid-coil current have >80 % difference..
Geometrical Adjustments	Ad-justments (e.g., disk height) [38]	Low	High	Fixed	13.6 %	Not effective at MF due to constructional complexity.
Compensation Toroids	(This Work)	High	Low to Moderate	High	>98.4 %	Easily tunable via magnets or auxiliary windings; validated experimentally.

5.2.1. UNBALANCING IN CURRENT SHARING

In a transformer feeding a high-current load, any current unbalancing within the winding between the parallel paths is not desired and must be mitigated as much as possible. Current unbalancing leads to extra losses and extra heat, especially at the hotspots of the transformer, raises the need for more complicated cooling systems, and generally increases the total cost of owning the transformer. The unbalancing has more severe consequences in high- or medium-frequency transformers.

There are several methods to deal with the high current in the secondary side of a step-down transformer. In most cases, having several parallel paths for the current in the secondary is essential. These parallel paths can be along the conductor like litz wires and CTCs with several strands and transposition to avoid extra losses. Alternatively, the parallel paths can be within the same disk in the winding. However, having more parallel paths, like parallel disks for high-current applications, is essential. For all of the techniques, current balancing is vital since having poor current sharing results in extra loss and rising temperatures, especially near the hotspots of the winding. This phenomenon is more severe in medium- or high-frequency transformers for two main reasons. Firstly, the higher the frequency, the higher the proximity and skin effect. Secondly, the unbalancing occurs because the difference in the inductances of parallel disks is more significant when the frequency is much higher than 50 Hz.

UNBALANCING DUE TO DIFFERENT INDUCTANCES

In the 2D core of Figure 5.4, the three parallel coils consist of only one turn each. The relative permeability of the core is significantly greater than 1 but still finite. The permeability of the core can be calculated as follows:

$$\mu_c = \mu_0 \cdot \mu_{rc} \quad (5.10)$$

where $\mu_0 = 4\pi \times 10^{-7}$ and μ_{rc} is the relative permeability of the core. The electrical conductivity of the coils is significantly greater than 1. The outer borders of the core are assumed to be magnetic insulation ($A_x = A_y = A_z = 0$). The inductance of each winding can be calculated as follows:

$$[V] = j\omega[L] \cdot [I] \quad (5.11)$$

where ω is the angular frequency. $[V]$, $[L]$, and $[I]$ are the matrices of voltage, inductance, and current, respectively. For the three-coil

problem, Equation 5.11 can be written as follows:

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \end{bmatrix} = j\omega \begin{bmatrix} L_{11} & L_{12} & L_{13} \\ L_{21} & L_{22} & L_{23} \\ L_{31} & L_{32} & L_{33} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix} \quad (5.12)$$

The elements of the inductance matrix inductance $[L]$ consist of two parts as follows:

$$[L] = [L_c + L_a] \quad (5.13)$$

According to 5.13, the inductances are made of two parts: one corresponding to the magnetic core (L_c) and one corresponding to the air (L_a). Since the relative permeability of the core is significant, $L_c \gg L_a$.

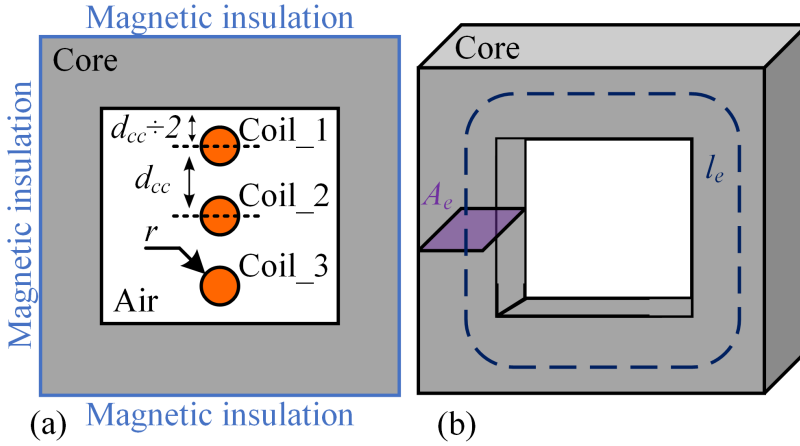


Figure 5.4: (a) The simple 2D problem with three parallel coils inside the core window, (b) Three-dimensional view of a magnetic core illustrating the effective magnetic path length l_e and the effective cross-sectional area A_e .

L_c can be calculated using the equivalent magnetic circuit as shown in Figure 5.5 as follows:

$$L_c = \frac{N^2}{R_c} = \frac{N^2 \mu_0 \mu_r A_e}{l_e} \quad (5.14)$$

where R_c is the equivalent reluctance of the core, A_e is the effective cross-sectional area of the core, and l_e is the effective length of the core, as shown in Figure 5.4.b.

Calculating L_a is not straightforward. There are several studies investigating the inductance in the core window, such as [39]. Moreover,

both L_c and L_a can be calculated using the finite element method (FEM). In the present study, L_c and L_a are obtained through FEM simulation using COMSOL. The relationship between inductance and magnetic energy is given by:

$$E = \frac{1}{2}LI^2 \quad (5.15)$$

where I is the current of the coil. L_c and L_a can be calculated as follows:

$$L_c = \frac{2}{I^2} \iiint_{\text{core}} W_{\text{mag}} dv \quad (5.16)$$

$$L_a = \frac{2}{I^2} \iiint_{\text{window}} W_{\text{mag}} dv \quad (5.17)$$

where W_{mag} is the time-averaged magnetic energy density, and the integrals are evaluated over the core and the window regions for L_c and L_a , respectively. The value of L_c remains the same, and the only variation appears in L_a among them. Therefore, the elements of the inductance matrix have values close to each other, and the matrix can be approximated as:

$$L = \begin{bmatrix} L_c + L_{a11} & L_c + L_{a12} & L_c + L_{a13} \\ L_c + L_{a21} & L_c + L_{a22} & L_c + L_{a23} \\ L_c + L_{a31} & L_c + L_{a32} & L_c + L_{a33} \end{bmatrix} \quad (5.18)$$

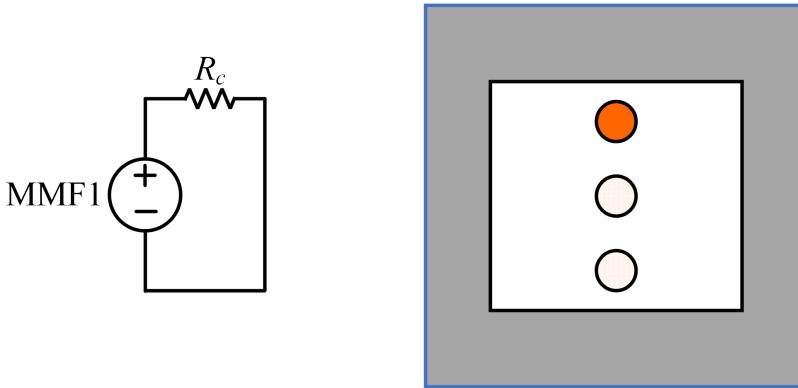


Figure 5.5: The equivalent magnetic circuit when Coil_1 is excited by a current and other coils are left open circuit ($I = 0$).

The condition number measures how much the output value of a function can change for a small change in the input. When the elements

of a matrix like $[L]$ are close to each other, the matrix might have a high condition number, especially if the differences between the elements make the matrix nearly singular (i.e., its determinant is close to zero). This implies that $[L]$ is ill-conditioned, and as a result, the inverse matrix $[L]^{-1}$ can contain elements with significant differences, even though the elements of $[L]$ differ only slightly.

In a winding with several parallel paths, the voltage across the parallel disks is fixed because the ends of the windings are rigidly connected to high-current busbars. Assuming the voltages are equal, the currents can be calculated from Equation (5.11) as follows:

$$[I] = \frac{1}{j\omega} [L]^{-1} [V] \quad (5.19)$$

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Therefore, the currents can be varied significantly even though the inductances are almost equal. Analysis based on [21] and FEM proves that in winding around a magnetic core without an air gap consisting of several parallel coils (disks), the inductance of the upper and lower coils is less than the inductance of the inner coil. Therefore, the upper and lower coils will carry more current than the inner coils. This situation is depicted in Figure 5.6.

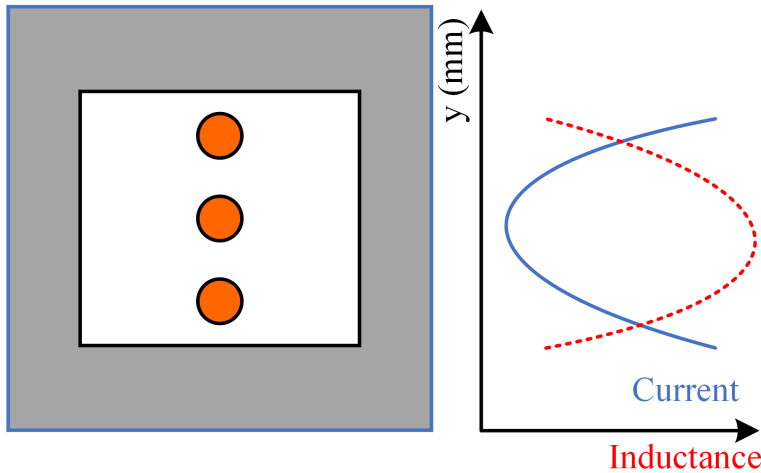


Figure 5.6: The current sharing and inductance comparison of parallel coils (disks) of a winding around a non-gapped magnetic core. The upper and lower coils have less inductance and higher current than the inner coil.

THE IMPACT OF THE AIR GAP

The equivalent magnetic circuit, when considering the presence of an air gap, is modified by introducing an additional reluctance r_g in series with the core reluctance [17, 39, 40]. The fringing effect, changes the field in the areas near the gap, can be modeled as a parallel branch consisting of a reluctance in series connection to an anti-MMF source in the equivalent magnetic circuit, as shown in Figure 5.7. The coils nearer to the gap will be more affected by the fringing effect. This impact is shown in Figure 5.8. The fringing effect leads to a non-uniform magnetic field distribution, causing variations in inductance across different coil positions. Coils closer to the gap experience reduced inductance due to the lower effective permeability in the fringing region, which increases the local magnetic reluctance. Since current follows the path of least impedance, lower inductance results in higher current flow in these coils. A theoretical justification for this can be found in magnetic circuit analysis and field distribution studies in gapped-core transformers [39, 40].

The impact of the air gap is illustrated in Figure 5.9 using COMSOL simulation based on FEM. In Figure 5.9.a, the core does not have any gap. Therefore, the flux uniformly flows the two legs (vertical parts) and two yokes (horizontal parts) of the core. In addition, the magnetic field inside the window is more intense near the upper and lower coils in comparison to the middle one. The higher field intensity, the lesser the inductance and lesser the impedance consequently. The lower impedance of two outermost coils leads to having higher current density through those coils. As soon as one of the legs of the core has a air gap, the situation differs from Figure 5.9.a, which is depicted in Figure 5.9.b. This also changes the magnetic field inside the window and has impact on the impedance of the coils. Nevertheless, for most of the real-world magnetic cores, both legs have air gaps as illustrated in Figure 5.9.c, where the magnetic field intensity is higher near the middle coil which is located near the air gap. Again, the lesser the inductance the higher the current density in the coil.

5.2.2. CURRENT UNBALANCING MITIGATION

Although there are other possibilities to achieve current balancing such as performing distributed air gaps through the magnetic core, the novel approach of the present study is to use series compensation inductors to achieve current balancing.

The impedance of a transformer winding is defined as follows:

$$Z = R_{\text{coil}} + j\omega L_{\text{coil}} \quad (5.20)$$

where $R_{\text{coil}} = R_{\text{coil,dc}} + R_{\text{coil,ac}}$ and L_{coil} is the inductance of the coil.

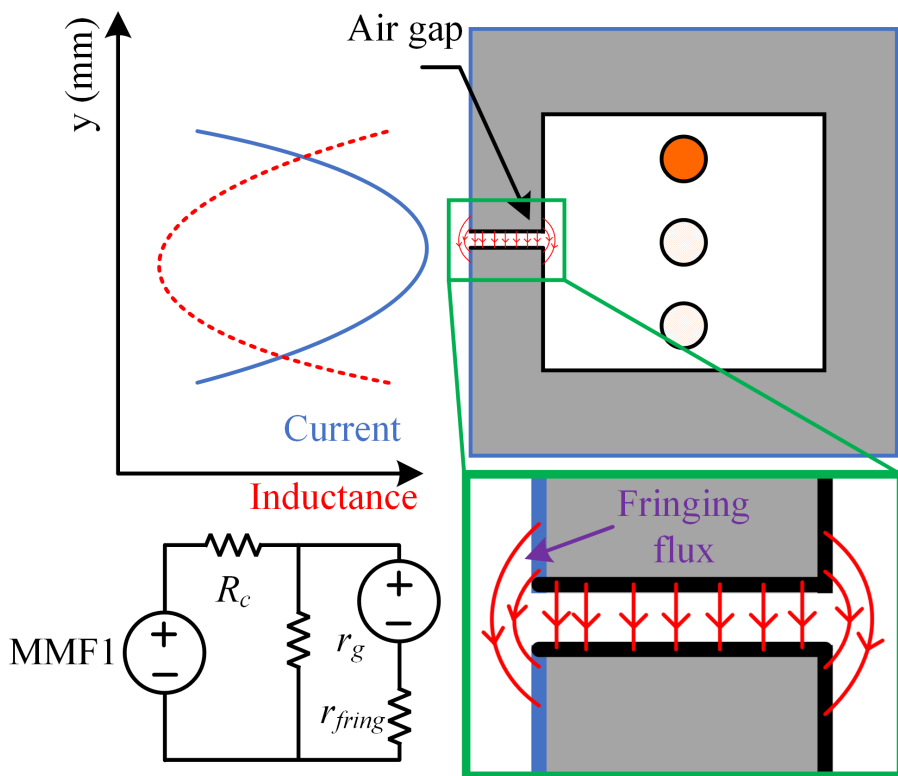


Figure 5.7: The equivalent magnetic circuit of a gapped magnetic core considering the fringing effect. The upper and lower coils have higher inductance and lesser current than the inner coil.

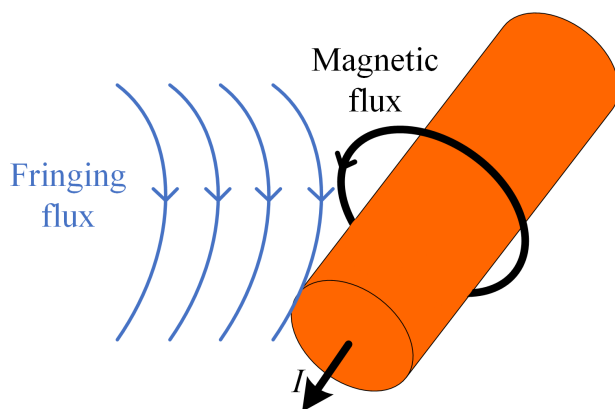


Figure 5.8: Fringing effect near a current-carrying wire (the coil).

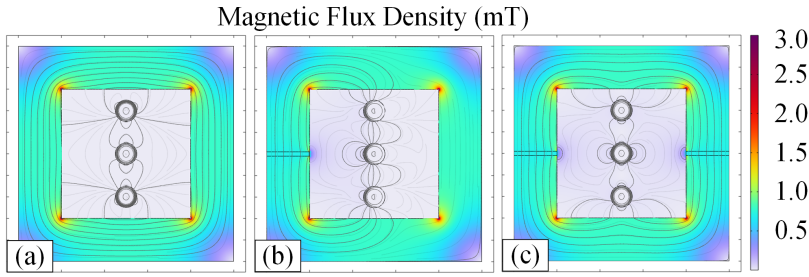


Figure 5.9: FEM simulation results of the impact of air gap in the core: a) without gap, b) Air gap on one leg, and c) air gaps on both legs. The main difference is in the field distribution in the window.

For a transformer, the inductive part is dominant since the winding is designed to carry current with minimal losses. In the proposed winding structure, several coils (disks) are connected in parallel. All of the coils can be considered as coupled inductors, as they are wound around the same magnetic core. Since they are connected in parallel and magnetically coupled with aiding flux linkage, the system can be modeled as parallel-connected aiding coupled inductors (Figure 5.10.a).

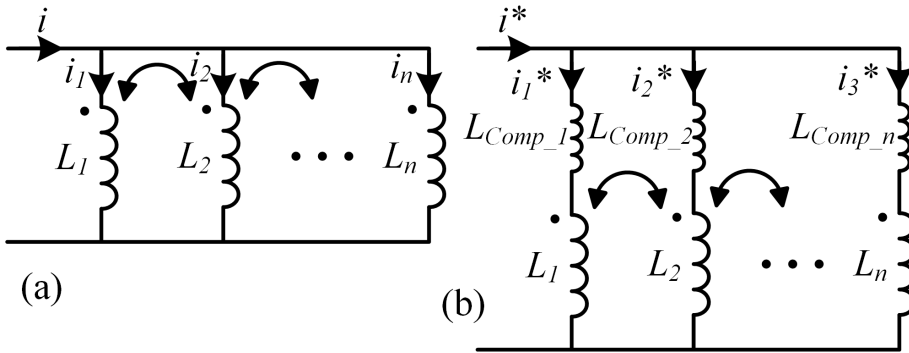


Figure 5.10: The parallel coils as parallel connected coupled inductors: (a) without compensation, (b) with the compensation toroids.

The currents can be derived from Equation (5.19). Adding a compensation toroid to each coil introduces an additional inductance specific to each coil, which does not share the same magnetic path with the others and, consequently, is not magnetically coupled with them. Therefore, the inductance matrix can be rewritten as:

$$[L^*] = \begin{pmatrix} L_c + L_{a11} + L_{comp1} & L_c + L_{a12} & \cdots & L_c + L_{a1n} \\ \vdots & \ddots & \ddots & \vdots \\ L_c + L_{an1} & L_c + L_{a2n} & \cdots & L_c + L_{ann} + L_{compn} \end{pmatrix} \quad (5.21)$$

where L_{comp} is the inductance of the compensation toroid added to the coil. The new inductance matrix $[L^*]$ determines the updated current distribution as follows:

$$[I^*] = \frac{1}{j\omega} [V] \cdot [L^*]^{-1} \quad (5.22)$$

If the value of the compensation inductances selected suitable, the currents will be balanced. It is noteworthy that adding more than needed compensation inductance leads to “overcompensation” which can also rise unbalancing.

The idea of using toroids to compensate for the differences between the inductances of the parallel disks is depicted in Figure 5.11. The method is described as follows. At the first turn of each coil, a toroidal magnetic core is placed to add the desired inductance to compensate for the differences between the inductances of parallel-connected coils. For the power transformers that do not have air gaps in the middle of the leg(s), the introduced compensation inductance will be higher for the top and bottom windings since these windings will carry more current without compensation. For the transformers made of cut cores similar to the prototype of the present study, the introduced inductance of the middle coil is selected higher since the middle coil is closer to the air gap of the core, has lesser inductance, and consequently carries more current. The FEM simulation of the prototype is shown in Figure 5.12. The inner coil carries more current since it has lesser inductance because of the presence of the gap in the magnetic core. Before applying the compensation toroids, the inner coil carried significantly more current in comparison to the other windings as shown in Figure 5.12.b. However, after applying the compensation toroids, the current densities of the different coils become more similar to each other (Figure 5.12.c).

The value of the current that each coil carries before applying the proposed compensation inductors shows significant imbalance. These values are shown in the Table 5.2. Nevertheless, after applying the proposed compensation inductors, the value of the currents will be balanced significantly as shown in Figure 5.12 and Table 5.2.

The proposed method also has significant impact on the loss amount and distribution. The value of the loss before and after applying the proposed method can be calculated as follows:

Table 5.2: The Simulation Results: The Value of the Added Compensation Inductances and the Balancing Summary at 1 kHz

Coil No.	Added Compensation Inductance (μH)	Equivalent Inductance (μH)	Current (mArms)	
			Before	After
1	0.00	416.23	92.37	98.00
2	2.03	413.05	88.36	97.72
3	5.21	411.02	151.17	97.50
4	2.03	413.05	88.36	97.72
5	0.00	416.23	92.37	98.00

$$\text{LOSS}_{\text{NC}} = \sum_{n=1}^N R_n \times I_n^2 \tag{5.23}$$

$$\text{LOSS}_{\text{C}} = \sum_{n=1}^N R_n^* \times I_n^{*2} \tag{5.24}$$

where N is the number of parallel coils, R_n is the AC resistance of the n^{th} coil before compensation, and R_n^* is the AC resistance of the n^{th} coil after compensation. The losses are compared in Figure 5.13 for the prototype at 1 kHz, illustrating that applying the compensation method reduces the total loss by 28%. Nevertheless, it is not the only positive impact of the proposed method. For the compensated case the loss in the middle point, which is most likely the hotspot of the winding, is less, resulting in a more balanced heat distribution and better overall thermal performance of the winding.

It is important to note that while the compensation toroids introduce additional inductance to each parallel winding, their values are carefully selected to be significantly smaller than the transformer’s leakage inductance, typically in the range of 0.1% to 1%. This ensures that the added inductance has a negligible effect on the overall transformer behavior and does not interfere with the converter’s performance in medium-frequency power conversion applications.

The compensation toroids are physically small and operate at relatively low flux densities. Given their limited core volume and the use of low-loss ferrite material (e.g., 3E25), the additional core losses introduced are negligible in comparison to the winding and transformer core losses. Therefore, the method provides a net improvement in efficiency by mitigating current unbalance and reducing associated localized losses.

The selection of the compensation toroid’s parameters, such as core geometry, material, and number of turns, should be based on the required compensation inductance L_{comp} , which in turn depends on the inductance mismatch ΔL between the parallel winding disks. A practical

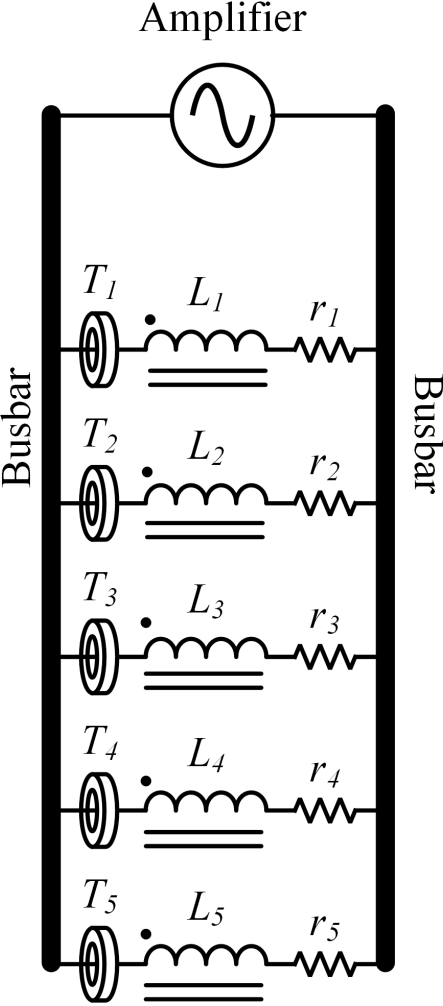


Figure 5.11: Proposed method for mitigating current unbalancing in the parallel coils using compensation toroids.

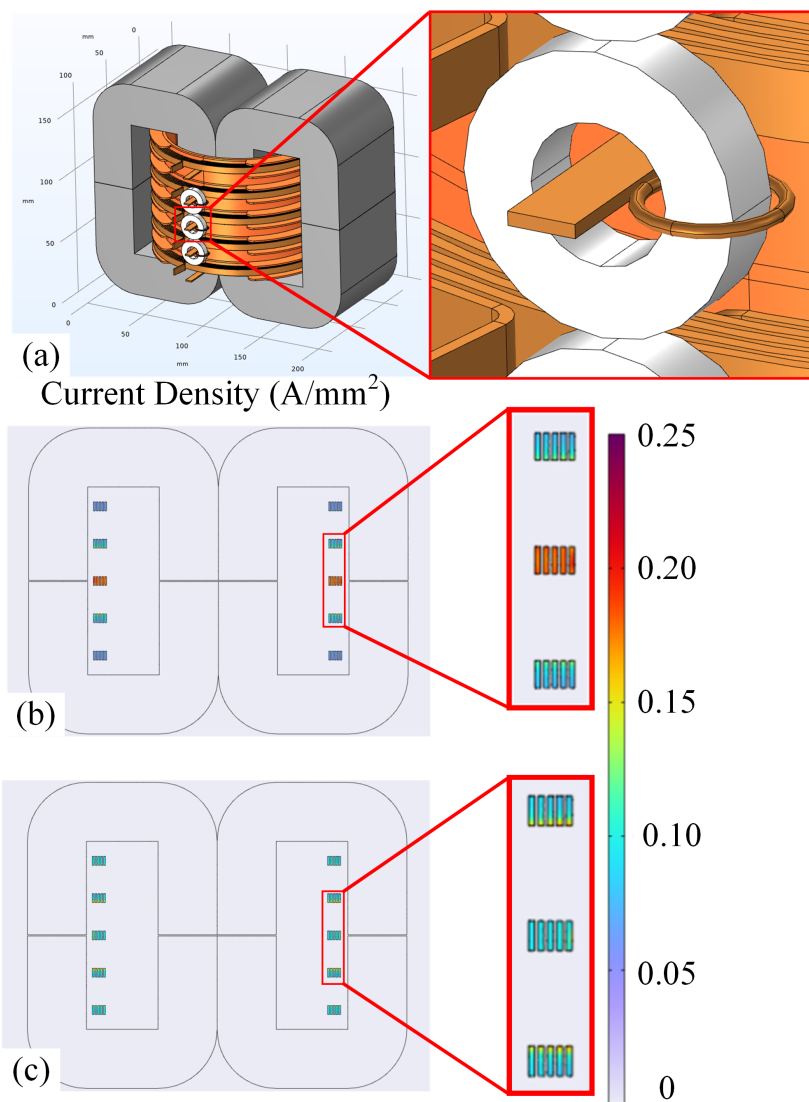


Figure 5.12: The FEM simulation for the prototype: (a) the 3D model and the compensation toroid, (b) the current density (A/mm^2) over parallel-connected coils around the core before applying the compensation toroids, and (c) the impact of applying compensation toroids.

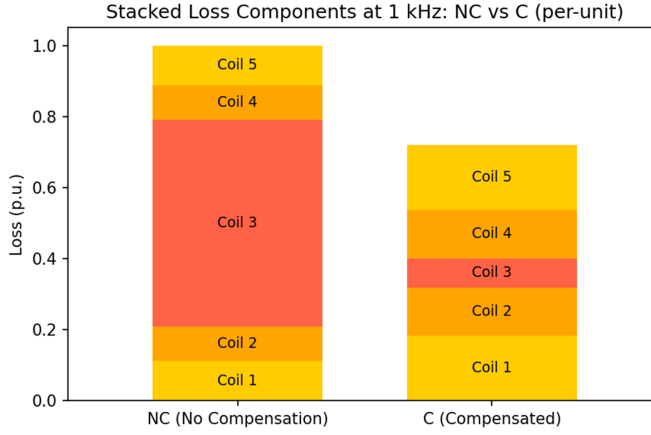


Figure 5.13: Comparison of the losses of each coil before and after compensation.

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rule of thumb is to keep $L_{comp} \ll L_{leakage}$ to avoid interference with the overall transformer behavior. The required L_{comp} can be calculated from:

$$L_{comp} = L_{ref} - L_i \quad (5.25)$$

where L_{ref} is the highest inductance among all parallel paths, and L_i is the inductance of the i^{th} disk.

The value of the compensation inductance affects the current sharing; therefore, it is essential to fine-tune this value. For fine-tuning, an auxiliary winding is employed to control the inductance of the compensation toroid. This method selected for the experimental validation since it is excellent for controlling the current sharing of the winding, especially in overcoming hotspot temperature rise issues.

5.2.3. EXPERIMENTAL VALIDATION

A prototype is made to prove the reasoning behind and analyze the unbalancing of the current sharing and mitigation methods. The winding is made of five parallel disks, each with five turns. The winding is made around a 3D-printed Polylactic Acid (PLA) bobbin. The coils' turns are made of a 1×5 mm rectangular enameled conductor to ensure the fixed position of each turn in the disk. All disks are connected to high-current busbars for a solid and reliable parallel connection.

The magnetic core comprises two AMCC1000 amorphous cores (four pieces) with a shell-type core structure. The core is made of several layers of Metglas alloy, which provides a good lamination feature.

However, the core is a cut-core, and there is a gap between the upper and lower parts, affecting the current sharing of the parallel disks. The prototype is shown in Figure 5.14.

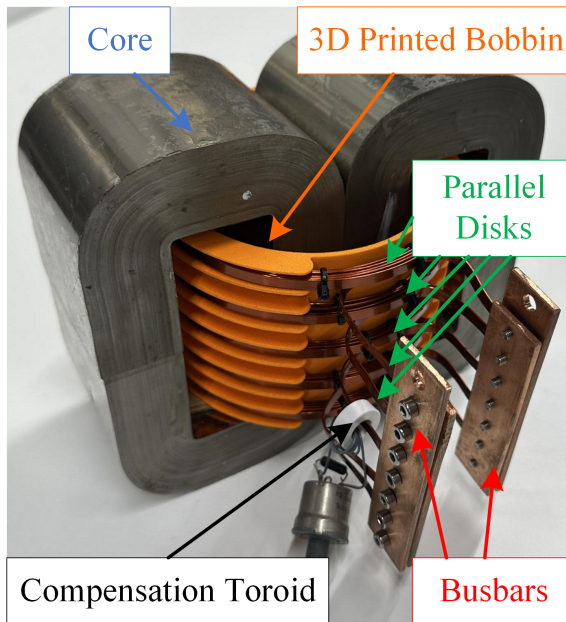


Figure 5.14: The prototype. Key components include the magnetic core, 3D-printed bobbin, parallel disks, busbars, and compensation toroid.

TEST SETUP

The measuring setup consists of seven different parts as follows:

1. Signal generator (Keysight 33500B)
2. Power supply (Delta Elektronika 15 V, 20 A)
3. Amplifier (Sony Xplod 500 W)
4. Oscilloscope (Tektronix 2 Series MSO 4x 200MHz 2.5GSPS)
5. Differential voltage probes (Picotech TA057)
6. Current probes (Keysight N2783B)
7. The prototype winding and core

The test table is shown in Figure 5.15 and the block diagram of the test setup is shown in Figure 5.16. The test is done with 1 kHz frequency. The maximum current is $5.43 A_{rms}$.

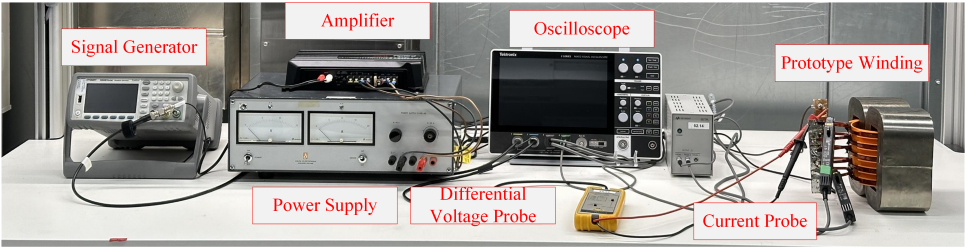


Figure 5.15: The test table.

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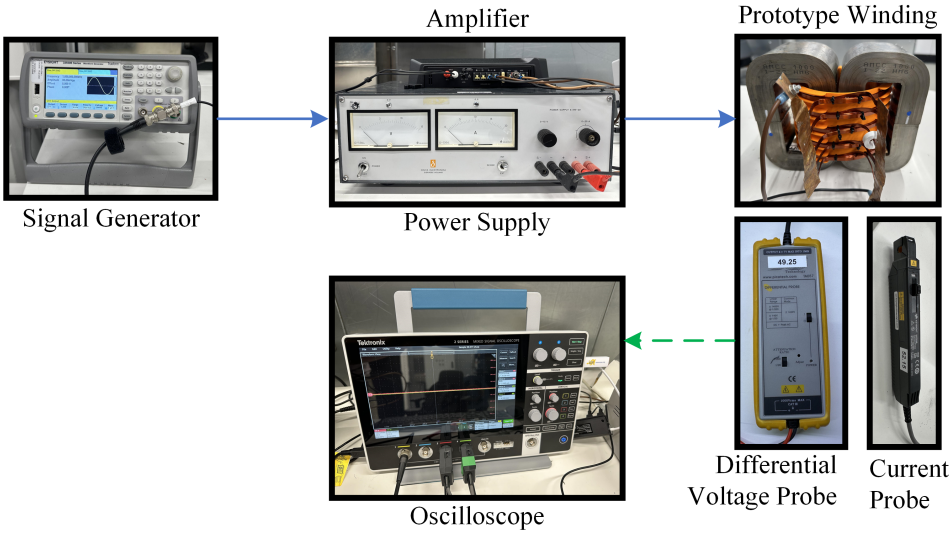


Figure 5.16: The test setup block diagram.

COMPENSATION TOROID FABRICATION AND FINE-TUNING

The compensation toroid using the auxiliary fine-tuning method is shown in Figure 5.18. The compensating toroid must be designed based on the expected current carried by each coil. However, the variation in current between the parallel paths primarily originates from differences in their inductances.

The toroid core used is a TN19/11/10 ferrite core made of 3E25 material, with an A_L value of $6420 \text{ nH}/N^2$, where N here is the number of turns. Ferrite was selected due to its low core losses and high electrical resistivity, making it a suitable and efficient material for medium-frequency applications where eddy current and magnetic losses must be minimized.

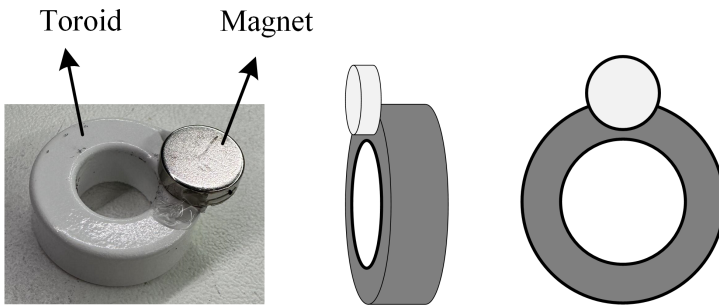


Figure 5.17: The compensation toroid using a permanent magnet (NdFeB) as a fine-tuning method. The magnetic material alters the local flux density, enabling minor adjustments to the inductance by varying the position and overlap of the magnet on the core.

5.2.4. EXPERIMENTAL RESULTS

The experimental results are as follows. The values of the inductances of the coils are as follows:

$$L_{\text{coils}} = [416.23 \quad 413.05 \quad 411.02 \quad 413.05 \quad 416.23] \mu\text{H} \quad (5.26)$$

The value of the inductances of the coils should be as equal as possible to guarantee a good current-sharing. Therefore, the values of compensation inductances must be as follows:

$$L_{\text{comp}} = [0 \quad 2.03 \quad 5.21 \quad 2.03 \quad 0] \mu\text{H} \quad (5.27)$$

The inductance of the compensation toroid can be adjusted from $0.46 \mu\text{H}$ (when the auxiliary winding is short-circuited) to $6.4 \mu\text{H}$

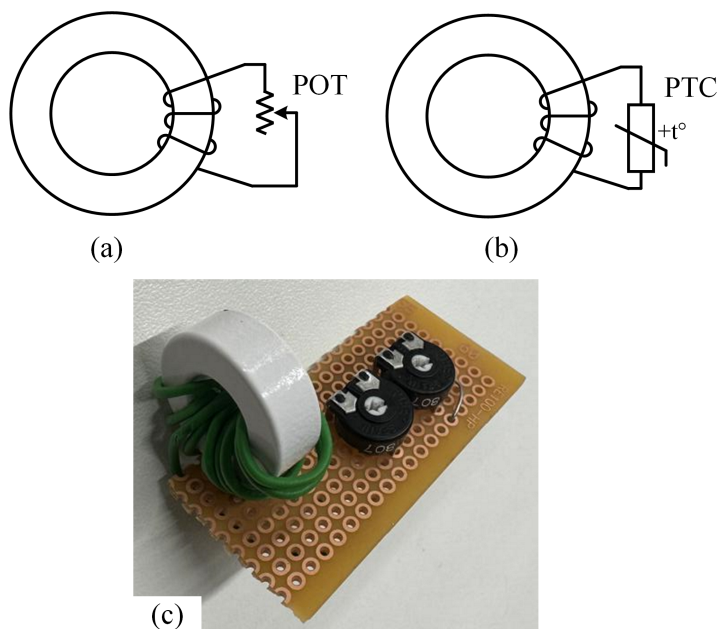


Figure 5.18: The compensation toroids using auxiliary winding as a fine-tuning method: (a) using a potentiometer, (b) using a positive-temperature-coefficient (PTC) thermistor, and (c) the prototype compensation toroid.

(when the auxiliary winding is open-circuited). For coils #1 and #5, no compensation toroid is required. For coils #2 and #4, the potentiometer in the auxiliary winding is adjusted to achieve a compensation inductance of $2.03\ \mu\text{H}$, while for coil #3, it is set to $5.21\ \mu\text{H}$. The flowchart of the presented compensation method is shown in Figure 5.19.

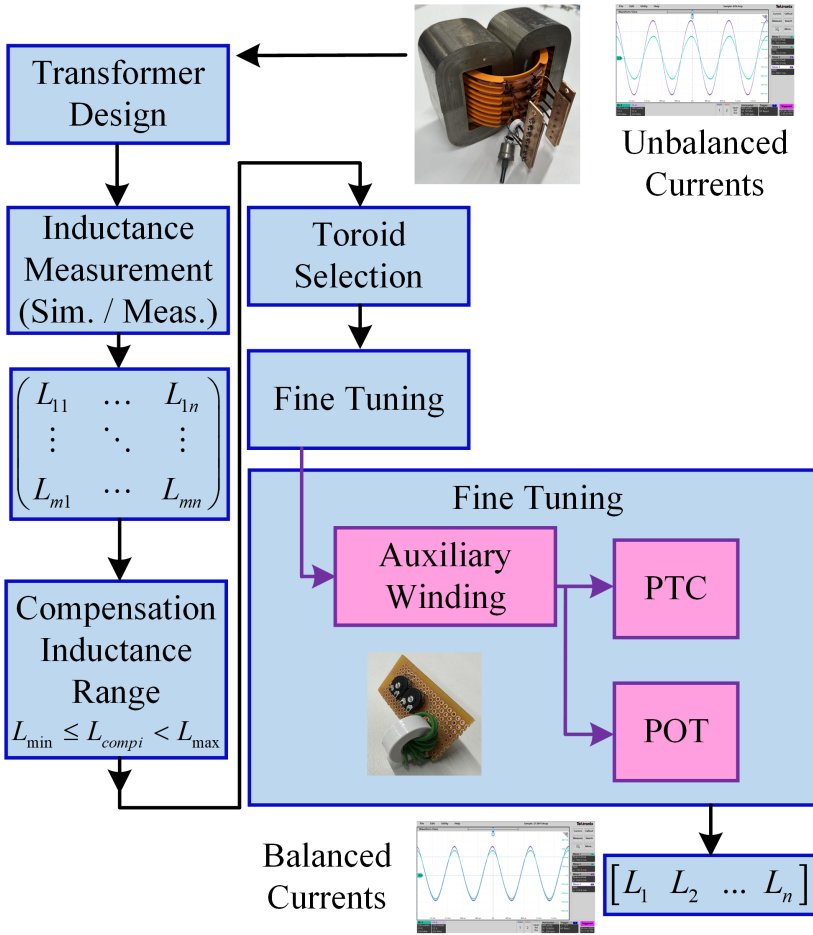


Figure 5.19: The flowchart of the presented method. The designed transformer might face unbalancing in current distribution. The compensation inductances are designed and fine-tuned to achieve current balancing.

The difference in current between the upper disk (D_1) and the middle disk (D_3) before applying the compensation toroid is illustrated in Figure 5.20. The figure also presents the trend after compensation,

highlighting the improved current sharing and achieved balance between disks #1 and #3.

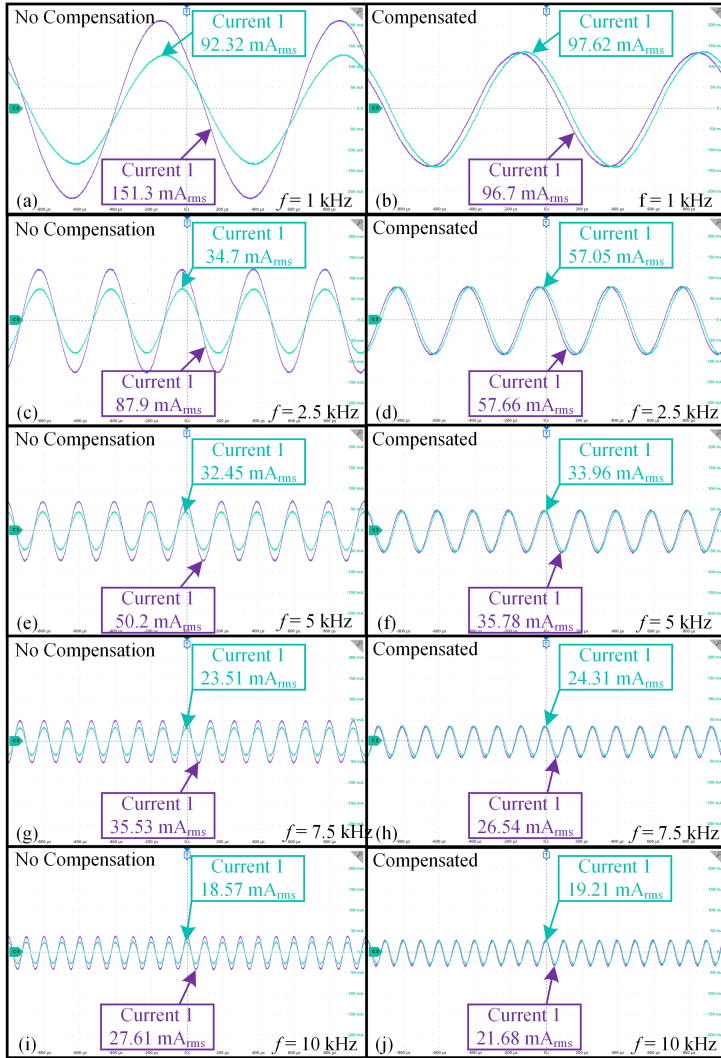


Figure 5.20: Experimental results before and after applying compensation toroids at 1 kHz, 2.5 kHz, 5 kHz, 7.5 kHz, and 10 kHz. Each row compares the unbalanced and compensated current waveforms for disks #1 and #3, demonstrating the effectiveness of the compensation method in improving current sharing across frequencies.

Although the experimental validation in this work has been conducted

at 1, 2.5, 5, 7.5 and 10 kHz, the proposed current balancing method is not limited to these frequencies. The selected frequency range is based on a system-level optimization for large-scale green hydrogen production using solid-state transformers (SSTs). In this application, high-current medium-frequency transformers (MFTs) are needed to supply large electrolyzers from medium-voltage offshore wind power. Due to the high secondary current levels, the transformer design shares similarities with arc furnace transformers, requiring careful trade-offs between frequency, size, and thermal management. Operation at higher frequencies leads to reduced transformer size but increases proximity and skin effects, which in turn require more complex cooling systems. Figure 5.21 shows that while the absolute difference between winding inductances decreases with frequency, the percentage mismatch becomes more significant. This variation necessitates frequency-specific compensation. As demonstrated in Figure 5.22, the compensation elements can be tuned to maintain effective current sharing across the frequency range. The compensated case maintains minimal imbalance across frequencies, demonstrating the effectiveness of the tuning method. These results confirm that the proposed compensation method remains effective under higher-frequency operation, provided that proper tuning is applied during the design phase.

5.3. CONCLUSIONS AND KEY FINDINGS

This chapter investigated the origin, modeling, and mitigation of parasitic effects in medium-frequency transformers (MFTs), with a particular focus on current unbalancing in high-current parallel disk windings used in solid-state transformer (SST) applications.

First, the physical mechanisms of parasitic resistances, inductances, and capacitances were reviewed, highlighting their increasing influence at medium frequencies due to skin and proximity effects as well as capacitive voltage redistribution. Analytical expressions and energy-based FEM methods were used to estimate these parameters, providing a consistent framework for integrating electromagnetic effects into circuit-level transformer models.

Second, the root cause of current unbalancing in parallel disk windings was shown to be the non-uniform distribution of leakage and window inductance, which becomes particularly critical in the presence of finite core permeability and air gaps. FEM analysis demonstrated that even small inductance mismatches between parallel paths can lead to large current deviations due to the ill-conditioned nature of the inductance matrix, especially at elevated frequencies. The location of the air gap strongly affects which disks experience lower inductance and therefore higher current density.

Third, a novel mitigation technique based on series-connected

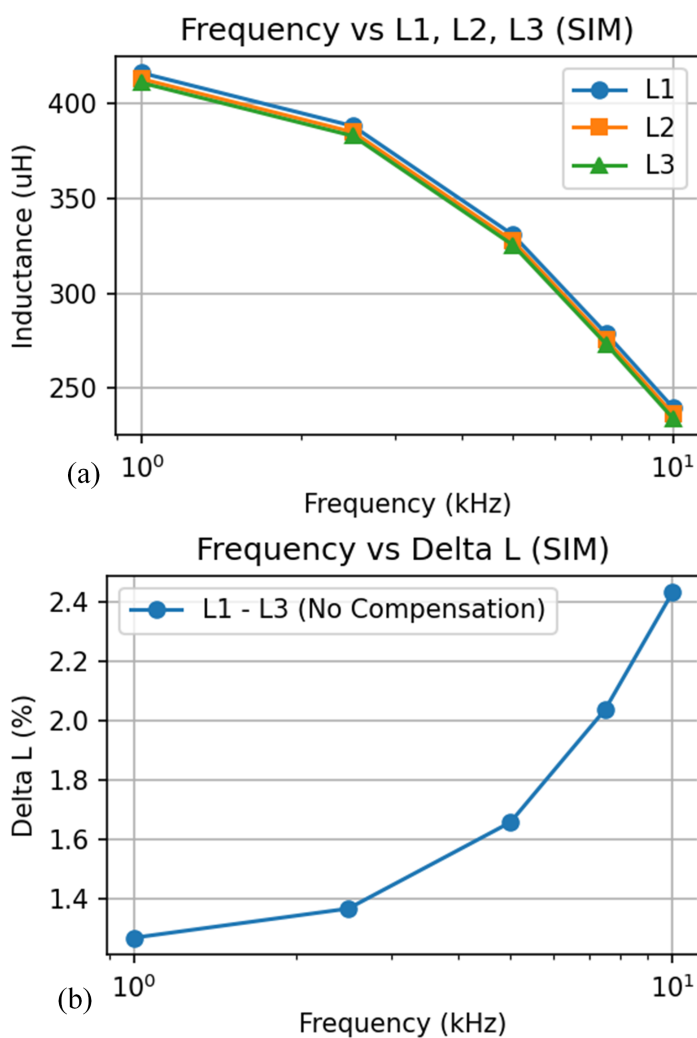


Figure 5.21: (a) Changes in the inductance of coils over frequency; (b) difference in inductance between coil #1 and coil #3 at various frequencies without compensation.

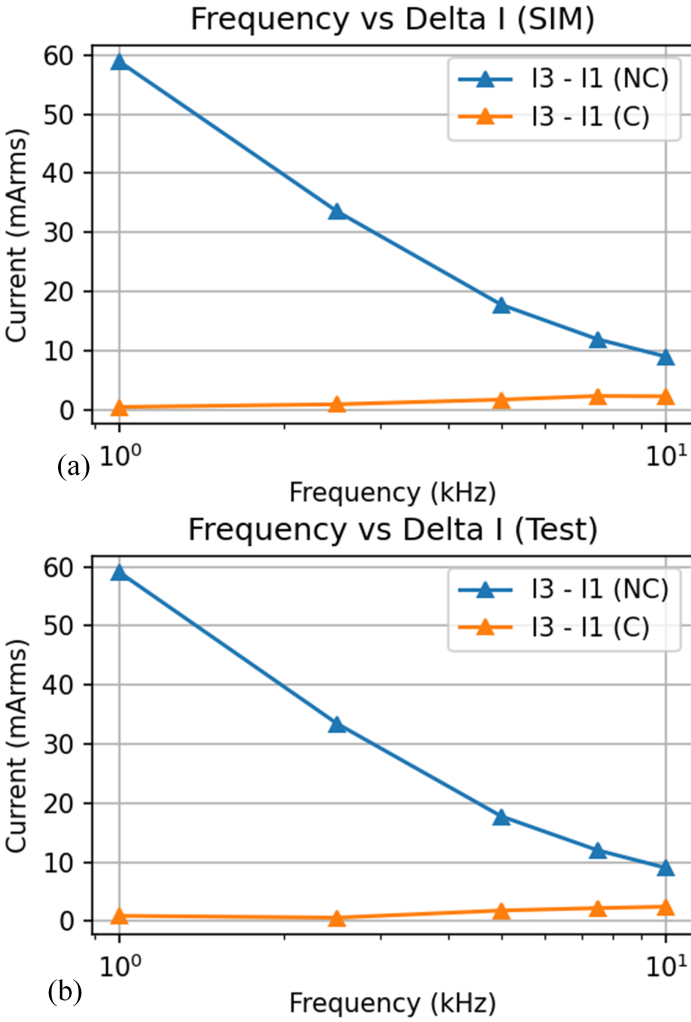


Figure 5.22: Simulation vs experimental results: (a) Difference between currents of coil #1 and coil #3 at various frequencies with (C) and without compensation (NC); (b) The same results from experiments.

compensation inductors using compact toroidal cores was proposed. By introducing small, uncoupled inductive elements in series with individual disks, the effective inductances of the parallel paths can be equalized without modifying the transformer geometry or winding layout. The method is modular, scalable, and compatible with existing transformer designs.

Simulation and experimental validation on a five-disk prototype confirmed that the proposed approach can improve current sharing from more than 80% imbalance to better than 98% uniformity, while simultaneously reducing total winding losses by approximately 28% and mitigating hotspot formation. The compensation inductances were kept well below the transformer leakage inductance (typically 0.1–1%), ensuring negligible impact on overall converter operation.

Finally, frequency-dependent measurements demonstrated that although absolute inductance differences decrease with frequency, relative mismatch becomes more critical, making compensation increasingly important at higher operating frequencies. The use of auxiliary windings or magnetic tuning enables frequency-specific or design-phase adjustment of the compensation inductors, maintaining effective current balancing across the tested frequency range up to 10 kHz.

Overall, the results show that series compensation using toroidal inductors is an effective and practical solution for disk-level current balancing in medium-voltage, medium-frequency transformers for SST-based energy systems, including large-scale green hydrogen production. The method offers a viable alternative to geometric redesign or complex conductor transposition techniques, providing improved thermal performance, reduced losses, and enhanced operational reliability.

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6

LOSSES, THERMAL MANAGEMENT, AND COOLING SYSTEM

The profound study of nature is the most fertile source of mathematical discoveries.

Joseph Fourier

This chapter provides a comprehensive analysis of the losses, thermal management, and cooling design in medium-frequency transformers used within solid-state transformer systems for green hydrogen production. It discusses the calculation and modeling of no-load and load losses, examines thermal behavior using both lumped thermal networks and finite element analysis, and outlines the design of effective cooling systems to maintain reliable operation. Experimental validation is also presented to confirm the accuracy of the thermal models under practical loading conditions.

Parts of this chapter have been published in:

Z. Li **R. Mirzadarani**, M. Ghaffarian Niasar, M. Itraj, L. van Lieshout, P. Bauer, Z. Qin, "Medium-Voltage Solid-State Transformer Design for Large-Scale H₂ Electrolyzers," in IEEE Open Journal of Power Electronics, vol. 5, pp. 936-955, 2024 [1].

There is an important trade-off between reducing the size of a transformer by increasing its operating frequency and managing its thermal performance. Higher operating frequencies can increase losses due to high-frequency effects such as the skin effect and proximity effect. Although transformers working at higher frequencies often use smaller magnetic cores, which can help limit core losses, many common core materials exhibit high losses at these frequencies. Therefore, the use of new materials becomes necessary, introducing challenges related to availability and cost. Smaller transformers also have less surface area to remove heat generated by these losses. As a result, balancing operating frequency with thermal performance is essential. Large-scale medium-frequency transformers therefore require properly designed cooling systems to ensure compliance with temperature-rise limits defined by standards for top-oil, average winding, and insulation systems, while also providing sufficient thermal margin in view of uncertainties in hotspot location and non-uniform loss distribution. This chapter introduces the different types of losses, thermal management considerations, and thermal analysis using lumped models and finite element methods (FEM), and it provides guidance on cooling system design.

6

6.1. NO-LOAD LOSS

No-load loss in a power transformer refers to the losses that occur when the transformer is energized but not supplying any load. These losses are mainly due to the magnetization of the core and include hysteresis and eddy current losses within the core material. No-load loss is primarily determined by the core material properties, the design of the core, and the operating voltage and frequency. Although the magnitude of no-load loss is generally small compared to load loss, it is continuously present as long as the transformer is energized, making it important in evaluating the transformer's efficiency, especially in applications where the transformer operates under light-load conditions for extended periods. Although there are some stray losses, such as dielectric loss, termination losses, and losses in the tank, clamps, and other conductive parts, the majority of the no-load loss comes from the magnetic core.

The core loss depends on many variables, such as the waveform and total harmonic distortion (THD), temperature, frequency, and other factors. For the 50/60 Hz sinusoidal waveform commonly used in the traditional grid, there are well-established methods precisely calculate the loss based on empirical analysis. However, for modern grid-connected solid-state transformers operating at medium or high frequencies, conventional analysis methods are not accurate enough. Determining the core loss has been the subject of various studies [2–17].

Previous works are normally based on one of the following approaches:

- **Equation-based methods** such as the Steinmetz method, the Generalized Steinmetz Method (GSM), the Improved Generalized Steinmetz Method (iGSM), and the Improved Improved Generalized Steinmetz Method (i2GSM).
- **Machine Learning (ML) or AI-based models**, which often use methods such as random forests, neural networks, and Long Short-Term Memory (LSTM) networks.
- **Mixed methods**, also called physics-informed methods, which often use equations to calculate the loss based on specific characteristics of the problem and apply ML methods, such as genetic algorithms, to categorize and optimize the solutions.

In this thesis, an equation-based model is employed to calculate core loss because the operating frequency is fixed. The frequency is not varied for control purposes; instead, controllability is achieved through phase-shift methods. The temperature also remains relatively stable, as the medium-frequency transformer (MFT) operates at or near its nominal operating point for the majority of its lifetime. This stability, combined with the advanced cooling system, ensures nearly constant temperature operation. However, the waveform is not purely sinusoidal. To address this challenge, the Fast Fourier Transform (FFT) technique is employed to decompose the waveform into its harmonic components. The corresponding loss for each harmonic is then calculated, and the total core loss is obtained by summing the losses of all harmonics.

To calculate core losses accurately in MFTs, several models based on the Steinmetz equation have been developed. These models relate the core loss to the frequency and the peak flux density of the waveform applied to the core. The classic Steinmetz equation provides a simple method to estimate losses under sinusoidal conditions, while more advanced methods such as the Generalized Steinmetz Method (GSE), Improved Generalized Steinmetz Method (iGSE), and Improved Improved Generalized Steinmetz Method (i2GSE) have been developed to handle non-sinusoidal waveforms. These methods use different strategies to incorporate the waveform shape into the calculation and improve accuracy under various operating conditions. Table 6.1 summarizes the main characteristics of these models.

These methods provide a practical way to calculate core losses in MFTs used in solid-state transformers under the non-sinusoidal conditions typically encountered in power electronic applications.

For sinusoidal waveforms, the Steinmetz Equation estimates core loss per unit volume as follows [2]:

$$P = kf^{\alpha}B_p^{\beta} \quad (6.1)$$

Table 6.1: Summary of Steinmetz-based core loss calculation methods

Method	Applicable Waveform	Description
Steinmetz Equation	Sinusoidal	Estimates core loss using frequency and peak flux density with empirical coefficients. Simple and widely used for 50/60 Hz applications.
Generalized Steinmetz Equation (GSE)	Non-sinusoidal	Extends Steinmetz equation to handle non-sinusoidal waveforms by considering the rate of change of flux density over the waveform cycle.
Improved Generalized Steinmetz Equation (iGSE)	Non-sinusoidal	Enhances GSE by improving the treatment of waveform-dependent factors, providing better accuracy for arbitrary waveforms with lower computational complexity.
Improved Generalized Steinmetz Method (i2GSE)	Non-sinusoidal	Further refines iGSE by addressing discontinuities and non-nonlinearity, offering high accuracy in calculating core losses under complex switching waveforms.

where P is the core loss density [W/m^3], f is the frequency [Hz], and B_p is the peak flux density [T]. k , α , and β are empirical Steinmetz coefficients determined from measurements.

For non-sinusoidal waveforms, the Generalized Steinmetz Equation is [18]:

$$P = \frac{1}{T} \int_0^T k_1 \left| \frac{dB(t)}{dt} \right|^\alpha B(t)^{\beta-\alpha} dt \quad (6.2)$$

where T is the period of the waveform [s], $B(t)$ is the instantaneous flux density [T], $dB(t)/dt$ is the derivative of flux density with respect to time [T/s], and k_1 can be calculated as follows:

$$k_1 = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos \theta|^\alpha |\sin \theta|^{\beta-\alpha} d\theta} \quad (6.3)$$

The Improved Generalized Steinmetz Equation simplifies GSE using piecewise approximations [19]:

$$P = \frac{1}{T} \int_0^T k_i \left| \frac{dB(t)}{dt} \right|^\alpha B(t)^{\beta-\alpha} dt \quad (6.4)$$

where k_i can be calculated as follows:

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos \theta|^\alpha 2^{\beta-\alpha} d\theta} \quad (6.5)$$

Other methods for core loss calculation, such as improved-improved Generalized Steinmetz Equation (i2GSE) [20], Modified Steinmetz Equation (MSE) [9], Natural Steinmetz Equation (NSE) [10], Improved

Steinmetz Equation (ISE) [11], Square Switching Loss Estimator (SSLE) [21], and improved General Core Loss Calculation (iGCC) [12], are also available in the literature. However, in this thesis, the focus is on the Steinmetz Equation (SE), Generalized Steinmetz Equation (GSE), and Improved Generalized Steinmetz Equation (iGSE) due to their practicality, widespread use, and clear traceability under the conditions relevant to medium-frequency transformer applications in solid-state transformers.

6.1.1. USING FFT BASED SE

In this subsection, the use of the Fast Fourier Transform (FFT) together with the Steinmetz Equation (SE) for calculating core losses is explained. This approach provides sufficiently accurate results while avoiding unnecessary complexity in the analysis.

The core material used in this study is amorphous 2605SA1 alloy. The formula for calculating the core loss for this alloy is [22, 23]:

$$P_{fe} = \rho_{fe} \cdot V_{fe} \cdot p_{fe0} \cdot \left(\frac{f}{f_0}\right)^\alpha \cdot \left(\frac{B_{pk}}{B_0}\right)^\beta \cdot \left(\frac{F}{F_0}\right)^{2(\alpha-1)} \quad (6.6)$$

where P_{fe} is the core loss [W], ρ_{fe} is the material density [kg/m³], V_{fe} is the core volume [m³], p_{fe0} is reference core loss per unit volume [W/m³], f is the frequency [Hz], f_0 is reference frequency [Hz], B_{pk} is the peak flux density [T], B_0 is reference flux density [T], F is the waveform factor, F_0 is reference waveform factor, and α, β is empirical loss coefficients.

In the system under study, the applied voltage has a fixed frequency and can be considered as a distorted sine wave at the operating fundamental frequency f , with a number of odd-order harmonics (3, 5, 7, ...). By using the FFT, the distorted waveform is decomposed into its harmonic components, and the core loss is calculated for each harmonic separately using the above formula. The total core loss is then calculated as the sum of the losses of each significant harmonic component:

$$P_{fe, \text{total}} = \sum_{i=1,3,5,7,\dots} P_{fe,i} = P_{fe,1} + P_{fe,3} + P_{fe,5} + P_{fe,7} + \dots \quad (6.7)$$

This method provides a practical and accurate approach to account for waveform distortion while maintaining the simplicity of the Steinmetz-based core loss calculation, making it suitable for analyzing medium-frequency transformers in solid-state transformer applications.

6.2. LOAD LOSS

The load loss is the dominant loss in a transformer. Load loss refers to the ohmic losses in the conductors of the transformer, mainly in the windings and busbars, due to the current passing through them. At higher frequencies, issues such as the skin effect and proximity effect become important and increase the effective resistance of the conductors. The task here is to calculate the AC resistance and use it in the following formula for each winding and busbar:

$$\text{Loss} = R_{ac} \times I^2 \quad (6.8)$$

It is essential to minimize this loss because the winding resistance increases with the total conductor length, which scales with the number of turns. Therefore, using as few turns as possible in the windings is desirable. However, this approach would require a larger magnetic core, which is not acceptable due to size and material limitations. Additionally, using fewer turns than a certain practical limit in a winding is not feasible. Thus, a trade-off exists between minimizing load loss and maintaining a practical and compact transformer design.

6

6.3. THERMAL MANAGEMENT AND COOLING DESIGN

The thermal management of medium-frequency transformers (MFTs) relies on analyzing heat transfer within their components, including windings, the magnetic core, and structural parts. Heat transfer in these systems involves conduction within solids and convection between solid surfaces and surrounding fluids. A clear understanding of these mechanisms is important for maintaining temperatures within safe limits and ensuring reliable transformer operation.

Heat conduction in solid materials follows the transient heat conduction equation:

$$\rho c_p \frac{\partial T}{\partial t} = \nabla \cdot (k \nabla T) + Q \quad (6.9)$$

where ρ is the material density, c_p is the specific heat capacity, k is the thermal conductivity, T is the temperature, t is time, and Q is the internal heat generation due to losses in the windings and core.

When fluid flow interacts with solid surfaces, convection also contributes to heat transfer. The governing equation for heat transfer in a fluid under laminar flow conditions is:

$$\rho c_p \left(\frac{\partial T}{\partial t} + \mathbf{u} \cdot \nabla T \right) = \nabla \cdot (k \nabla T) + Q, \quad (6.10)$$

where $\mathbf{u}$ is the fluid velocity field, and the term $\mathbf{u} \cdot \nabla T$ represents convective heat transport.

The fluid velocity field is described by the Navier–Stokes equations for incompressible laminar flow:

$$\rho \left(\frac{\partial \mathbf{u}}{\partial t} + \mathbf{u} \cdot \nabla \mathbf{u} \right) = -\nabla p + \mu \nabla^2 \mathbf{u}, \quad (6.11)$$

$$\nabla \cdot \mathbf{u} = 0, \quad (6.12)$$

where p is the pressure and μ is the dynamic viscosity. These equations form the basis for coupled thermal-fluid analysis in MFTs and are used to evaluate temperature distribution and cooling performance under different operating conditions.

Mathematical analysis using explicit formulas provides useful insight for simple geometries and boundary conditions. However, for real transformer structures, analytical solutions often become impractical due to geometric complexity and non-uniform heat generation. In such cases, numerical methods, including the finite difference method (FDM) and the finite element method (FEM), are applied.

FDM uses discretization of the governing equations on structured grids and is effective for simple domains with regular geometries. However, it can face challenges in handling complex geometries and boundary conditions often present in transformer analysis.

FEM offers greater flexibility by allowing the use of unstructured meshes that can accurately represent complex geometries and non-uniform material properties. FEM discretizes the domain into smaller elements where the governing equations are solved locally, enabling detailed evaluation of temperature distributions and heat transfer within the transformer. This makes FEM more suitable than FDM for the thermal analysis of MFTs.

In this study, a combined approach is used. Mathematical analysis is applied for validation in simplified cases, while FEM is employed for detailed analysis of temperature distribution and thermal resistance extraction in components with complex geometries. These thermal resistances are then used to construct lumped thermal equivalent circuit models, providing an efficient means to evaluate the thermal behavior of MFTs under different load and cooling conditions.

This approach balances computational efficiency and accuracy, supporting the design and optimization of cooling systems for medium-frequency transformers in solid-state transformer applications.

6.3.1. VALIDATION OF FEM USING ANALYTICAL AND PYTHON-BASED SOLUTIONS

To validate the FEM model and its applicability for analyzing the thermal and fluid behavior of medium-frequency transformers, several

benchmark problems were implemented and compared against analytical solutions and Python-based numerical methods. This multi-case validation ensures the accuracy of the FEM approach under various thermal and fluid conditions relevant to transformer applications.

TRANSIENT HEAT CONDUCTION IN A 1D SLAB

The first validation case studied transient heat conduction in a 1D slab, a classic problem with a well-known analytical solution using Fourier series. A plane wall with thickness $2L$ and initial uniform temperature T_i was suddenly exposed to a fluid at temperature T_∞ under convective cooling with heat transfer coefficient h . The governing heat conduction equation with appropriate initial and boundary conditions was solved using:

- Analytical solution (Fourier series), computing the temperature at the slab center over time.
- FEM using COMSOL, with matching geometry, properties, and conditions, extracting the transient temperature at the center.
- Python-based numerical solution, using explicit finite difference and Fourier evaluation for direct comparison.

The dimensionless temperature, which can be calculated as follows:

$$\theta = \frac{T - T_\infty}{T_i - T_\infty} \quad (6.13)$$

was used for consistent comparison, parameterized by the Fourier number $Fo = \frac{\alpha t}{L^2}$ and Biot number $Bi = \frac{hL}{k}$.

Observations: Excellent agreement was observed across FEM, analytical, and Python methods, confirming the FEM implementation's correctness and its suitability for extracting thermal resistances in lumped thermal models. The analytical and FEM results are compared in Figure 6.1.

TRANSIENT HEAT CONDUCTION IN A 2D RECTANGULAR PLATE

A second validation case considered transient heat conduction in a 2D rectangular plate. The geometry was a $0.1\text{ m} \times 0.1\text{ m}$ steel plate initially at 373.15 K . The left and right edges were cooled convectively ($h = 100\text{ W/m}^2 \cdot \text{K}$, $T_\infty = 293.15\text{ K}$), while the top and bottom edges were insulated. The 2D transient heat conduction equation was solved using FEM in COMSOL and Python-based FDM.

Observations: FEM results aligned with Python simulations, confirming the FEM model accurately captures transient thermal behavior under 2D conduction with convective boundary conditions. The analytical and FEM results are compared in Figure 6.2.

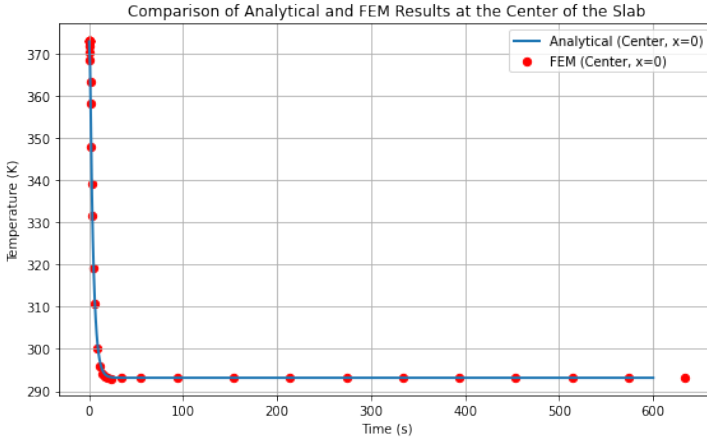


Figure 6.1: Comparison of analytical (Fourier series) and FEM results for transient heat conduction in a 1D slab at the center ($x = 0$). The analytical solution uses the Fourier series method, while the FEM results are obtained using COMSOL simulations.

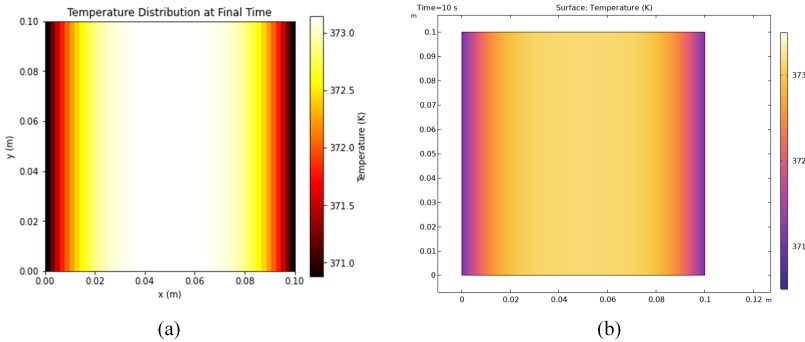


Figure 6.2: Comparison of temperature distribution in a 2D rectangular plate under transient heat conduction conditions. (a) Python-based numerical simulation using the finite difference method (FDM) at the final time step. (b) FEM results obtained using COMSOL at the same time instant.

2D HEAT CONDUCTION WITH INTERNAL HEAT GENERATION

A 2D heat conduction problem with internal heat generation in a square copper plate ($0.1\text{ m} \times 0.1\text{ m}$) was evaluated under uniform heat generation ($1 \times 10^6\text{ W/m}^3$) with all edges held at 293.15 K .

Observations: FEM and Python-based FDM results were consistent, validating the FEM implementation for scenarios involving volumetric

heating relevant to transformer windings. The analytical and FEM results are compared in Figure 6.3.

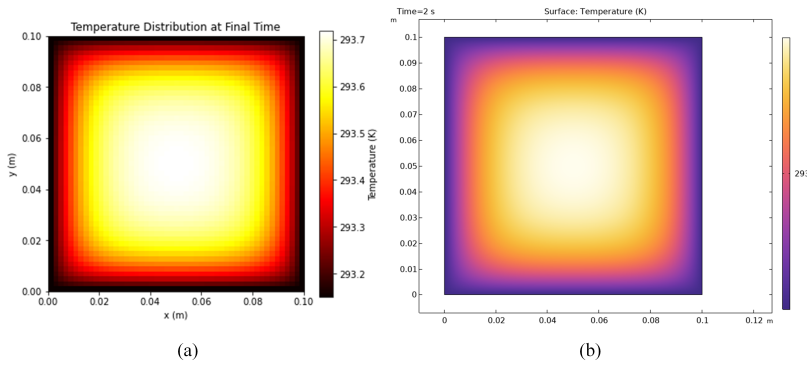


Figure 6.3: Comparison of temperature distribution in a 2D square plate with internal heat generation under transient conditions. (a) Python-based numerical simulation using the finite difference method (FDM) at the final time step. (b) FEM results obtained using COMSOL at the same time instant.

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STEADY-STATE 3D HEAT CONDUCTION IN A CUBE

A steady-state 3D heat conduction case in a 1 m cube with one face at 100°C, the opposite face at 0°C, and adiabatic other faces was analyzed. The FEM results matched the expected linear temperature profile:

$$T(x) = T_H \left(1 - \frac{x}{L} \right) \quad (6.14)$$

confirming the FEM model's reliability for 3D steady-state conduction. The analytical and FEM results are compared in Figure 6.4. The plot shows excellent agreement between the analytical linear temperature profile, numerical finite-difference (Gauss-Seidel) solution, and FEM simulation results, validating the correctness of the implemented models.

2D POISEUILLE FLOW BETWEEN PARALLEL PLATES

To validate fluid flow modeling, 2D Poiseuille flow [24] was studied under steady, incompressible laminar conditions with a constant pressure gradient and no-slip walls. FEM, Python FDM, and analytical solutions aligned closely, confirming correct implementation for forced convection analysis relevant to transformer cooling. The analytical and FEM results are compared in Figure 6.5.

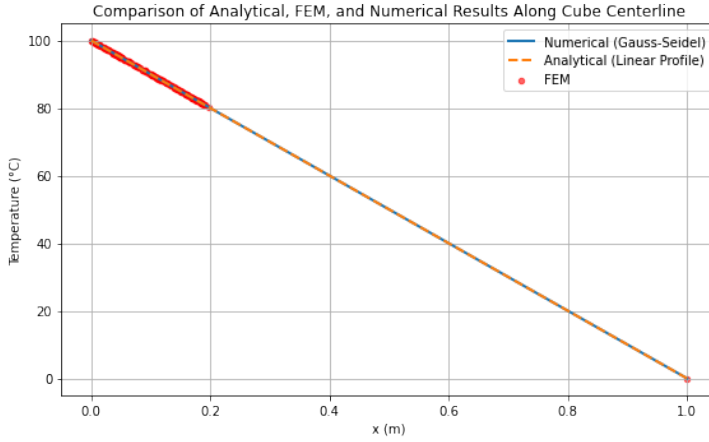


Figure 6.4: Comparison of analytical, FEM, and numerical (Gauss-Seidel) results along the centerline of a 3D cube under steady-state heat conduction.

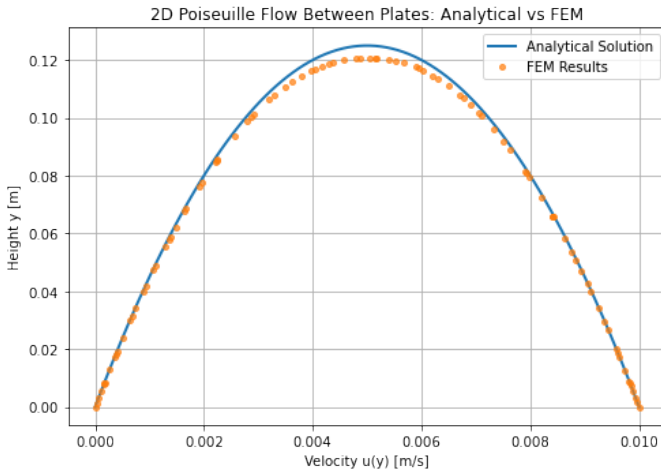


Figure 6.5: Comparison between the analytical solution and the FEM results for 2D Poiseuille flow between parallel plates.

SUMMARY OF OBSERVATIONS

- FEM results showed close agreement with analytical and Python-based solutions across all validation cases.
- The cases confirmed FEM's capability to handle transient and steady-state heat conduction, volumetric heating, and laminar flow relevant to transformer applications.

- The validation supports using FEM for extracting thermal resistances and analyzing detailed thermal and fluid behavior in transformer windings, cores, and cooling channels under realistic operational conditions.

6.3.2. LUMPED THERMAL NETWORK ANALYSIS OF TRANSFORMER LOSSES AND COOLING

This subsection explains the method used to derive the thermal resistances (R_{th}) and capacitances (C_{th}) of the transformer, construct the lumped thermal model, and analyze its thermal behavior under operational conditions using Python-based simulations. The thermal network follows a node-based representation in which the core, HV winding, and LV winding are modeled as thermal nodes with associated heat sources, and heat is transferred through thermal resistances to the surrounding oil and tank, and finally to the ambient. In other words, the dominant heat-transfer path can be represented in a single-line form as:

loss generation $\rightarrow$ winding/core $\rightarrow$ oil $\rightarrow$ tank $\rightarrow$ ambient.

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LOSS AND THERMAL RESISTANCE EXTRACTION

The primary power losses in the transformer are calculated, including:

- No-Load loss which is mainly the Core losses (P_{core})
- Load loss which mainly consists of two elements:
 - High-voltage (HV) winding losses (P_{HV})
 - Low-voltage (LV) winding losses (P_{LV})

These losses are determined from electrical simulations and measurements under nominal operating conditions. To model the thermal behavior, thermal resistances (R_{th}) are extracted using FEM simulations, where controlled heat generation is applied, and the resulting temperature rise under steady-state conditions is recorded. The thermal resistance for each component is calculated using the following equation:

$$R_{th} = \frac{\Delta T}{P} \quad (6.15)$$

where ΔT is the steady-state temperature rise due to power dissipation P . The FEM simulation using COMSOL 2D analysis is shown in Figure 6.6.

In this work, COMSOL is not used to model the complete transformer thermal behavior across all operating points. Instead, the FEM model is used as a parameter-extraction tool to obtain representative (effective) thermal resistances by applying controlled heat sources and recording the steady-state temperature rise, as in Equation (6.15). This approach

was selected because a full 3D conjugate heat-transfer model including all transformer sections, oil flow, tank, piping, and boundary conditions would substantially increase model complexity and computational cost, while offering limited additional benefit for the objectives of this chapter, i.e., system-level thermal dynamics and design-oriented iteration using a lumped thermal network.

It is important to note that thermal resistances and capacitances obtained from a reduced FEM domain can depend on the surrounding structures that are not explicitly included in the COMSOL model. Neighboring sections may provide additional parallel conduction paths, modify contact areas, and alter the local convection field in the oil, leading to a change in the effective heat-transfer coefficients and therefore in the extracted values of R_{th} . For this reason, the reported R_{th} values should be interpreted as effective parameters associated with the specific geometry, interfaces, and boundary conditions used during extraction. The lumped thermal network formulation in Equation (6.17) is then used to integrate these effective parameters at system level, where the impact of neglected couplings is mitigated through (i) selecting boundary conditions that represent the physical interfaces between components, and (ii) validating the resulting temperature dynamics against prototype measurements (Chapter 7). Where required, the uncertainty introduced by the reduced-domain extraction can be handled by sensitivity analysis or correction factors on the dominant R_{th} paths.

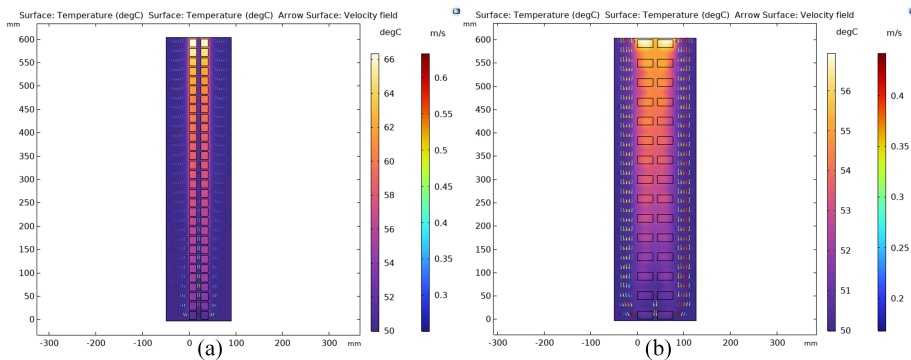


Figure 6.6: COMSOL FEM simulation results for steady-state thermal analysis used in thermal resistance extraction. (a) Primary side: surface temperature distribution (°C) and velocity field. (b) Secondary side: surface temperature distribution (°C) and velocity field.

CALCULATION OF THERMAL CAPACITANCES

The thermal capacitances (C_{th}) for the core and windings are calculated using the material properties and geometrical parameters:

$$C_{th} = mc_p \quad (6.16)$$

where m is the mass of the component and c_p is the specific heat capacity. These capacitances are important for capturing the transient thermal behavior and determining the thermal time constants of the system.

LUMPED THERMAL NETWORK CONSTRUCTION

An equivalent lumped thermal network is constructed with:

- Nodes representing the core, HV winding, and LV winding temperatures.
- Thermal resistances connecting these nodes to each other and to the ambient.
- Power losses applied as heat sources to the corresponding nodes.
- Thermal capacitances associated with each node to capture transient effects.

The structure of the lumped thermal network allows the system of thermal balance differential equations to be solved explicitly:

$$C_{th,i} \frac{dT_i}{dt} = \sum_j \frac{T_j - T_i}{R_{th,ij}} + P_i \quad (6.17)$$

where T_i and T_j are node temperatures, and P_i is the heat generation at node i . The lumped model of the transformer in Simulink is shown in Figure 6.7

TRANSIENT THERMAL ANALYSIS

The transient thermal behavior is simulated over a time period of 1.1 to 1.4 hours sufficient to observe the heating curves and the time required to reach steady-state conditions. The time-dependent temperature rise in the core, HV winding, and LV winding is computed using an explicit or implicit time-stepping method in Python, and the results are compared against steady-state FEM simulations for validation.

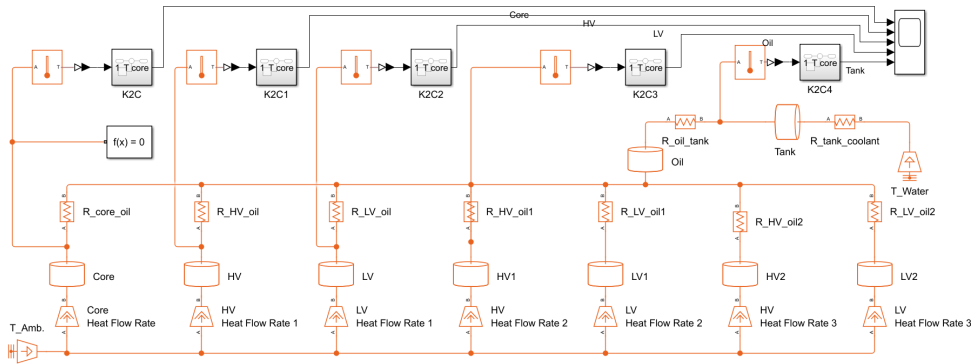


Figure 6.7: Lumped thermal network model of the transformer implemented in Simulink. The model includes nodes for the core, high-voltage (HV) windings, low-voltage (LV) windings, and the oil tank, connected via thermal resistances. Heat sources representing power losses are applied to each node, while thermal capacitances enable capturing transient thermal behavior.

RESULTS AND DISCUSSION

The simulation results show:

- The LV winding exhibits the highest steady-state temperature due to its higher power losses.
- The thermal time constants of the core and windings are consistent with expected values based on the calculated R_{th} and C_{th} .
- The transient response demonstrates smooth and stable heating behavior, confirming the model's numerical stability.
- Comparison with FEM-based steady-state temperatures validates the accuracy of the lumped thermal network in predicting hotspot and component temperatures under operational load conditions.

The lumped thermal model provides a computationally efficient yet accurate method for analyzing transformer thermal performance, enabling effective assessment of cooling system adequacy and thermal design under operational scenarios [25–28]. The results of python-based model and the Simulink lumped model simulation are compared in Figure 6.8

6.4. EXPERIMENTAL VALIDATION USING A PROTOTYPE

The thermal analysis and simulation results described in the previous sections were validated using a prototype winding under controlled

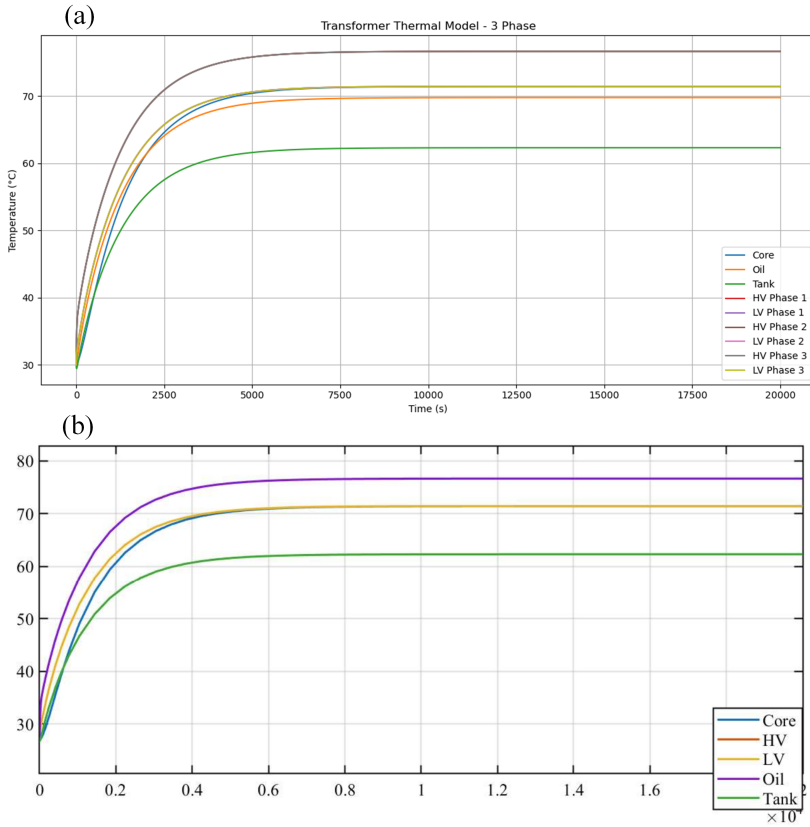


Figure 6.8: Transient thermal response of the transformer under operational loading conditions. (a) Python-based lumped thermal network simulation showing the temperature rise in the core, HV and LV windings, oil, and tank until steady-state. (b) Simulink-based lumped thermal network simulation of the same case for cross-validation, demonstrating consistent thermal time constants and steady-state temperatures with the Python model.

test conditions. This validation was performed to confirm the accuracy of the lumped thermal network model, the finite element analysis results, and the practical thermal behavior of the system under operational loading.

In this experiment, a prototype winding was excited by nominal DC current, and the temperature rise of each disk was recorded over time. The winding tested is one of the primary windings of prototype #2 described in Chapter 7, consisting of 16 disks in series (D1 to D16). The

purpose of this measurement was to confirm the hotspot temperature rise and to identify the location of the hotspot within the winding. The lumped model of the prototype winding under the test is shown in Figure 6.9. The results of python-based analysis and the simulink model are compared in Figure 6.10. The figure shows the temperature rise of the HV winding over time under nominal loading, demonstrating excellent agreement between the two methods in capturing the transient and steady-state thermal behavior of the system.

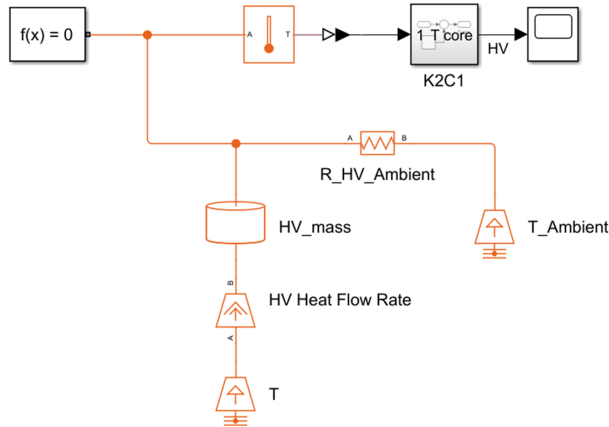


Figure 6.9: Lumped thermal network model of the prototype winding under test, implemented in Simulink. The model includes the HV winding with its thermal capacitance, heat generation source, and thermal resistance to the surrounding air, representing the thermal behavior of the winding during the validation test.

The experimental results for 16 disks are shown in Figure 6.11.

The results show that disk D5, which is the fifth disk from the top, exhibited the highest steady-state temperature among all disks. This indicates that D5 is the hotspot within the winding, a result that aligns with the FEM simulation predictions. The hotspot at this location is explained by the buoyancy effect within the transformer tank, which leads to higher oil temperatures near the top of the winding, increasing the local thermal resistance and reducing heat transfer at that point.

This experimental validation confirms that the thermal models used, including the lumped thermal network and FEM simulations, accurately capture the transient and steady-state thermal behavior of the transformer winding. Additionally, the test provides practical insight into the thermal distribution within the winding under operational conditions, supporting the effectiveness of the developed thermal analysis and

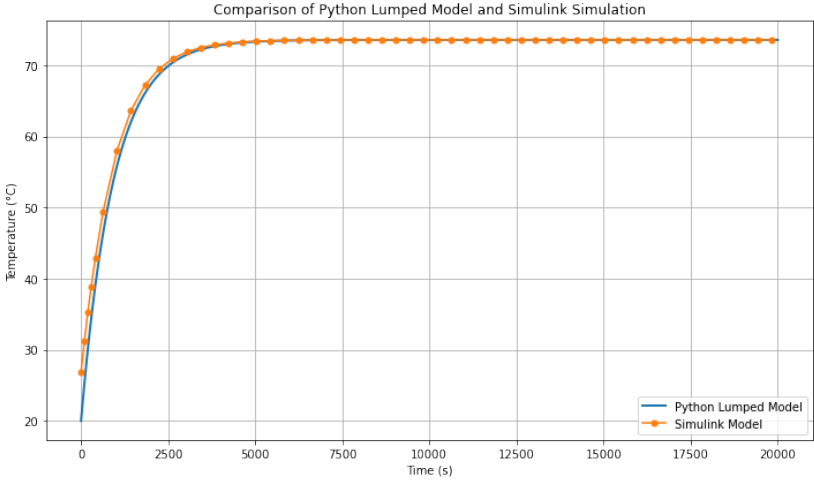


Figure 6.10: Comparison between the Python-based lumped thermal network model and the Simulink model for the prototype winding under test conditions.

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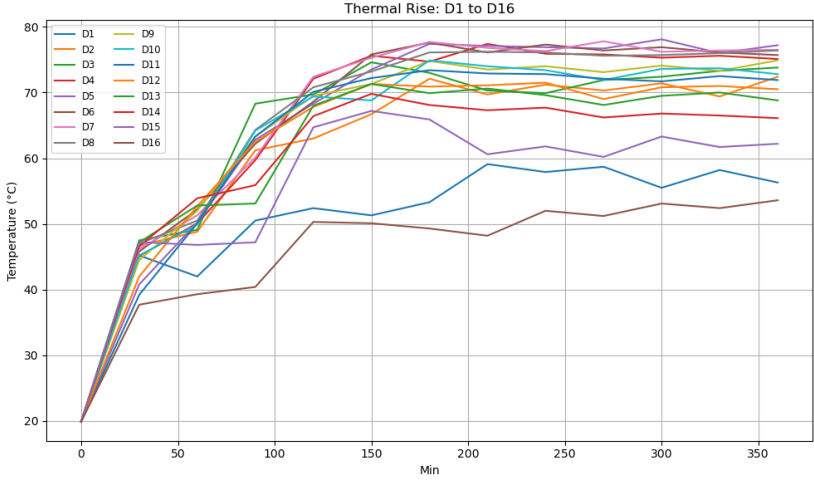


Figure 6.11: Measured thermal rise of each disk (D1 to D16) in the prototype winding under nominal DC current excitation. The temperatures were recorded over time for each disk to evaluate the transient and steady-state thermal behavior of the winding.

cooling design methods for medium-frequency transformers in solid-state transformer applications.

6.5. KEY FINDINGS AND CONCLUSIONS

This chapter analyzed the dominant loss mechanisms and the thermal behavior of the medium-frequency transformer (MFT), and it established a practical workflow for cooling-system design based on a combination of FEM parameter extraction and lumped thermal network simulation. The main findings are summarized as follows:

- **Frequency-thermal trade-off:** Increasing operating frequency reduces the magnetic component size but generally increases winding AC resistance due to skin and proximity effects and can also increase core losses depending on the selected material and excitation waveform. Therefore, frequency selection must be considered jointly with thermal constraints and cooling capability.
- **Dominant loss contributors:** Under nominal operation, the **load loss** in windings and current-carrying parts is the dominant contributor to total losses, while **no-load loss** is mainly associated with the magnetic core. For the non-sinusoidal excitation in solid-state transformer operation, waveform distortion must be accounted for in the core-loss model (e.g., via FFT-based harmonic decomposition).
- **Cooling criteria beyond a single hotspot value:** Cooling-system design is not solely driven by a single predicted hotspot temperature. In practice, **hotspot locations are uncertain** due to non-uniform loss distribution, manufacturing tolerances, local oil-flow patterns, and interface/contact variations. Therefore, the cooling concept must ensure compliance with **standard temperature-rise limits** (top-oil and winding) while maintaining adequate thermal margin against localized overheating.
- **Role of FEM in this chapter:** The FEM model is used as a **parameter-extraction tool** to obtain effective thermal resistances under representative boundary conditions. These effective parameters enable fast system-level iteration using lumped thermal networks, avoiding the cost and complexity of full 3D conjugate heat-transfer simulations for every operating point.
- **Transient thermal dynamics:** The lumped thermal network predicts smooth and stable heating behavior. The dominant thermal time constant is on the order of [1–2] ks, and approximately **95% of steady-state temperature is reached after about [4–5] ks** ($\approx$ [1.1–1.4] h). This timescale is important for defining test durations and interpreting short-term overload behavior.
- **Experimental confirmation:** Prototype measurements confirmed the predicted transient and steady-state temperature behavior

and indicated a **repeatable hotspot location** (disk D5 in the tested winding), consistent with the expected buoyancy-driven oil-temperature stratification in the tank.

Overall, the proposed workflow (loss calculation → FEM-based extraction of effective thermal parameters → lumped-network simulation → prototype validation) provides a computationally efficient and traceable approach to evaluate thermal limits and cooling adequacy in MFTs for solid-state transformer applications.

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7

PROTOTYPING AND EXPERIMENTAL SETUP

*In theory, there is no difference between theory and practice.
In practice, there is.*

Yogi Berra

This chapter outlines the comprehensive prototyping and experimental validation process of a three-phase medium-voltage, medium-frequency transformer (MV-MFT) specifically designed for solid-state transformer (SST) applications in green hydrogen production. It includes design requirements, optimization methodology, winding and core configuration, finite element simulations for electromagnetic fields and thermal behavior, as well as fabrication steps. Extensive experimental tests, such as Bode analysis, partial discharge measurement, insulation breakdown, and thermal rise tests, are conducted to validate the design. Finally, the transformer is assembled and evaluated under SST system-level testing conditions.

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Z. Li **R. Mirzadarani**, M. Ghaffarian Niasar, M. Itraj, L. van Lieshout, P. Bauer, Z. Qin, "Medium-Voltage Solid-State Transformer Design for Large-Scale H2 Electrolyzers," in IEEE Open Journal of Power Electronics, vol. 5, pp. 936-955, 2024 [2].

L. Bolzonella **R. Mirzadarani** M. Ghaffarian Niasar, "Medium Frequency Transformer Optimization and Insulation Testing for Solid State Transformer Applications," IECON 2025- 51st Annual Conference of the IEEE Industrial Electronics Society, Madrid, Spain, 2025, pp. 1-6 [3]

This chapter presents the prototyping process and experimental setup developed for the medium-voltage, medium-frequency transformer (MV-MFT) used in the solid-state transformer demonstrator. The design objectives, boundary conditions, and optimization criteria are first introduced, followed by a discussion of candidate configurations and the selection procedure. The optimized design is then analyzed through electromagnetic and electrostatic simulations to verify magnetic flux distribution, electric field stress, and insulation margins. The chapter continues with the design of conductor terminations and concludes with a description of the experimental tests, including partial discharge measurements, breakdown tests, transformer performance evaluation, and SST system-level validation. The configuration of FlexH2 innovation concept for MW-level system and the demonstrator diagram are illustrated in Figure 7.1.

The procedure of prototyping is divided into three steps, as shown in Figure 7.2: design space analysis, topology optimization, and fine-tuning with verification.

7.1. DEMONSTRATOR REQUIREMENTS

This section gives the design of a 30 kVA (25 kW) three-phase dry-type medium-frequency transformer of the demonstrator for the “GROW H2 SST Demonstrator” project.

The specifications of the demonstrator are as follows:

- Input power: $P = 25 \text{ kW}$ (apparent power: 30 kVA)
- Transfer Ratio (TR) = 46.2
- Primary line current: $I_{p_L} = 7.07 \text{ A}$
- Secondary line current: $I_{s_L} = 326.6 \text{ A}$
- Nominal voltage: $U_N = 2.67 \text{ kV}$
- Nominal frequency: $f_N = 1 \text{ kHz}$
- THD below 5%
- Leakage inductance of 7.7-10.2 mH
- $\int V \cdot dt = 0.6 \text{ Vs}$

The value 0.6Vs means that the RMS voltage is 2.665kV, which is 8.8% higher than the expected U_N (2.45kV) due to the square-shaped waveform of the input voltage. Additionally, electric and magnetic fields will have to be maintained below safety margins to prevent breakdown and core saturation, and thermal behavior should be bounded to prevent thermal runaway. Numerical values will depend on material and insulation selection.

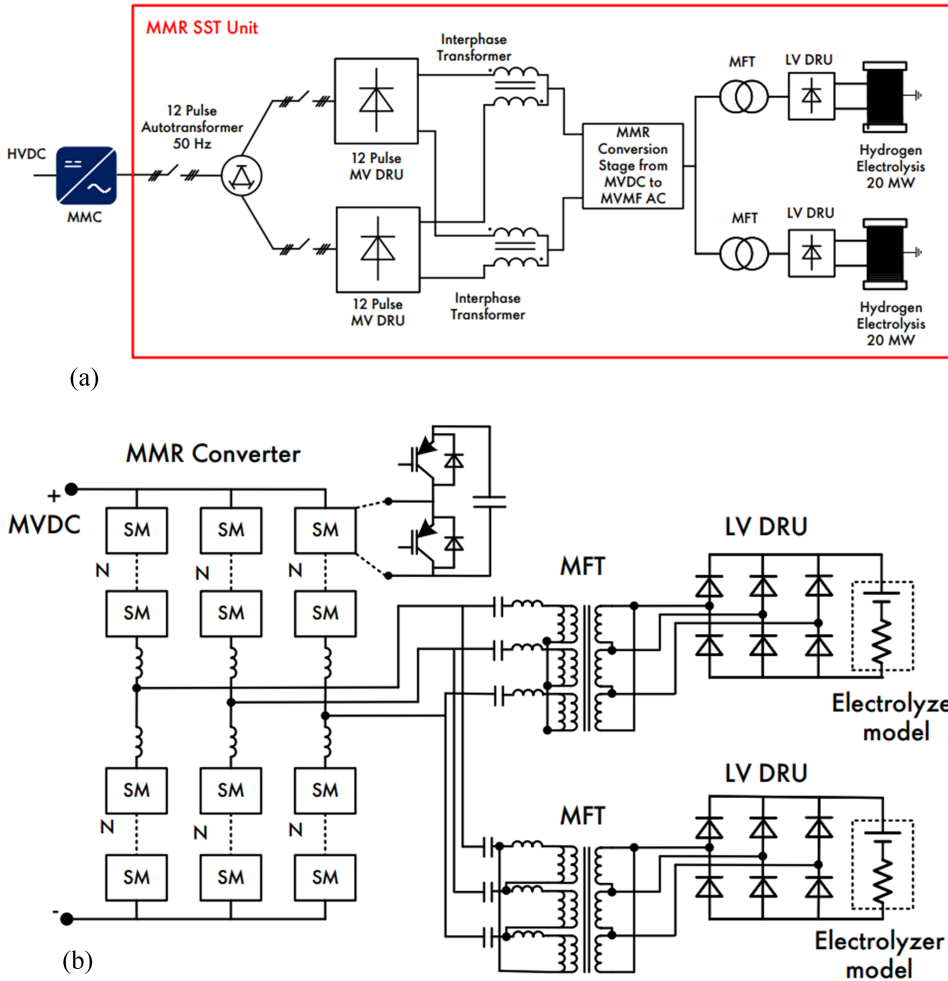
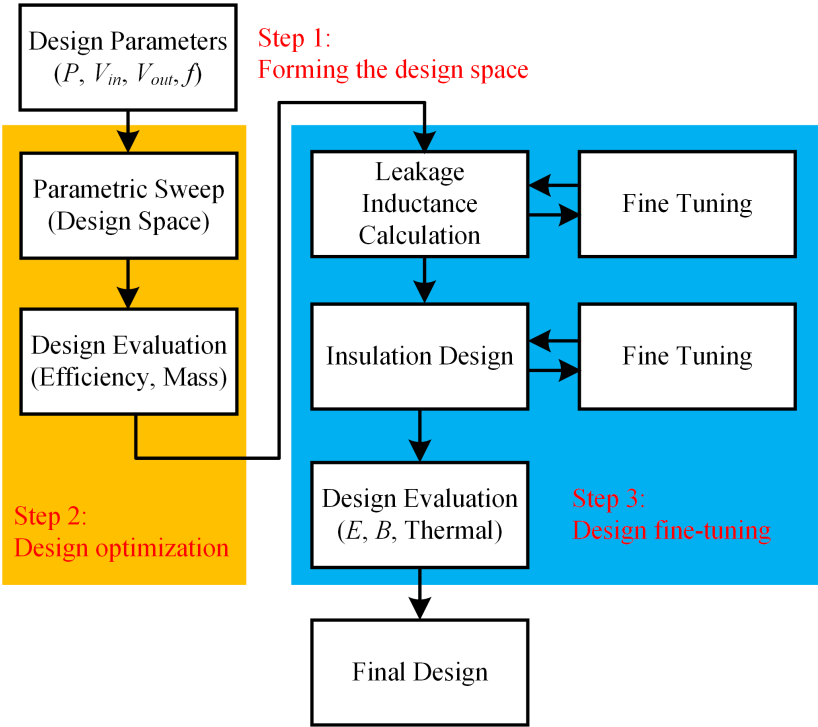


Figure 7.1: a) Configuration of FlexH2 innovation concept MV-level solid-state transformer, b) The circuit diagram of MMR and connected system in the demonstrator.

7.2. DEMONSTRATOR DESIGN

This section analyses different topologies and design parameters to generate a table of possible designs, which will then be evaluated in terms of mass and efficiency. From the design considerations presented in the Literature Review chapter, four main topologies can be identified for the core and winding configurations, as displayed in Figure 7.3:

- Core-type cylindrical winding



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Figure 7.2: Design methodology pipeline consisting of three main steps: (1) defining the design space by identifying requirements and design parameters, (2) performing topology optimization through parametric sweep and performance evaluation, and (3) fine-tuning the design considering leakage inductance, insulation constraints, and electromagnetic/thermal behavior.

Table 7.1: Conversion of voltage–time area to equivalent rms voltage (demonstrator input)

Quantity	Symbol	Value
Voltage–time area (given)	$\int V dt$	0.6 Vs
Operating frequency	f	1 kHz
Expected nominal rms voltage	U_N	2.45 kV
Equivalent rms voltage (square waveform)	$U_{rms,sq}$	2.665 kV
Absolute increase	$\Delta U = U_{rms,sq} - U_N$	0.215 kV
Relative increase	$\Delta U/U_N$	8.8%

- Core-type disk winding
- Shell-type cylindrical winding
- Shell-type disk winding

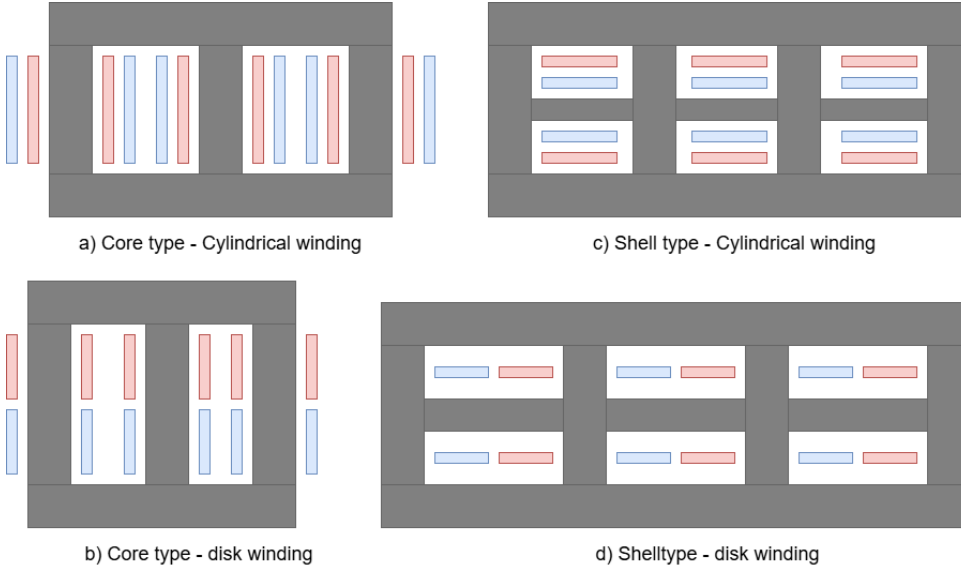


Figure 7.3: Core and shell-type transformer configurations with cylindrical and disk windings. Red and blue blocks represent primary and secondary windings, receptively.

To analyze each topology, a default case with four turns in the secondary winding is considered. This will allow us to highlight the drawbacks and advantages of each configuration, to then select the best approach when performing the design optimization. Additionally, some parameters are provided ad hoc from the other components of the SST. In particular, it is noted:

- $\int V \cdot dt = 0.6Vs$
- RMS primary voltage 2.665 kV due to the square-shaped waveform

7.2.1. WINDING DESIGN

To perform the winding analysis, a base case is examined, which is then further developed to explore additional options. For the case with $N_s = 4$ secondary turns, the number of primary turns is calculated as 185, based on the required transformer ratio of 46.2. The phase currents on the primary and secondary sides are derived as:

$$I_p = \frac{I_{pL}}{\sqrt{3}} = \frac{7.07}{\sqrt{3}} = 4.08 \text{ A} \quad (7.1)$$

$$I_s = \frac{I_{sL}}{\sqrt{3}} = \frac{326.6}{\sqrt{3}} = 188.56 \text{ A} \quad (7.2)$$

To account for over-current scenarios, a safety margin of 10% is used; therefore, the design current values used are 4.5 A (primary) and 207 A (secondary). The conductor cross-sectional area is constrained by a maximum current density of 4 A/mm². Therefore, the minimum surface area required for the primary conductor is:

$$a_p = \frac{I_p}{J_{p,\max}} = \frac{4.08}{4} = 1.02 \text{ mm}^2 \quad (7.3)$$

At elevated frequencies, the skin effect becomes significant, limiting current penetration into the conductor. For a copper conductor operating at 1 kHz, the skin depth is given by:

$$\delta = \sqrt{\frac{1}{\pi \mu f \sigma}} \approx \frac{0.066}{\sqrt{f}} = 2.087 \text{ mm} \quad (7.4)$$

where μ and σ represent the permeability and electrical conductivity of the material, respectively. Maintaining conductor dimensions within 4 mm ensures the skin effect remains negligible. To ease the winding and assembly procedure, a rectangular conductor is selected.

Due to the large number of turns, the primary winding is constructed using two disk types: Type 1 (multi-turn) and Type 2 (single-turn). This modular winding approach eases the construction process and conductor placement on the bobbin.

For the base case, each Type 1 disk contains 11 turns. A total of 16 Type 1 and 15 Type 2 disks yields:

$$N_{\text{total}} = 16 \times 11 + 15 \times 1 = 176 + 15 = 191 \text{ turns} \quad (7.5)$$

To reach the required 185 turns, six Type 1 disks are reduced by one turn each:

$$N_{\text{adjusted}} = 191 - 6 = 185 \text{ turns} \quad (7.6)$$

Spacing assumptions include 1 mm between turns and 2 mm between disks. An additional 2 mm margin is added on either side to accommodate the bobbin structure. Figure 7.4 illustrates the resulting winding structure using a 2 × 1 mm conductor. The inter-disk gap (G_{dd}) is set at 4.2 mm.

For the secondary winding, a crucial consideration is to distribute the high current across multiple parallel paths to minimize conduction

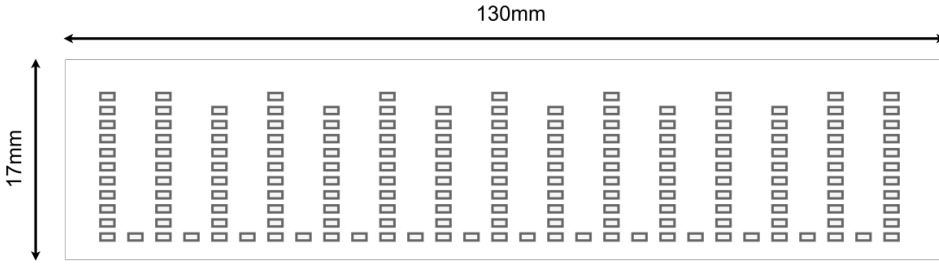


Figure 7.4: Primary winding example for $N_s = 4$ and 2x1 mm conductor.

losses. An initial estimate assumes 15 parallel conductors, resulting in approximately 14 A per path. To maintain the current density within the maximum allowable limit of 4 A/mm², the minimum conductor cross-sectional area is calculated as:

$$a_s = \frac{I_{\text{per path}}}{J_{\text{max}}} = \frac{14}{4} = 3.5 \text{ mm}^2 \quad (7.7)$$

The spacing between turns and adjacent parallel paths is chosen to ensure that the overall cross-sectional dimensions of the primary and secondary windings remain comparable, thereby ensuring symmetry in magnetic coupling and thermal behavior. An example of the secondary winding layout for the case $N_s = 4$, using 2 × 2 mm rectangular conductors, is shown in Figure 7.5.

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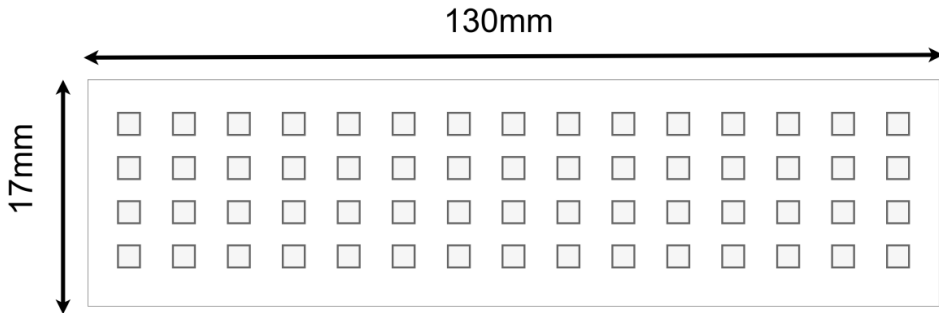


Figure 7.5: Secondary winding example for $N_s = 4$ and 2x2 mm conductor.

The total mass of the conductors can be estimated using the mean length per turn (MLT), the number of turns, and the cross-sectional area of the conductors. For each winding, the total conductor mass is given by:

$$M_{CP} = 3 \cdot MLT_p \cdot N_p \cdot a_{cp} \cdot \rho \quad (7.8)$$

$$M_{CS} = 3 \cdot MLT_s \cdot N_s \cdot a_{cs} \cdot PPS \cdot \rho \quad (7.9)$$

where a_c is the conductor cross-sectional area, ρ is the material density, and PPS is the number of parallel paths in the secondary winding. The factor of 3 accounts for all three phases in the medium-frequency transformer (MFT).

7.2.2. CORE DESIGN

At the operating frequency of 1 kHz, conventional core materials such as ferrite and electrical steel are generally not optimal for a high-efficiency design. Ferrites are limited by their relatively low saturation flux density, whereas electrical steel typically exhibits elevated core losses at kHz frequencies due to eddy-current effects. Nevertheless, for the prototype in this chapter, electrical steel is selected as a pragmatic choice driven by availability, mechanical robustness, and ease of fabrication; the prototype is primarily intended for integration and insulation/functional validation rather than loss-optimized continuous operation at 1 kHz.

Grain-oriented Electrical Steel has a maximum flux density of up to 2.0 T; however, due to high-frequency losses, $B_m = 1.8$ T is considered for the design of the core's cross-sectional area to maintain a safety margin and avoid core saturation. Using this value, the minimum required core cross-sectional area is calculated as:

$$A_c = \frac{\int V dt}{k_c B_m N_p} = \frac{0.6}{0.79 \times 1.8 \times 185} = 2916 \text{ mm}^2 \quad (7.10)$$

where $k_c = 0.54$ is the stacking factor for the selected lamination thickness. Assuming a square cross-section, this corresponds to a minimum core leg width of approximately 54 mm.

Core window dimensions are topology-dependent and will be refined in later stages of the design. Initial insulation and spacing assumptions are taken from the minimum insulating distances specified in IEC 60076-11:

- Core leg to phase gap: $G_{pl} = 20$ mm
- Inter-phase gap: $G_{pp} = 20$ mm
- Core yoke to phase gap: $G_{py} = 20$ mm
- Primary to secondary gap: $G_{ps} = 20$ mm

It is important to note that the closest distance from the core leg to the winding occurs along the diagonal of the square cross-section. This is illustrated in Figure 7.6.

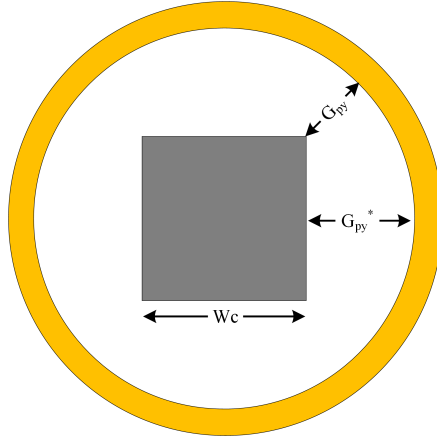


Figure 7.6: Phase to core leg gap with square core section.

The dimension s of the transformer are illustrated in Figure 7.7. Based on the selected winding dimensions, 130 mm in height and 17 mm in width for both primary and secondary windings, the window sizes can be estimated for each configuration listed before. This provides a preliminary framework for evaluating the design options.

For case (a), corresponding to a core-type configuration with cylindrical windings, the window dimensions ($W_{wa} \times H_{wa}$) are calculated as:

$$W_{wa} = 2(2W_p + G_{ps} + G_{pl1} + \frac{1}{2}G_{pp}) = 2(2 \cdot 17 + 20 + 32.4 + 10) = 222.8 \text{ mm} \quad (7.11)$$

$$H_{wa} = H_p + 2G_{py} = 130 + 2 \cdot 20 = 170 \text{ mm} \quad (7.12)$$

For case (b), the core-type configuration with disk windings:

$$W_{wb} = 2W_p + G_{pp} + 2G_{pl1} = 2 \cdot 17 + 20 + 2 \cdot 32.4 = 118.8 \text{ mm} \quad (7.13)$$

$$H_{wb} = 2H_p + G_{ps} + 2G_{py} = 2 \cdot 130 + 20 + 40 = 320 \text{ mm} \quad (7.14)$$

For case (c), representing a shell-type transformer with cylindrical windings:

$$W_{wc} = H_p + 2G_{pl} = 130 + 2 \cdot 32.4 = 194.8 \text{ mm} \quad (7.15)$$

$$H_{wc} = 2W_p + 2G_{pl} + G_{ps} = 2 \cdot 17 + 2 \cdot 32.4 + 20 = 118.8 \text{ mm} \quad (7.16)$$

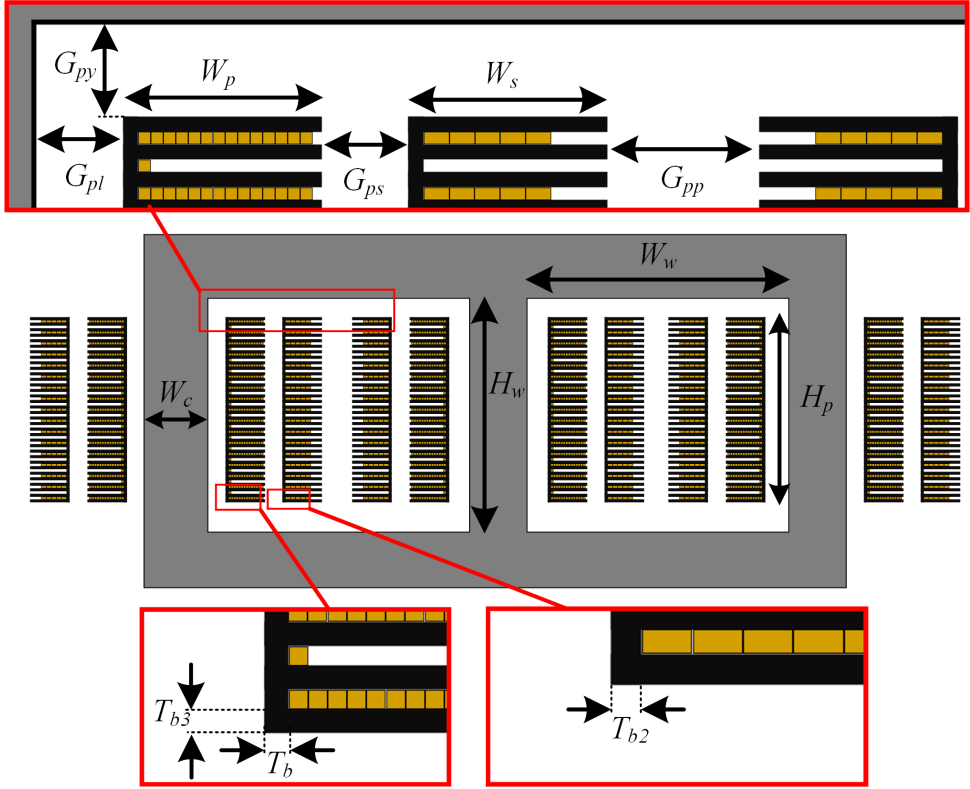


Figure 7.7: Dimensions of the transformer.

For case (d), the shell-type configuration with disk windings, the side window dimensions are:

$$W_{wd2} = H_p + 2G_{pl} + G_{ps} = 130 + 2 \cdot 32.4 + 20 = 214.8 \text{ mm} \quad (7.17)$$

$$H_{wd2} = W_p + 2G_{pl} = 17 + 2 \cdot 32.4 = 81.8 \text{ mm} \quad (7.18)$$

Since one of the key goals of SST design is mass reduction relative to conventional LLC-based transformers, the core mass for each configuration is calculated using a fixed core leg width of 60 mm and the corresponding window dimensions. Assuming a magnetic core density of 7300 kg/m^3 , the resulting masses are summarized in Table 7.2.

Although shell-type configurations offer certain electrical advantages, such as reduced magnetic path, they are not selected in this design due to the substantial increase in core mass, exceeding 40%, when

compared to core-type configuration. Thus, a core-type configuration has been selected.

Table 7.2: Core volume and mass depending on topology

MFT topology	Volume [m ³]	Mass [kg]
Core-type cyl winding	6.4e ⁻³	46.2
Core-type disk winding	6.7e ⁻³	48.2
Shell type - cyl winding	9.8e ⁻³	71.5
Shell type - disk winding	9.4e ⁻³	68.8

7.2.3. DESIGN OPTIMIZATION

With the previous analysis in mind, it is now possible to identify the key parameters used for the optimization phase, as described in the design methodology. Some design features are fixed based on prior validation of their performance and feasibility:

- Core-type configuration;
- Cylindrical winding structure;
- Rectangular conductor cross-section.

The following parameters are selected for the optimization sweeps:

- **Number of secondary turns (N_s):** Prior analysis indicates that 4 or 5 turns are sufficient to meet design requirements.
- **Primary current density:** Controlled by varying the conductor cross-section from 0.5 mm to 2.5 mm in both dimensions.
- **Secondary current density:** Influenced by conductor size (1.5 mm to 3 mm per side) and the number of parallel paths, which helps distribute the current and reduce per-wire loading.
- **Geometrical spacings:** Fine-tuning of the core width, the gap between primary and secondary windings, inter-phase spacing, and conductor-to-core clearance.
- **Conductor material:** Both copper and aluminum are considered viable; copper offers superior conductivity but at a higher density compared to aluminum.
- **Number of Turns per Disk Type 1 (NTDP):** Set to 11 for designs with $N_s = 4$, and 14 for those with $N_s = 5$.

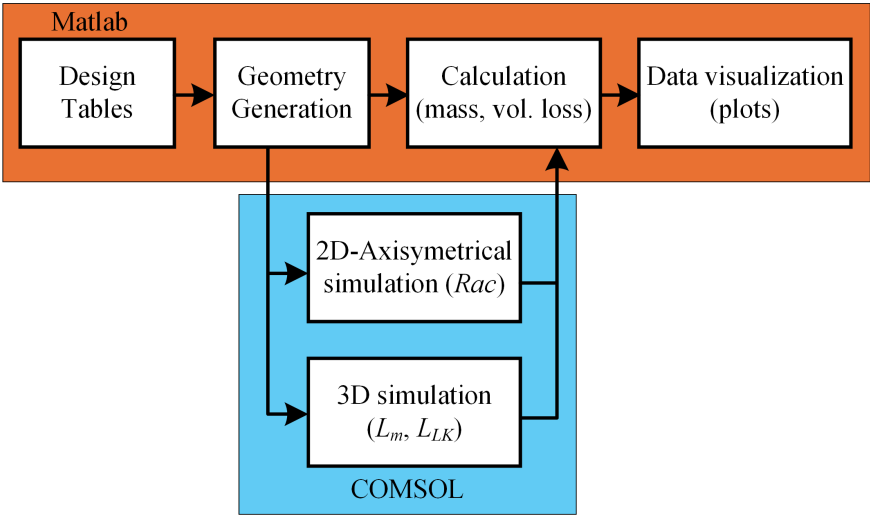


Figure 7.8: Optimization block diagram

In addition to these electrical and geometric design parameters, a bobbin is required to structurally support and guide the windings during the manufacturing process. The optimization block diagram is shown in Figure 7.8.

First, the design tables are generated in MATLAB. Based on the design space identified in this section, a table is created containing the possible permutations of parameters (conductor material and size, number of secondary turns, window and core size, etc). The result is a table containing approximately 1800 designs. An example of a table entry is shown in Table 7.3, summarizing the parameters needed to create a transformer model, where W_{cp} and W_{cs} are the widths of the conductors, H_{cp} and H_{cs} are the heights of the conductors, PPs is the number of parallel paths on the secondary, NTDP is the number of turns in a Type 1 disk, NDP is the total number of disks in the primary winding, and H_p and H_s are the sizes of the windings.

T_{b1} to T_{b3} represent the extra space the bobbin would take around the windings. G_{pc} to G_{wy} are the insulation distances. Cmat indicates the conductor material (1 for copper and 2 for aluminum), and W_c is the total conductor width. All sizes are listed in millimeters.

Table 7.3: Example entry for design table

Wcp	Hcp	Wcs	Hcs	PPs	NTDP	NDP	Hp	Hs
0.5	2.5	2.5	1.5	17	14	31	128	128
Tb1	Tb2	Tb3	Gpc	Gps	Gww	Gwy	Cmat	Wc
2.4	2.1	2.5	16	20	20	20	1	54

A design sweep algorithm is then implemented in MATLAB, retrieving each table entry, generating the data for parametrization and creation of a COMSOL model to measure winding ac resistance, and finally computing the efficiency and weight of the design. The result is a plot comparing the mass-efficiency for each design to facilitate in identifying the best trade-off for this specific application. The first step is to calculate the geometric sizes of the core to compute the core weight and generate the COMSOL model.

From the data on the table, the code computes the sizes of the width of the window as follows:

$$W_w = 2 \times \left(G_{pl} + T_b + W_p + T_{b2} + G_{ps} + T_b + W_s + T_{b2} + \frac{G_{ww}}{2} \right) \quad (7.19)$$

Where G_{pl} is the gap between primary and core due to geometry, T_b is the bobbin thickness (main cylinder), W_p is the width of the primary winding, T_{b2} is the insulation over the winding, G_{ps} is the gap between primary and secondary windings, W_s is the width of the secondary winding, and G_{ww} is the gap between adjacent winding windows (split equally).

The height of the window can be calculated as follows:

$$H_w = 2 \times (G_{wy} + T_{b3}) + H_p \quad (7.20)$$

Where G_{wy} is the gap between winding and upper/lower yoke, T_{b3} is the bobbin thickness at the top and bottom, and H_p is the height of the primary winding. The front view of the core is shown in Figure 7.9.

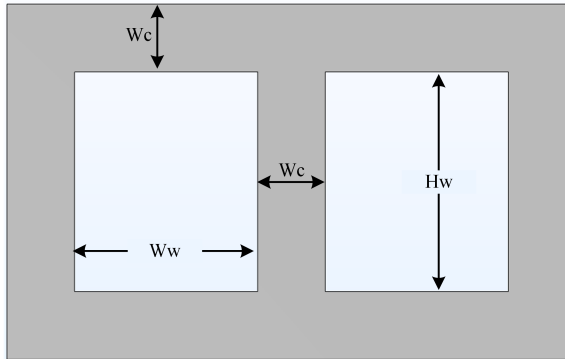


Figure 7.9: Front view of the core

From the window size and core width, the core volume can be calculated (a 0.99 scaling factor is used to compensate for rounded edges):

$$V_c = ((W_w \cdot 2 + W_c \cdot 3) \cdot (H_w + 2 \cdot W_c) - 2 \cdot (H_w \cdot W_w)) \cdot W_c \cdot 0.99 \quad (7.21)$$

And finally the core mass, using Metglas density $\rho_c = 7.3 \text{ g/cm}^3$ can be calculated as follows:

$$M_c = V_c \cdot \rho_c \quad (7.22)$$

The conductor length is approximated as the number of turns multiplied by the average length of turns in a winding, allowing to compute the mass of each conductor as follows:

$$M_p = 3 \cdot l_p \cdot W_{cp} \cdot H_{cp} \cdot \rho \quad (7.23)$$

$$M_s = 3 \cdot PPS \cdot l_s \cdot W_{cs} \cdot H_{cs} \cdot \rho \quad (7.24)$$

Where PPS is the number of parallel paths in the secondary. Therefore, the total mass of the transformer, excluding insulation and terminations, is as follows:

$$M_t = M_c + M_p + M_s \quad (7.25)$$

The next step is to use COMSOL Livelink to MATLAB to perform a FEM simulation, computing the AC losses at 1kHz for both conductors. Once the conductor ac resistance is found, the losses are calculated for primary and secondary windings at 1kHz as follows:

$$P_p = 3 \cdot R_p \cdot I_p^2 \quad (7.26)$$

$$P_s = 3 \cdot R_s \cdot I_s^2 \quad (7.27)$$

Core loss is estimated analytically as follows [4]:

$$P_c = W_c \cdot P_{c0} \cdot \left(\frac{f}{f_0}\right)^\alpha \cdot \left(\frac{B_m}{B_{m0}}\right)^\beta \cdot \left(\frac{k_f}{k_{f0}}\right)^{2(\alpha-1)} \quad (7.28)$$

The transformer total efficiency then is computed as follows:

$$\eta = \frac{25 \text{ kW}}{25 \text{ kW} + P_c + P_p + P_s} \quad (7.29)$$

7.2.4. DESIGN SELECTION

The results of the design sweep are shown in Figure 7.10. From the plot it can clearly be seen that the designs with 5 secondary turns and 15 or 17 parallel paths on the secondary winding seem to provide the highest efficiency. The trend noticed before during the initial analysis was confirmed, showing significant mass reduction for 5 secondary turns compared to designs with 4 turns.

However, leakage inductance should be ensured within the required range (7.7 mH for the outer phases and 10.2 mH for the inner phase). In order to verify this, the designs around the knee point (35–41 kg and above 97.5% efficiency) are selected, and a 3D COMSOL simulation is performed as described in Section 4.2.3 for each design. The lighter design that meets the requirement is **ID365**. Additionally, a heavier design can be identified as **ID122**, as both possess identical leakage inductance (7.7 mH for the side phases and 10.2 mH for the central one). Table 7.4 provides a comparison between the two possible options.

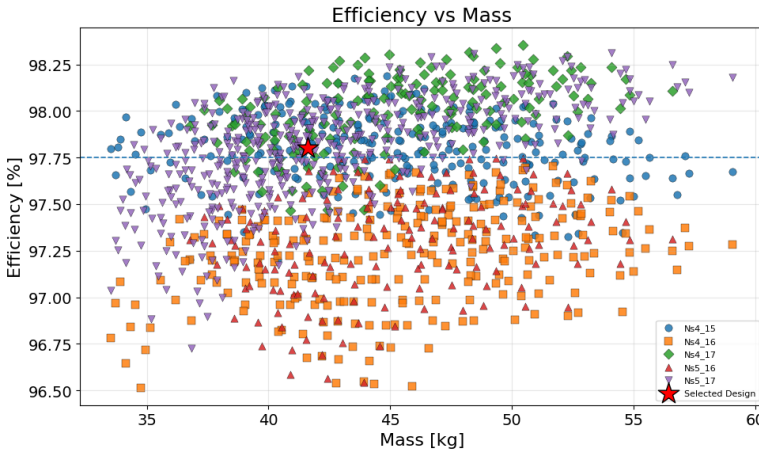


Figure 7.10: Optimization results for transformer design: mass vs efficiency.

The key distinction is the conductor material: **ID365** uses aluminum, achieving 97.5% efficiency at a mass of 38.4 kg, and **ID122** uses copper, with 97.8% efficiency at 41.6 kg. While the aluminum design appears more favorable in terms of mass, its integration poses several practical issues due to material mismatch with the copper-based PECs and resonant converter:

- Galvanic corrosion: electrical contact between copper and aluminum in the presence of moisture creates a galvanic cell, leading to accelerated corrosion of the aluminum;

Table 7.4: Comparison table between selected designs

Parameter	ID122	ID365
N_p	5	5
N_s	231	231
Core Width	54 mm	54 mm
Primary conductor	2 × 1 mm	2 × 1 mm
Secondary conductor	2 × 2.5 mm	2 × 2.5 mm
Parallel paths	17	17
Gap between disks	6.4 mm	6.4 mm
Primary to core distance	37.5 mm	37.5 mm
Conductor material	Cu	Al
Total mass	41.6	38.4
Efficiency	97.8%	97.5%

- Thermal expansion: differing thermal expansion coefficients can degrade the quality of mechanical and electrical contacts under temperature fluctuations;
- Oxidation: aluminum forms a non-conductive oxide layer, which increases contact resistance, whereas copper oxide remains conductive and less problematic.

These issues can reduce the quality of the electrical connection and can result in increased losses, heating, and arcing. While compensation methods exist—such as bi-metallic connectors and antioxidant compounds—the increased complexity does not present an appealing trade-off. This is especially relevant given that, at below 50 kg, the SST solution is considerably lighter than the average LLC transformer (typically 100–300 kg for a 25 kW unit, depending on insulation type, ratings, and materials used). With these considerations in mind, **design 122** is chosen. The core and windings are illustrated in Figure 7.11. The parameters of the final selected design are summarized in Table 7.5.

7.3. SIMULATIONS

With the design established, detailed simulations can now be carried out to evaluate the electric and magnetic field distributions within the MFT, ensuring that issues such as insulation breakdown, core saturation, or other unwanted effects do not arise. These simulations aim to verify the absence of insulation breakdown, core saturation, and other undesirable phenomena that could compromise performance.

7.3.1. ELECTRIC AND MAGNETIC FIELD SIMULATIONS

To evaluate the magnetic field distribution within the core, a 2D axisymmetric finite element model is implemented using COMSOL Multiphysics. The simulation is performed at an operating frequency

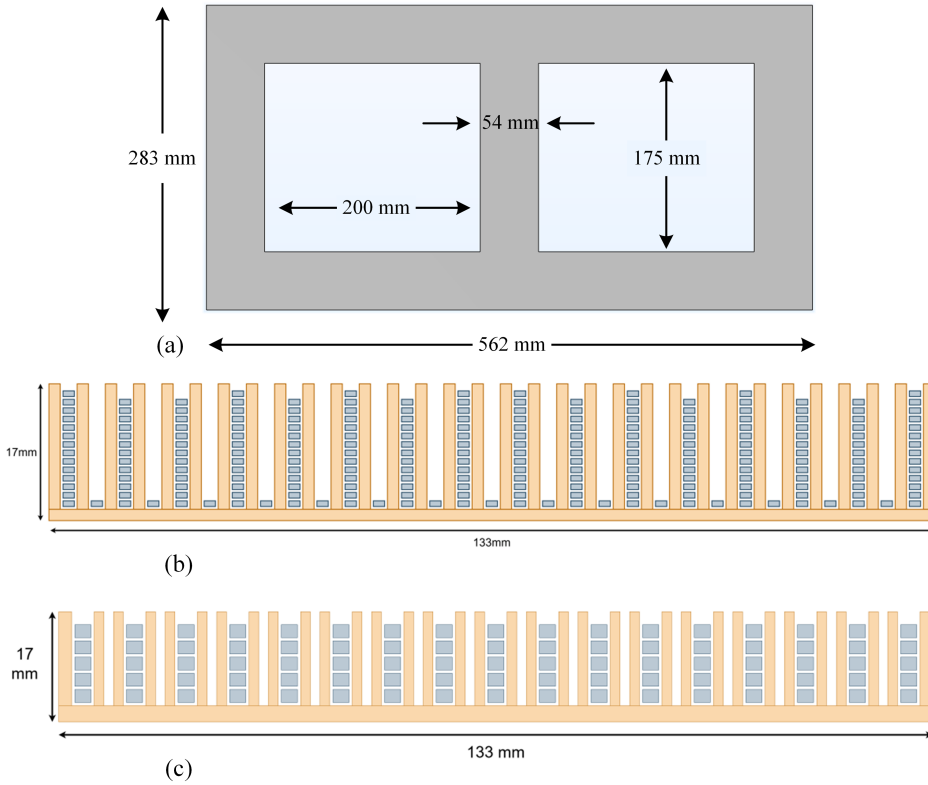


Figure 7.11: The schematic of the demonstrator transformer: a) the core, b) the primary winding, and c) the secondary winding.

of 1000 Hz and focuses on a single transformer phase, including the polycarbonate (PC) bobbins, to reduce computational complexity. The simulation geometry and meshing strategy are shown in Figures 7.12.a and b.

The worst-case conditions are expected to occur at the region of minimum distance between the core leg and the surrounding structure, particularly at the 45° angular position (see Figure 7.6). Due to the axisymmetric nature of the simulation, the rectangular core is approximated as a cylindrical shape. The equivalent radius of this cylinder is computed using the core flux utilization factor k_f as:

$$R_{eq} = \frac{\sqrt{2}}{2} \cdot k_f \cdot W_c \quad (7.30)$$

To preserve the magnetic path length, the height of the core leg is adjusted according to:

Table 7.5: Summary table final design

Parameter	Description	Value
Np	primary coil turns	231
Ns	secondary coil turns	5
Wp	primary coil radius	15.4 mm
Ws	secondary coil radius	13.75 mm
wcs, hcs	size of secondary conductor	2×2.5 mm
wcp, hcp	size of primary conductor	1×2 mm
Hp, Hs	Height of coil	128 mm
PPs	Parallel Paths secondary	17
NT_dp	Number of type 1 disks	14
Ndp	Number primary disks	31
MLTp1	mean length of turn (primary) type 1 disks	429 mm
MLTp2	mean length of turn (primary) type 2 disks	383.8 mm
MLTp2	mean length of turn (secondary)	674.5 mm
tb1, tb2, tb3	Size of PC bobbin	2.4, 2.1, 2.5 mm
Ww	core window width	198.7 mm
Hw	core window height	173 mm
Gwy	Gap between windings and yoke	20 mm
Gww	Gap between phases	20 mm
Gps	Gap primary to secondary	20 mm
Gpc	Gap primary to core	20 mm
Gddp	Gaps between primary disks	6.4 mm
Wc	Core width	54 mm
Rc	Equivalent core radius	30.4 mm
Vc	Core volume	0.0047 m ³
Mp	Mass of primary coil	5.43 kg
Ms	Mass of secondary coil	7.65 kg
Mc	Mass of core	28.54 kg
Mtr	Mass of total transformer	41.6 kg
Pp	Primary load losses	190.5 W
Ps	Secondary load losses	190.8 W
Pc	Core losses	186 W
Ptr	Total transformer losses	568.3 W

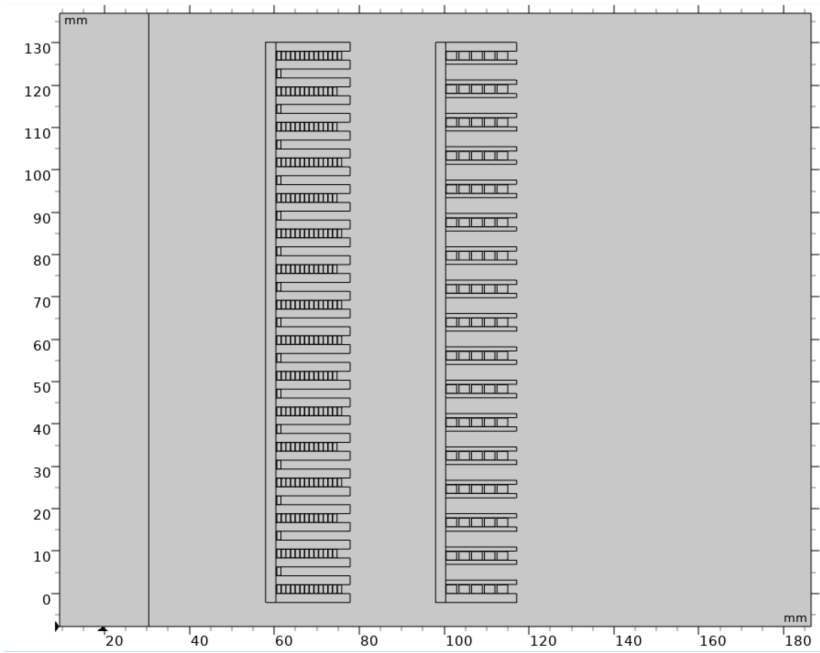
$$H_c = 1.5 \cdot (H_w + W_c) + W_w + W_c \quad (7.31)$$

where H_w and W_w are the window height and width, and W_c is the core width (Figure 7.13).

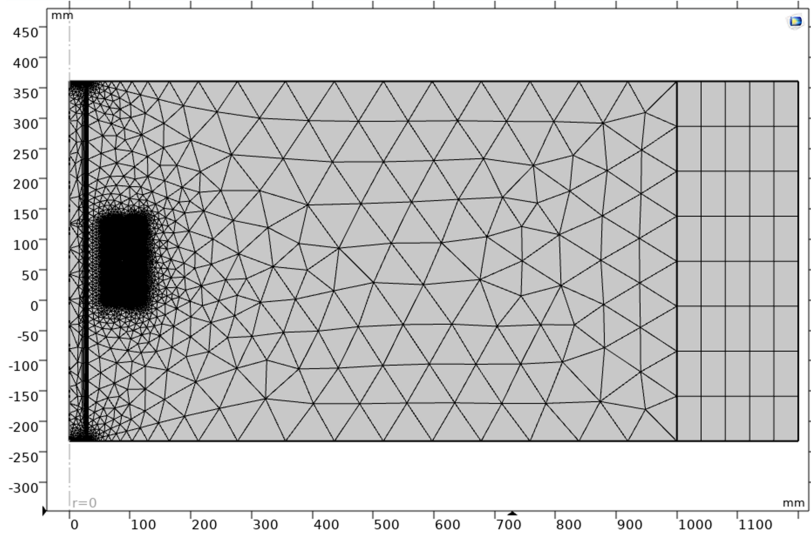
A periodic boundary condition is applied to the surrounding air domain to simulate an infinite repeating structure. This approach a closed-loop core in a 2D-Axisymmetric domain. The material properties used in the simulation are listed in Table 7.6. Notably, the Metglas core is modeled as a solid block with an equivalent electrical conductivity to account for lamination effects. The adjusted conductivity is calculated as:

$$\sigma_c = \frac{44 \times 10^6}{\left(\frac{R_{eq}}{w_l}\right)^2} \quad (7.32)$$

where $R_{eq} = 30.46$ mm and $w_l = 250$ μ m, yielding a conductivity of



(a)



(b)

Figure 7.12: COMSOL geometry and mesh for magnetic field simulation: a) the model, b) the mesh performance.

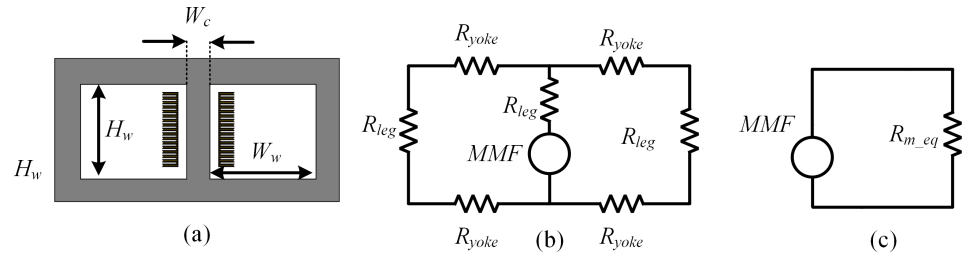


Figure 7.13: (a) Magnetic core geometry defining H_w , W_w , and W_c . (b) Corresponding magnetic reluctance network. (c) Equivalent magnetic circuit with lumped reluctance R_{m_eq} .

$\sigma_c = 2963 \text{ S/m}$.

Table 7.6: Material properties used in magnetic field simulation

Material	Rel. permeability μ_r	Conductivity $\sigma \text{ [S/m]}$	Rel. permittivity ϵ_r
Air	1	1×10^{-5}	1
Copper	1	4.2×10^7	∞
Metglas	4×10^3	2963	∞
Polycarbonate (PC)	1	1×10^{-13}	2.6

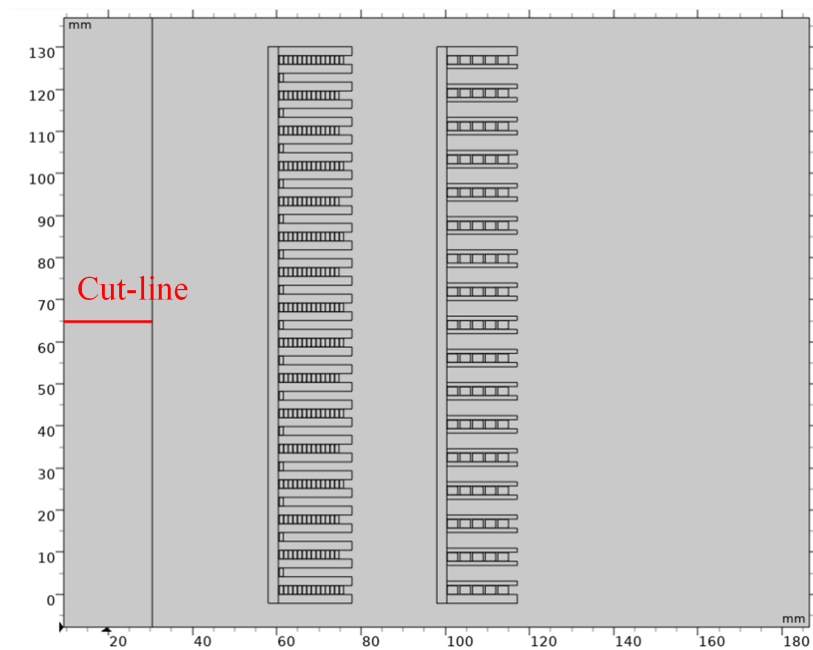
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The primary winding is defined using individual conductors and is excited with a peak voltage of 3776 V, while the secondary windings are left open-circuited. A frequency-domain study is conducted at 1000 Hz. The magnitude of the magnetic field within the core is evaluated at the height of the windings. Since the model treats the core as a solid block, the resulting field is scaled by a stacking factor of 0.79 to reflect the effect of lamination.

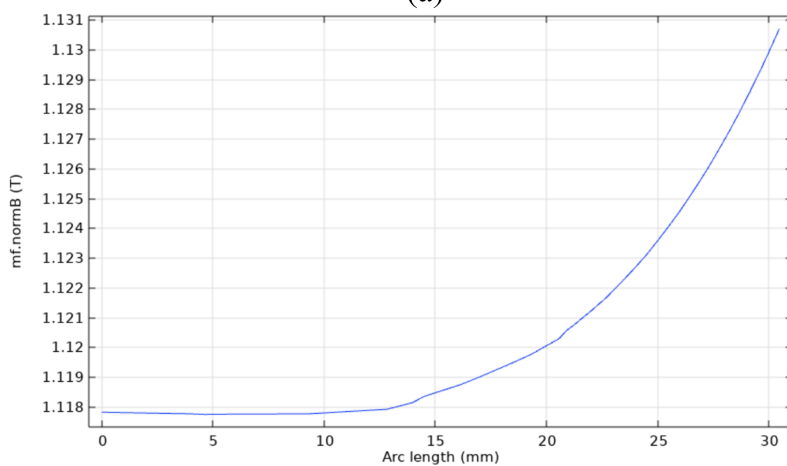
The resulting magnetic field distribution is shown in Figure 7.14. The peak magnetic flux density within the core reaches 1.13 T, remaining below the saturation threshold and providing a 9.6% margin for overvoltage or transient conditions.

To further complement this, the electric field distribution within the transformer air gaps is also investigated. For this simulation, a single window and phase are modeled using a 2D electrostatics study, as illustrated in Figure 7.15. The study evaluates steady-state field conditions under worst-case voltage stress, corresponding to the peak primary-side excitation.

The primary winding is modeled with a linear voltage distribution across its turns, assuming a uniform potential gradient of V_p/N_p , where V_p is the peak voltage and N_p the number of primary turns. The secondary windings and the core are held at ground potential. The



(a)



(b)

Figure 7.14: a) COMSOL geometry for magnetic fields simulation - primary and secondary windings. Red line shows the cut-line, b) Magnetic field intensity within the core leg, measured along the cut-line.

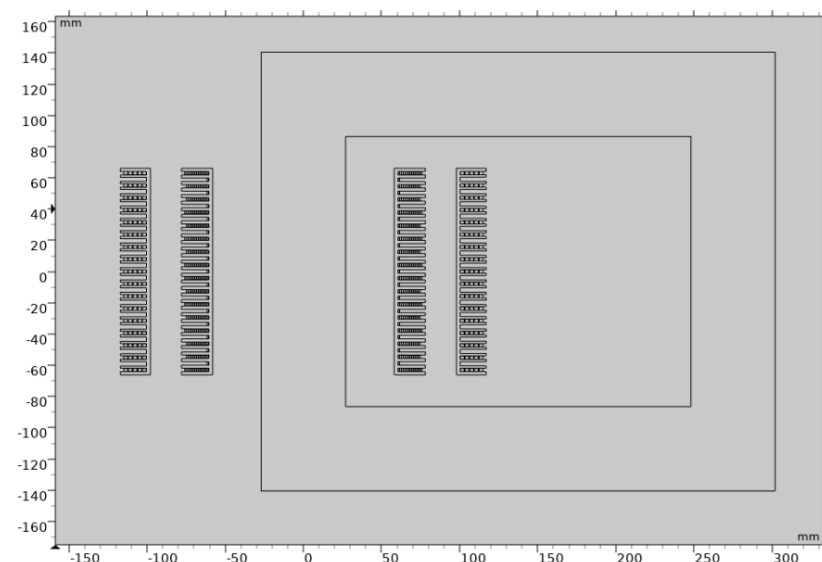


Figure 7.15: COMSOL geometry for electric field simulation for a single window and phase.

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simulation results are shown in Figure 7.16. Across the majority of the domain, the electric field magnitude remains below 0.4 kV/mm. A localized peak of approximately 1.8 kV/mm is observed near sharp conductor edges.

While this value is above the dielectric strength threshold, the simulation neglects enamel insulation on the wires. In order to verify this, partial discharge testing is performed on a first prototype, discussed in Chapter 4. In case of discharge behavior due to local field enhancements, resin insulation can be applied on the high-voltage coils.

The selected material is Araldite CY-225, mixed in a 1:1 ratio with HY-225 hardener, which possesses a relative permittivity of $\epsilon_r = 4$. The high-voltage coils are embedded in a rectangular cross-section of epoxy resin, with each side extending approximately 3–5 mm beyond the outer boundary of the windings and the polycarbonate bobbin. Simulation results indicate that the addition of the epoxy layer reduces the peak electric field at sharp conductor edges to approximately 0.4 kV/mm, below the breakdown strength of the solid-air interface.

The electric field between turns and between turn and core is measured in Figure 7.16.

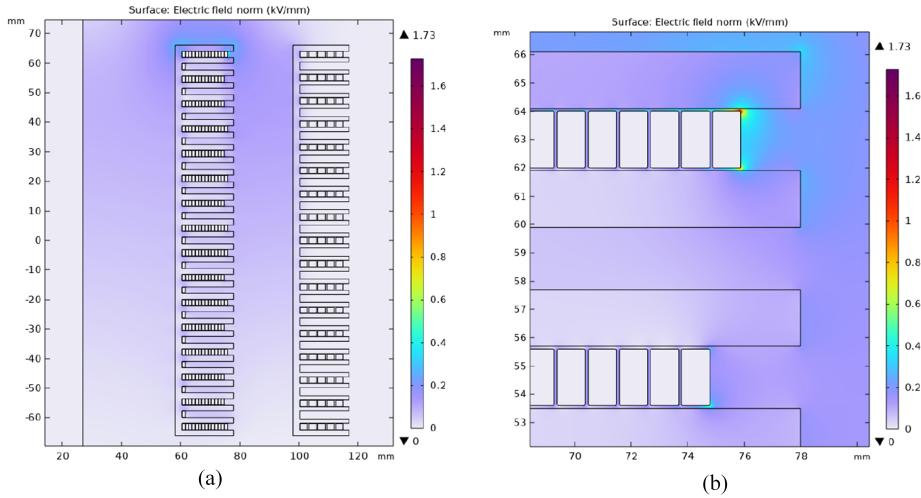


Figure 7.16: Electric field simulation: (a) around the primary coil, (b) around sharp points.

7.3.2. BUSBAR LOSSES SIMULATION

The skin depth in copper at 1 kHz was calculated to be approximately 2 mm. In order to avoid additional losses, a $3 \times 25 \times 150$ mm copper bus bar is selected to assemble the termination, ensuring even current distribution. Bolts are used to tighten two bars together and maintain the connection quality between the secondary parallel coils and the termination busses.

To estimate the power losses at the secondary winding terminations, a single terminal is modeled in COMSOL Multiphysics, as illustrated in Figure 7.18. In the simulation, a 1 V excitation is applied across the parallel conductor paths, with the opposing terminal (top of the bar) grounded. The study is performed in the frequency domain at 1 kHz to account for skin and proximity effects.

The resulting AC resistance of a single terminal is calculated as 0.023 mΩ. Since each phase termination involves two such terminals in parallel, the total resistance per phase termination is 0.046 mΩ. Given a secondary current of 200 A, the corresponding power loss per phase termination is:

$$P_{\text{loss}} = I^2 R = 200^2 \times 0.046 \times 10^{-3} = 1.84 \text{ W} \quad (7.33)$$

Since the core and conduction losses are estimated at 800 W from analytical calculations, the termination losses can be neglected for efficiency and thermal profile modeling.

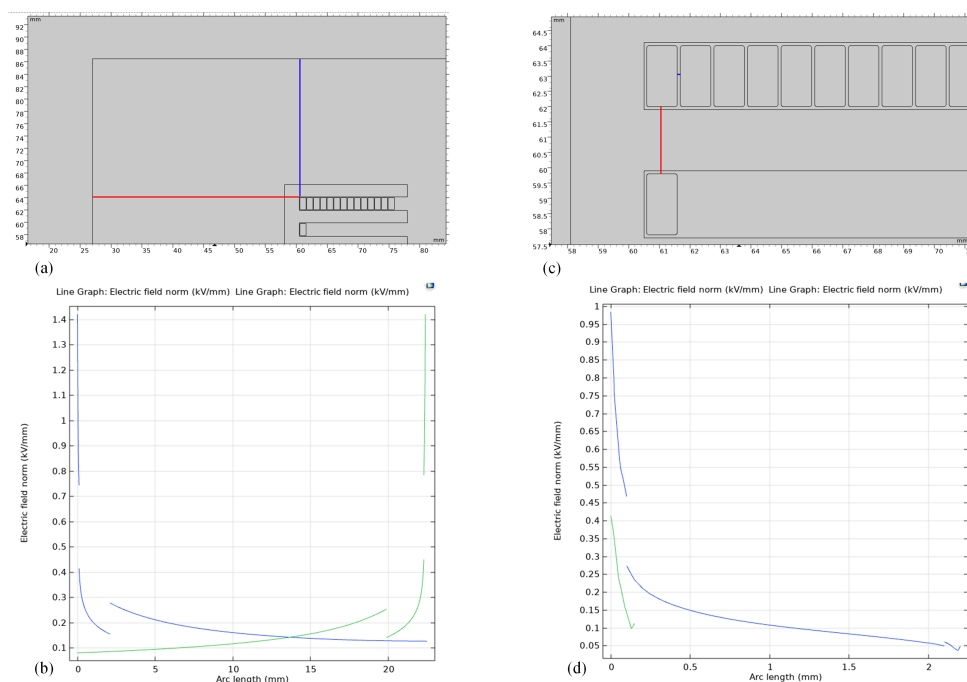


Figure 7.17: a) Cut-lines used for turn to leg (red) and turn to yoke (blue), b) Electric field magnitude between first turn and core leg (blue) and yoke (green), c) Cut-line used for turn-to-turn (blue) and turn-to-disk (red) electric field simulations, and d) Electric field magnitude between two turns in disk (blue) and between two disks (green).

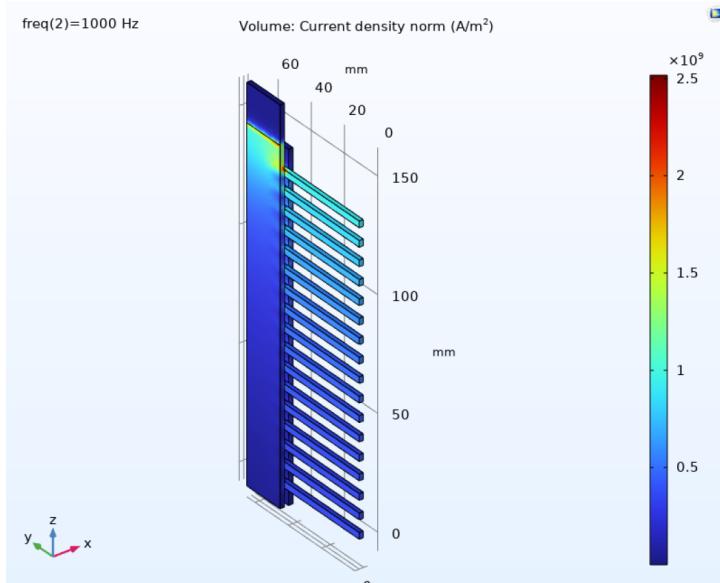


Figure 7.18: Termination geometry and current density when 1 V is applied to the parallel wires.

7.4. FABRICATION

The fabrication process described in this section consists of two parts. First, the fabrication of the MFT is detailed and then, implementation of the resonant tank is presented.

7.4.1. MFT FABRICATION

The first step in the fabrication process is the 3D printing of the polycarbonate bobbins for support of the conductors. In order to meet supplier availability, the secondary conductor size was changed to 2.8×1.8 mm. Since the surface area remains unchanged, the only modification required was a slight enlargement of the bobbin guides.

Once the bobbins are printed, the conductor is wound around the guides. In order to maintain tension on the wire and ensure minimal spacing between turns—where high electric fields could arise—the bobbin is placed above the copper wire bobbin, as shown in Figure 7.19. While rotating the PC bobbin and winding the conductor, tension is maintained using the weight of the copper itself.

Terminations are then added to the secondary coils, connecting each parallel path using a 25×3 mm copper bus bar. Special attention is required to avoid external particles from the workshop environment polluting the assembly, as these imperfections can cause local increases



Figure 7.19: Coil assembly process: a) bobbin 3D printing, b) printed primary bobbin, c) the winding procedure, d) close up of the conductors, e) close up of completed primary, f) close up of secondary coils with termination

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in electric field and degrade performance, possibly resulting in local discharges or breakdown.

Having verified the fabrication, simulation coordination, and thermal behavior of the single-phase prototype, the complete three-phase assembly with the core can be completed. Due to manufacturer availability and project schedule, the desired Metglas core could not be sourced. In order to perform loading and insulation testing for a three-phase design, an alternative material solution is necessary to provide magnetic coupling of the coils.

Two material options are available: H085-23 (Grain-Oriented Electrical Steel: GOES) [5], and M400-50A (Non-Grain-Oriented: Non-GOES) [6]. Table 7.7 lists the material properties provided by the manufacturer, showing how the grain-oriented material results in lower losses due to the molecular arrangement. This makes GOES ideal for magnetic cores in transformers, where the magnetic flux is predictable and aligned with the crystalline structure. On the other hand, non-grain-oriented steel possesses random orientation of the material crystals, resulting in higher losses, especially at higher frequencies—but with reduced material cost.

While the performance of GOES and Non-GOES at higher frequencies is reduced compared to amorphous materials such as Metglas, electrical

Table 7.7: Core material properties

Material	Lamination thickness	Density	Max core losses at 1.5T	Min mag. polarization
M400-50A	0.5 mm	7.7 kg/dm ³	4.0 W/kg	1.53T at 2500 A/m
H085-23	0.23 mm	NA	0.6 W/kg	1.90T at 900 A/m

steel cores are a more economical alternative that can be effectively used for insulation testing [2]. The steel core assembly can then be used for three-phase PD testing, applied voltage testing, and other loading tests, as suggested in the next chapter. A study can then be performed comparing the effectiveness and suitability of GOES and Non-GOES magnetic cores for medium-frequency applications.

The two core options are manufactured by Decoup S.A.S. The manufacturing process involves casting the aforementioned material, which is then cut into strip plates. The composition and dimensions of each strip are shown in Figure 7.20. Through-hole screws are used to assemble the structure and maintain pressure across the sheets, minimizing the inter-lamination gap/interface. The dimensions of each strip are as follows: During the assembly process, the phase coils are inserted onto the construction, and insulation distances are maintained using polycarbonate spacers.

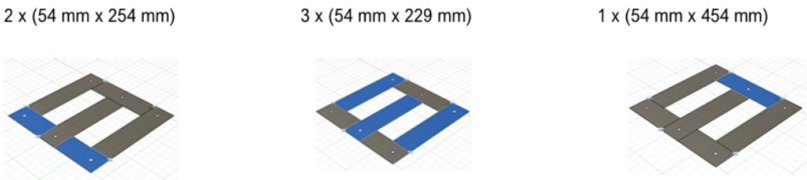


Figure 7.20: Schematic of the steel strips used in the core assembly.

The transformer fabrication process is shown in Figure 7.21. The assembly begins by preparing a wooden frame to hold the components in place during construction. The preformed modular winding segments are then positioned on the frame in their designated locations. The winding terminals are aligned and connected according to the required electrical configuration. Once the windings are secured, the laminated magnetic core segments are inserted through the coil openings to complete the magnetic path. The structure is finally tightened to ensure proper mechanical stability and electrical insulation before further testing.

An alternative assembly approach is illustrated in Figure 7.22. In this method, the magnetic core is first assembled as a complete structure. The upper yoke is then removed to create sufficient clearance for

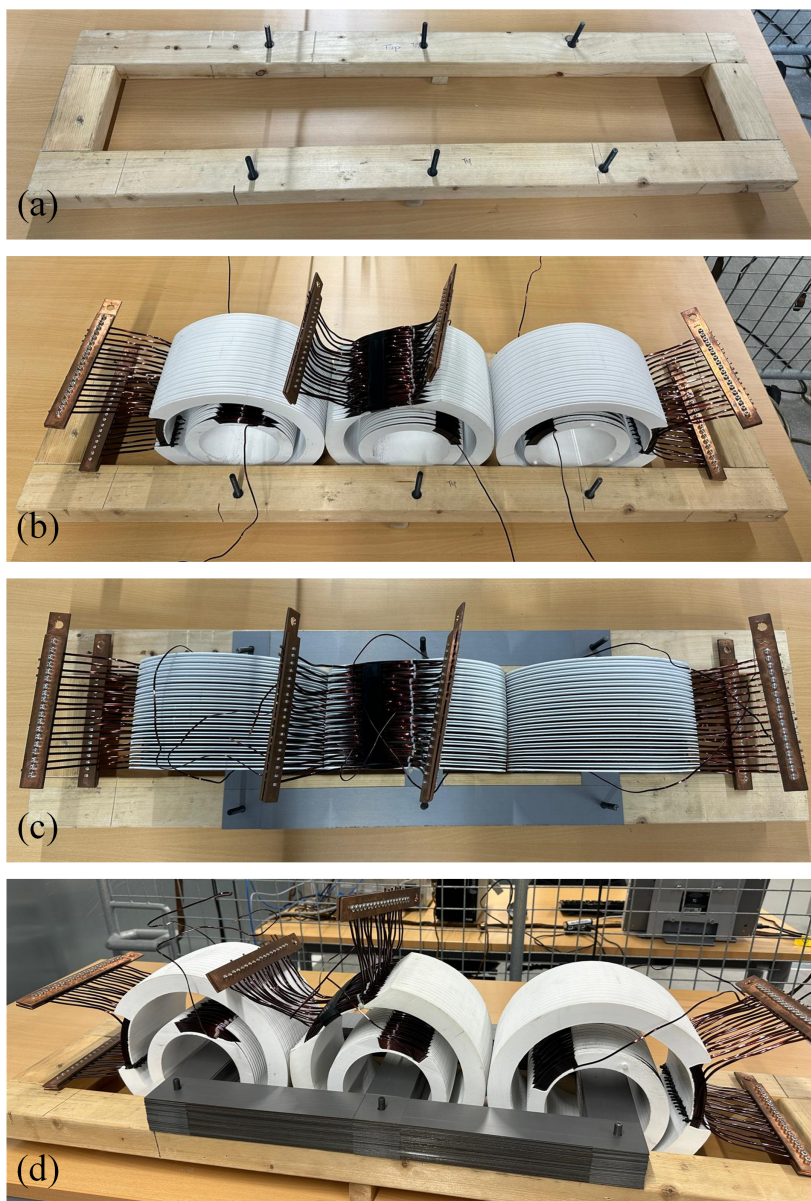


Figure 7.21: Transformer fabrication process: (a) wooden assembly frame, (b) placement of modular winding segments, (c) alignment and interconnection of windings, and (d) insertion of the magnetic core.

inserting the winding modules. After the windings are positioned and aligned, the upper yoke is reinstalled piece by piece to restore the magnetic path. This procedure simplifies the handling of tightly fitted windings and allows for better control of the mechanical tolerances during insertion. The assembled MFT is shown in Figure 7.23.



Figure 7.22: Alternative assembly method where the magnetic core is first fully assembled, and: a) the upper yoke is removed to allow winding insertion; (b) after positioning the windings, the upper yoke is reinstalled piece by piece to complete the magnetic path.



Figure 7.23: Assembled medium-frequency transformer (MFT) with winding modules secured and core clamped in place.

7.4.2. RESONANT TANK FABRICATION

The fabrication of resonant tank depends on the parasitic parameters of the transformer. Since the MMR SST is a LLC resonant converter, the value of the leakage inductance and magnetizing inductance of each phase must be measured to select suitable capacitor. The resonant frequency, leakage inductance, and series capacitance are related as follows:

$$f_r = \frac{1}{2\pi \times \sqrt{L_s \times C_s}} \quad (7.34)$$

where L_s and C_s are series inductance and capacitance, respectively. The measured parameters of the fabricated MFT are summarized in Table 7.8. The difference between the inner and outer phases is mainly caused by asymmetric magnetic paths in a three-phase core-type transformer. The middle phase has a symmetric main flux return path and confined leakage flux, whereas the side phases experience longer, asymmetric return paths and less constrained leakage-flux closure, leading to different magnetizing and leakage inductances.

Table 7.8: Characteristics of MFT at 1 kHz

Parameter	Phase 1	Phase 2	Phase 3
Magnetizing Inductance (mH)	165.7	254.9	170.7
Leakage Inductance (mH)	8	10.2	7.7

7.5. TEST

Before proceeding to the complete assembly, in order to ensure no defects are already present in the fabricated phase, preliminary testing is performed on this prototype as a single phase air core transformer. Once proper design and fabrication is verified, then the other phases and the magnetic core can be all assembled into the final design.

7.5.1. BODE ANALYSIS

A Bode analysis is performed, measuring parameters such as winding resistance, inductance, and gain over the frequency spectrum. The measurement is then compared with the expected values from FEM simulations. The setup is shown in Figure 7.24, using an Omicron Lab Bode 100 in a one-port configuration.

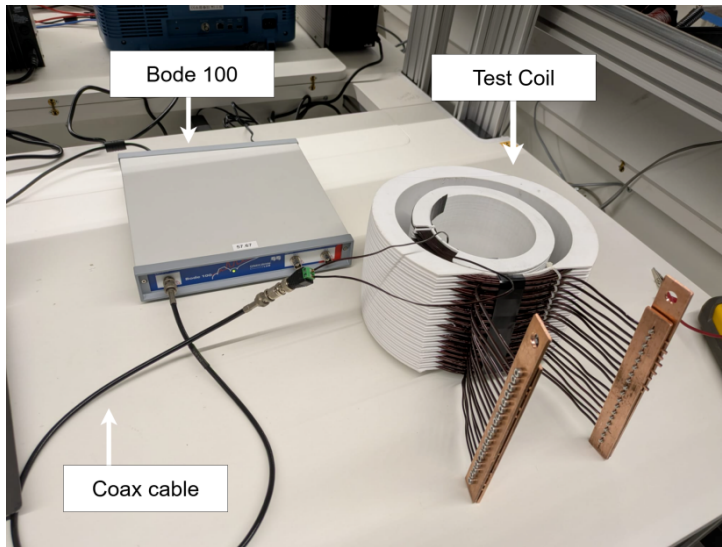


Figure 7.24: Measurement setup showing test transformer and Bode 100, connected in one-port configuration via coaxial cable.

The relevant measurement results are presented in Figure 7.25 and summarized in Table 7.9, where they are compared to the values obtained from electromagnetic simulations. As shown, the majority of parameters fall within a 10% margin of the expectations, confirming the validity of the design and fabrication.

An exception is observed in the measured secondary winding resistance, which deviates substantially from the simulated value. This discrepancy is attributed to the extremely low magnitude of

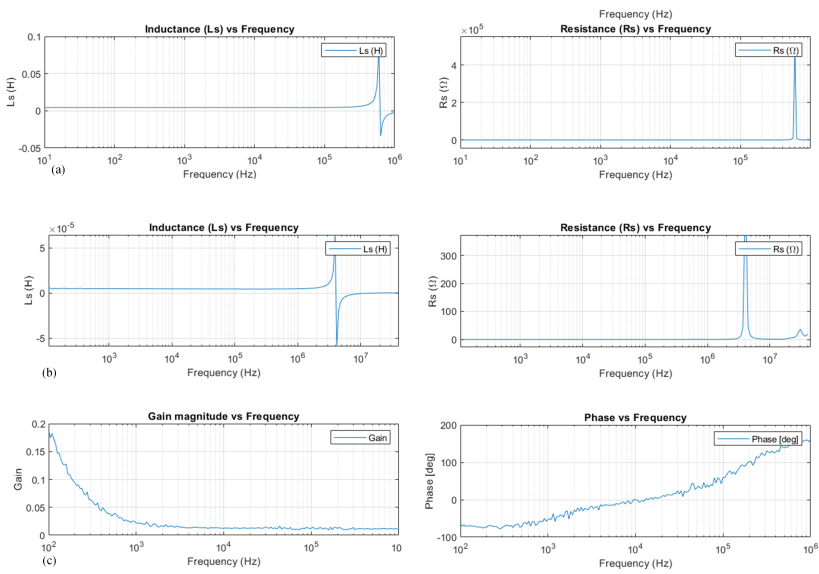


Figure 7.25: Bode measurements results for test phase: a) Primary resistance and inductance over frequency, b) Secondary resistance and inductance over frequency, c) Transformer gain and phase over frequency.

Table 7.9: Bode measurements for test phase at 1 kHz

Parameter	Primary		Secondary	
	Measured	FEM	Measured	FEM
R_{DC}	1.51 Ω	1.63 Ω	11.7 m Ω	0.95 m Ω
R_{AC}	1.26 Ω	1.53 Ω	41 m Ω	1.59 m Ω
L_{lk}	3.20 mH	3.07 mH	3.54 μ H	3.11 μ H
L_{mag}	1.34 mH	1.58 mH	1.41 μ H	1.42 μ H
Gain	21.50 m	21.64 m	46.5 μ H	46.2 μ H

the resistance (on the order of 1 m Ω), making the measurement highly sensitive to interconnection effects, including contact resistance between the coil terminations and the probe. Despite this, the results support the successful and accurate construction of both test windings. Additionally, the Bode plots in Figure 7.25 reveal that the self-resonant frequency of the winding occurs at approximately 800 kHz.

The self-resonant frequency around 800 kHz confirms that parasitic winding capacitances become relevant only far above the 1 kHz operating frequency. Consequently, capacitive effects have a negligible influence on the transformer behavior in the frequency range of interest.

7.5.2. PARTIAL DISCHARGES MEASUREMENT

Partial discharge (PD) refers to localized electrical discharges that only partially bridge the insulation between electrodes and typically occur at points where the electric field is intensified, such as voids, interfaces, or sharp edges within the insulation. Although initially small, PD can gradually deteriorate insulation materials, leading to insulation failure if left undetected.

Partial discharge is measured in pC; IEC 60076-11 dictates a maximum of 10 pC in its dry-type transformer standard. Standard IEC 60270 covers partial discharge measurements for high-voltage systems [7]. Figure 7.26 shows the test circuit: a high-voltage AC source energizes the primary winding of the transformer under test through a series resistor (R), which limits inrush currents and suppresses transients.

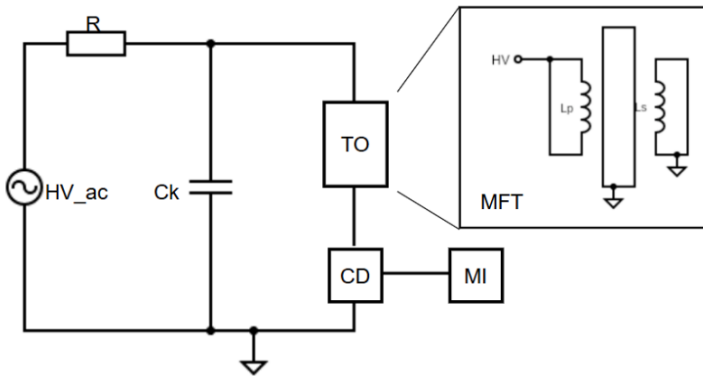


Figure 7.26: PD test setup diagram, showing the Test Object (TO), Coupling Device (CD) and measuring Instrument (MI). The primary coil is connected to the HV source at both connections, ensuring no turn-to-turn voltage. Secondary winding and core are connected to ground.

The partial discharge measurements presented in this section are performed according to IEC 60270 using a sinusoidal voltage waveform at line frequency (50 Hz). While this testing approach is widely adopted for conventional transformers, it does not fully replicate the electrical stress conditions present in medium-frequency SST operation, where the insulation is subjected to non-sinusoidal voltages with higher fundamental frequencies and fast switching transitions.

Therefore, the reported PD results should be interpreted as an indication of insulation integrity and manufacturing quality under standard test conditions. The PD performance under actual converter-driven excitation may differ due to the influence of waveform shape, dv/dt , and operating frequency. A detailed investigation of PD behavior

under representative SST excitation is beyond the scope of this work and is recommended for future research.

Both HV terminals of the transformer are energized at the same voltage, ensuring that no turn-to-turn voltage is present and focusing the test on the primary-to-secondary and primary-to-core insulation. To test the latter, a mock-up core of equivalent dimension is used, emulating the presence of the Metglas core. A coupling capacitor (C_k) is connected in parallel to provide a low-impedance path for high-frequency PD pulses while isolating the measurement system from the high test voltage.

The PD pulses generated in the transformer's insulation system are extracted via a coupling device (CD), typically a high-frequency measuring impedance, and fed into a measuring instrument (MI) for analysis. The secondary winding and core are grounded to prevent undesired induction. A picture of the physical setup is also provided in Figure 7.27.

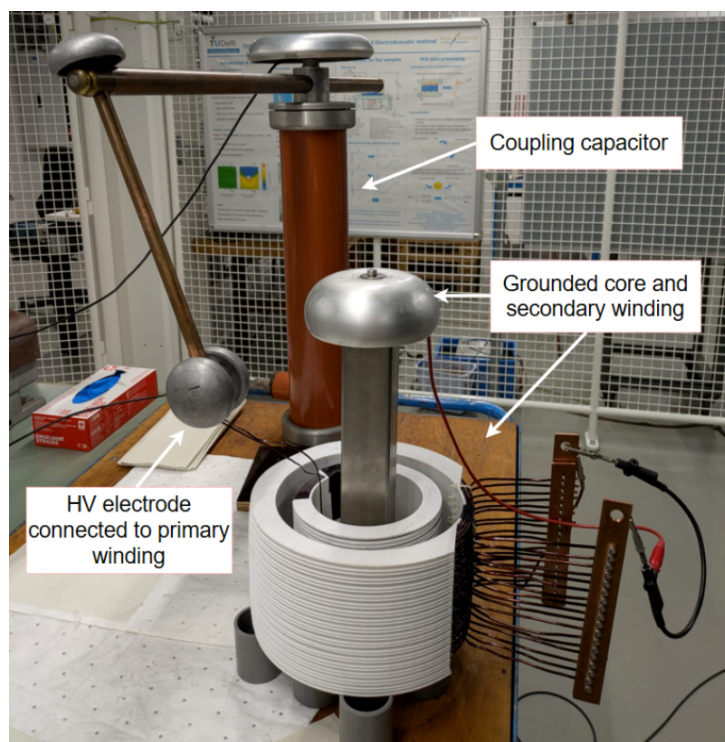


Figure 7.27: Partial discharge physical setup.

The applied voltage waveform, together with the resulting PD measurement, is shown in Figure 7.28. The applied voltage is initially

raised to $1.8 \times U_r$, with U_r indicating the rated rms voltage of the test object, for 30 seconds, after which the voltage is reduced to $1.3 \times U_r$ and maintained at this level for 180 seconds, before reducing it to zero. Throughout the test, the PD level is monitored. Since the level remains well below 10 pC, the test is passed, confirming that the air (and wire enamel) insulation is sufficient to withstand the applied voltage even in case of temporary over-voltage.

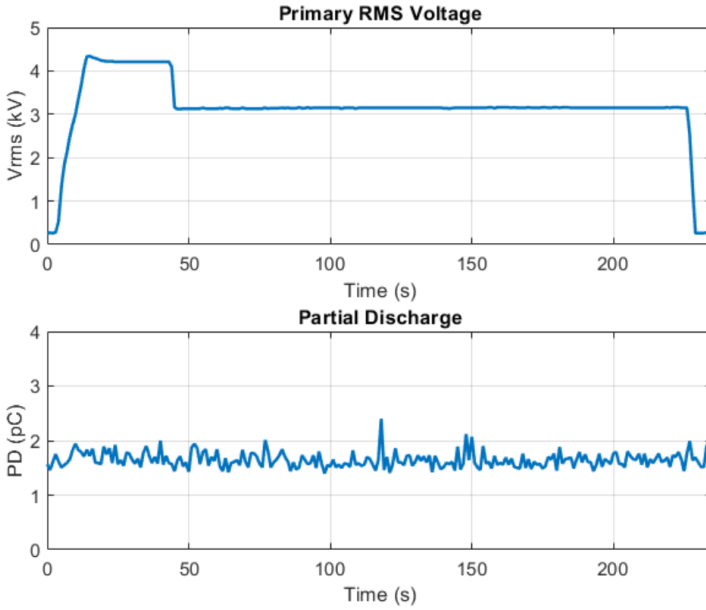


Figure 7.28: PD test result, applied voltage (top), PD discharge (bottom).

7.5.3. INSULATION BREAKDOWN TEST

To determine the dielectric failure threshold of the insulation system, an insulation breakdown test is conducted. According to IEC 60076-11, for transformers rated below 3.6 kV, the insulation is required to withstand at least 10 kV. In this test, high voltage is gradually increased until dielectric breakdown occurs. This event is typically characterized by a sudden surge in current, arcing, and a collapse of the voltage waveform. The voltage at which breakdown occurs represents the maximum RMS voltage the insulation can withstand under ideal conditions.

Due to the destructive nature of this test, it is performed on a scaled-down prototype rather than the full transformer coil, in order to minimize material costs while still capturing the relevant failure modes. The experimental setup and sacrificial test sample are shown in

Figure 7.29, and the test results are presented in Figure 7.30.

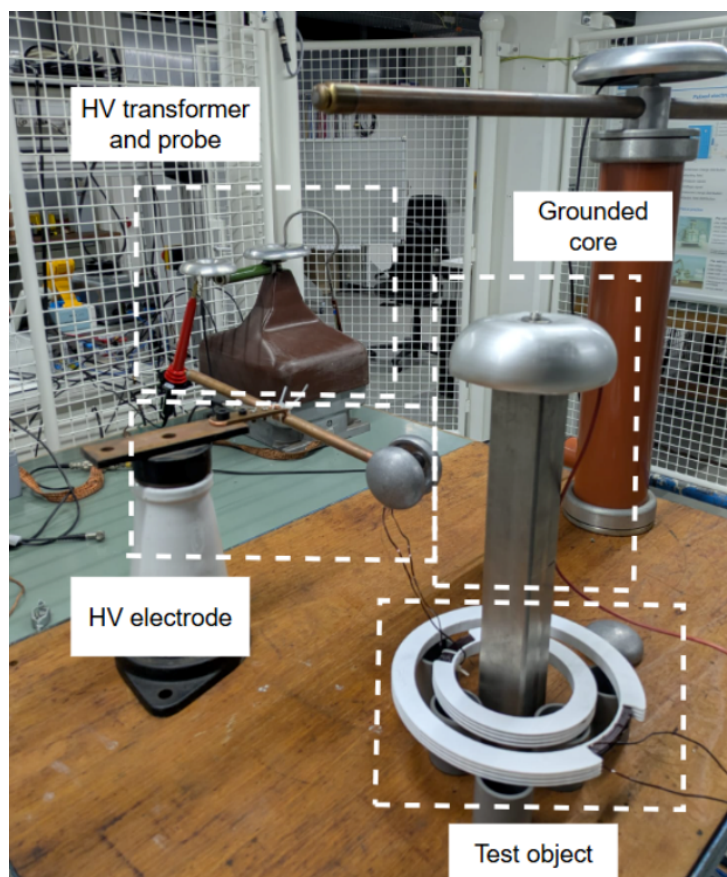


Figure 7.29: Insulation Breakdown Test laboratory setup.

During the first trial, breakdown occurred at $19.7 \text{ kV}_{\text{rms}}$, evidenced by a clear flashover and corresponding voltage collapse. In a subsequent test on the same sample, the breakdown voltage dropped to 17.3 kV . This reduction is attributed to localized degradation of the insulation from the initial breakdown event—burnt residues, conductive particles, and surface tracking reduce the dielectric strength and facilitate earlier discharge in subsequent tests. Since the insulation failure level is above the 10 kV required by IEC standards, it can be concluded that the insulation design of the prototype is sufficient.

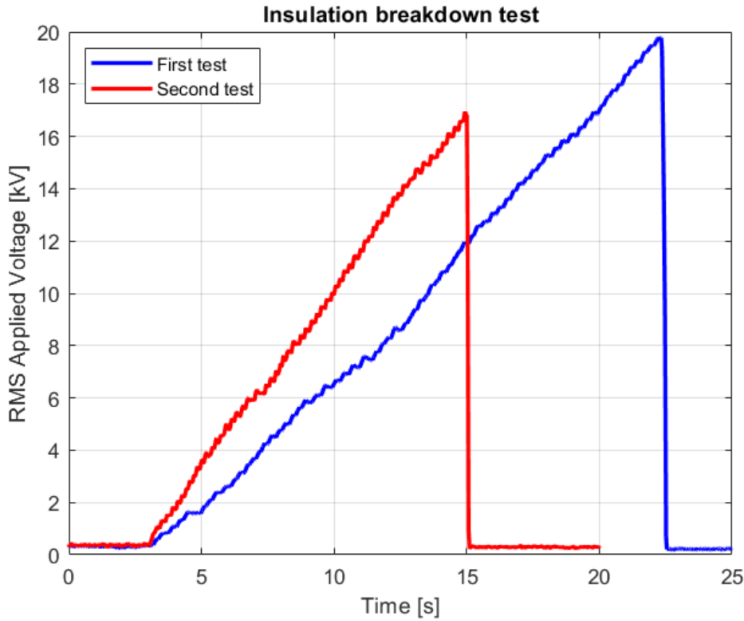


Figure 7.30: Insulation Breakdown Test laboratory setup.

7.5.4. THERMAL RISE TEST

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In order to validate the previously mentioned thermal simulations, a thermal rise test is performed on a single primary coil as outlined by the IEC standards. The coil is connected to a DC source, with a current output equal to the coil's rated current (7 A). With a resistance of approximately $1\ \Omega$, 7.6 V is initially applied to maintain the rated current; throughout the test, as the temperature of the conductor rises, the voltage is gradually increased to compensate for the increase in resistance and maintain constant current.

Temperature is measured using a type-K thermocouple on each disk at intervals of 30 minutes until steady-state conditions are reached. While the thermocouple has a limited accuracy of 4%, it is sufficient to evaluate the temperature profile of the design. Figures 7.31, 7.32, and 7.33 show the physical setup and the measured current, voltage, and temperature readings throughout the test.

As can be seen from the temperature readings, the hottest disks settle at a steady temperature of approximately 80°C , corresponding to a temperature rise of around 60°C . This is within the maximum temperature requirements dictated in Table 2 of IEC 60076-11 for Class B insulation.

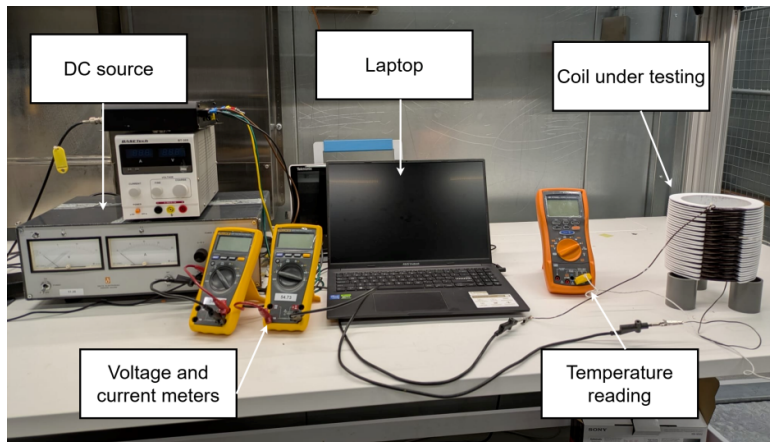


Figure 7.31: Thermal rise test setup.

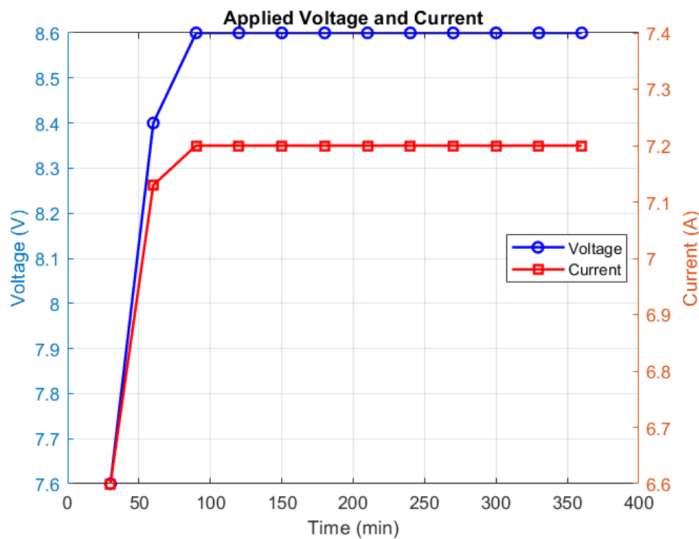


Figure 7.32: Applied voltage and current.

7.5.5. TRANSFORMER TESTS

SHORT-CIRCUIT TEST

As shown in Fig. 7.34, the setup of the short-circuit test consists of six parts as follows:

- **50 Hz grid** and protection,
- **50 Hz insulation transformer and Variac** to isolate the circuit and ramp up the applied voltage to the test circuit,

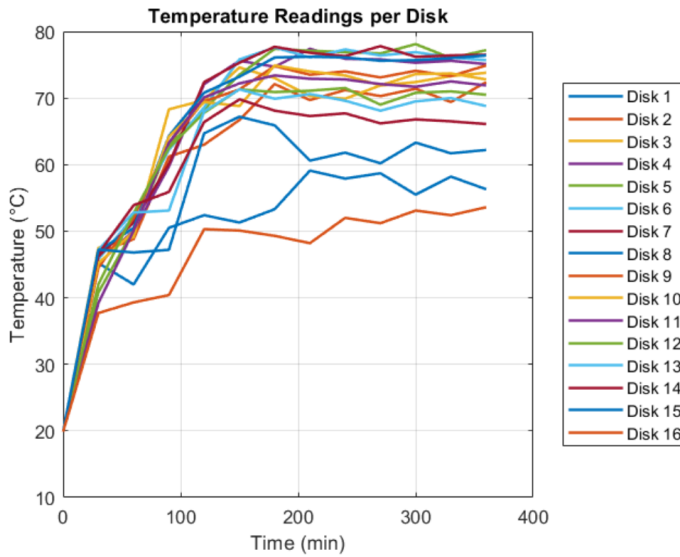


Figure 7.33: Temperature readings per disk.

- **Measurement equipment** including oscilloscope, current probes, differential voltage probes,
- **Compensation capacitors** to set the resonant frequency at 1 kHz for short-circuit test,
- **MFT** assembled as shown in Fig. 7.23,
- **Short-circuit connection** to provide low-impedance load for the MFT to mimic the real short-circuit situation.

7

The test setup is shown in Fig. 7.35. The frequency response of the transformer, measured by Bode 100 Vector Network Analyzer (VNA), in case of short-circuit is shown in Fig. 7.36. The impedance response shows a characteristic minimum around 1 kHz, corresponding to the short-circuit condition used for validation. At higher frequencies, the impedance increases due to leakage inductance, reaching a resonance peak near 120 kHz, which results from the interaction between the transformer leakage inductance and stray capacitances. To verify the analysis, especially for the compensation capacitors, a system-level Simulink simulation has been implemented as shown in Figure 7.37. The experimental results are shown in Fig. 7.38.

As can be seen from Fig. 7.38, the three phase currents are not perfectly identical. The small deviation in amplitude is consistent with the leakage inductance differences between the outer and central

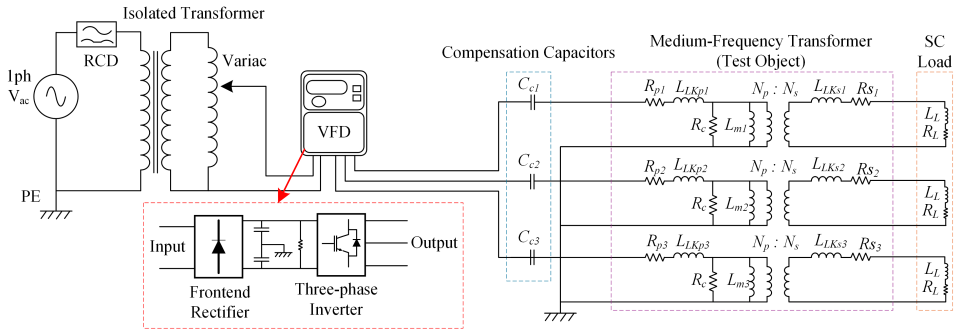


Figure 7.34: Experimental setup for three-phase excitation of the medium-frequency transformer (MFT) using a single-phase-fed VFD. Each phase is series-compensated with a tuned capacitor to achieve resonance at the operating frequency. Three high-value resistors (R_{VG}) form a virtual-neutral network connected at the transformer side of the capacitors to stabilize the floating star point without tripping the RCD protection.

phases obtained from the FEM analysis. In three-limb core-type transformers, the central leg naturally has a slightly different leakage path and mutual coupling than the outer legs, leading to a small dispersion in phase reactances.

7

7.6. CONCLUSIONS AND KEY FINDINGS

This chapter presented the complete prototyping and experimental workflow of the 30 kVA, 1 kHz three-phase MV-MFT developed for the SST demonstrator, including requirement definition, topology down-selection, multi-parameter optimization, fabrication, and test setup implementation. The main outcomes can be summarized as follows:

- A core-type configuration with cylindrical windings and rectangular conductors was selected as the most suitable trade-off between mass, practicality, and leakage inductance requirements for the demonstrator.
- A parametric sweep combining analytical sizing with COMSOL-based AC-loss extraction yielded a set of candidate designs, from which the final design (ID122) was selected as a practical solution, prioritizing reliable copper-based interconnections and integration with the resonant converter.
- Electromagnetic simulations confirmed that the peak magnetic flux density remains below saturation and that the electric-field stress

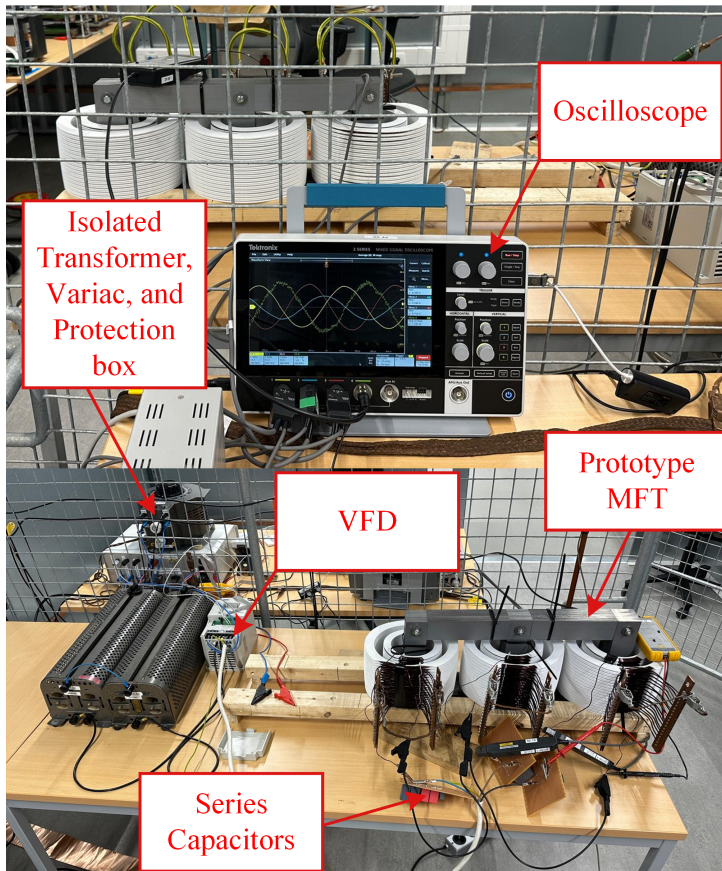


Figure 7.35: Laboratory setup for three-phase short-circuit testing of the prototype medium-frequency transformer at 1 kHz: (top) front view with oscilloscope displaying the measured phase currents and voltage waveforms, (bottom) overview of the complete test bench including supply, compensation capacitors, load resistors, and measurement equipment.

is generally below critical levels; localized field enhancement near sharp conductor edges motivated the use of epoxy encapsulation for additional margin.

- Measurements on the fabricated prototype validated key electrical characteristics. The Bode analysis showed agreement with FEM within approximately 10% for most parameters, and the self-resonant frequency around 800 kHz confirmed negligible capacitive effects at the 1 kHz operating point.

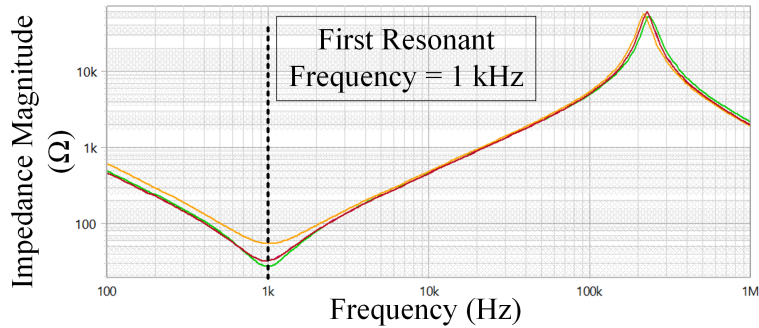


Figure 7.36: Measured frequency response of the transformer under short-circuit conditions using the Bode 100 Vector Network Analyzer (VNA).

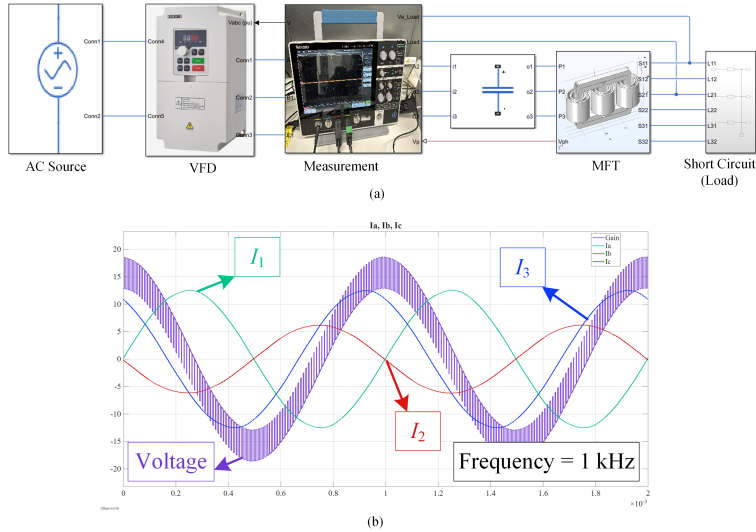


Figure 7.37: a) System-level simulation in Simulink, b) Simulated phase currents I_a , I_b , and I_c and primary-side voltage waveform at 1 kHz obtained from the system-level Simulink model with series-compensation capacitors.

- Insulation integrity was verified through PD measurements (below 10 pC as required by IEC 60076-11) on a representative single-phase setup. In addition, a destructive breakdown test on a sacrificial scaled sample provided an estimate of the dielectric strength margin relative to the IEC withstand requirement.

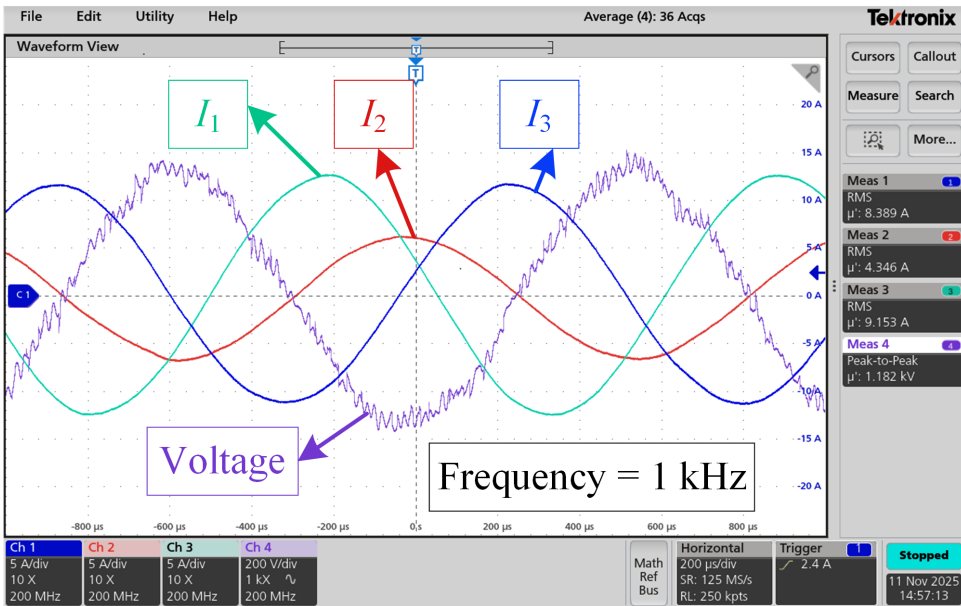


Figure 7.38: Experimental waveforms of the three-phase currents I_1 – I_3 and primary voltage of the prototype MFT during the 1 kHz short-circuit test.

- A single-coil thermal rise test showed a steady-state temperature rise of approximately 60°C, remaining within the limit for Class B insulation according to IEC 60076-11.
- Finally, three-phase short-circuit excitation with series compensation demonstrated stable operation at 1 kHz. The observed small current imbalance is consistent with the measured/simulated leakage inductance differences between the middle and outer phases in a three-limb core-type structure.

Overall, the prototyping results and experimental observations confirm that the developed MV-MFT is suitable for integration into the SST demonstrator and provides a validated baseline for the subsequent system-level tests and performance evaluation presented in the next chapter.

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8

CONCLUSION

As you start to walk on the way, the way appears.

Rumi

This section summarizes the principal outcomes of the research, focusing on transformer design methodology, modeling accuracy, and experimental validation under medium-frequency, non-sinusoidal voltage stress relevant to SST-based green hydrogen applications. The objective was to develop practical design methods and validation procedures for transformers operating under non-sinusoidal, medium-frequency voltage stress and high power density constraints.

8.1. KEY FINDINGS

This thesis investigated the design, modeling, and experimental validation of medium-voltage, medium-frequency transformers (MV-MFTs) for solid-state transformer (SST) applications in large-scale green hydrogen production. The main findings are as follows.

- Solid-state transformer architectures are suitable alternatives to conventional line-frequency transformer systems for supplying large electrolyzers, offering reduced size and weight together with improved controllability and modularity.
- Among the evaluated SST topologies, modular multilevel converter (MMC)-based structures provide favorable trade-offs between efficiency, scalability, harmonic performance, and system complexity for medium-voltage hydrogen production systems. Especially, modular multilevel resonant converters (MMR) shows promising results.

- Practical electromagnetic design procedures were developed for medium-voltage MFTs, including core selection, winding configuration, and frequency optimization, showing that increasing operating frequency does not always lead to overall system benefits due to increased losses and cooling requirements.
- Conventional insulation coordination practices based on 50/60 Hz standards cannot be directly applied to SST-based transformers because of non-sinusoidal voltage waveforms and different fault scenarios.
- The application of semiconductive coatings and screening techniques was shown to effectively reduce peak electric field stress and mitigate partial discharge activity in dry-type MFT designs.
- 3D-printed insulating components can be used in MFT structures when appropriate materials, geometries, and surface treatments are applied, enabling fast prototyping and compact insulation structures.
- A new current balancing technique was developed and experimentally validated, demonstrating improved current sharing and reduced thermal stress in parallel conductor paths.
- Lumped thermal network models, validated using FEM simulations and experimental measurements, provide accurate prediction of steady-state and transient temperature rise in compact MFT structures.
- A down-scaled prototype MFT was successfully designed, manufactured, and tested, and the experimental results confirmed the validity of the proposed electrical, insulation, and thermal design methodologies.

8.2. ANSWER TO THE RESEARCH QUESTIONS

The research questions formulated in Chapter 1 were addressed through system-level analysis, electromagnetic and thermal modeling, and experimental testing using a down-scaled prototype.

The first research question concerned the design requirements of high-power MV-MFTs for green hydrogen applications. Chapters 2 and 3 analyzed different SST topologies and transformer configurations under realistic operating conditions. Key design parameters, including voltage and current ratings, insulation distances, magnetic loading, and cooling constraints, were derived from the requirements of modular multilevel converter (MMC) and resonant modular multilevel converter (MMR) based SST systems. The analysis showed that three-phase MFTs offer

clear advantages in terms of weight, volume, and current handling compared with single-phase solutions. Frequency-dependent trade-offs between size, losses, and cooling complexity were quantified, leading to recommended operating ranges for different SST topologies.

The second research question addressed the applicability of existing transformer standards and the definition of suitable test requirements. The results show that conventional standards such as IEC 60076 cannot be directly applied to MV-MFTs operating in SST systems, mainly due to differences in voltage waveforms and thermal loading conditions. However, selected parts of existing standards, including insulation coordination, temperature rise limits, and partial discharge measurement procedures, remain applicable when interpreted with appropriate margins. A practical validation approach combining IEC 60076, IEC 60270, and IEC 60076-11 test principles was applied to the demonstrator and shown to provide meaningful assessment of insulation and thermal performance under medium-frequency operation.

The third research question focused on the nature of high-frequency dielectric stress and its impact on insulation design. Chapter 4 demonstrated that insulation performance in dry-type MV-MFTs is mainly limited by partial discharge and surface flashover rather than bulk breakdown. Finite-element field simulations showed that local geometric features can dominate peak electric field stress. The use of semiconductive screening and stress-cone termination was shown to effectively control electric field distribution and suppress partial discharge activity. Experimental measurements confirmed a significant reduction in partial discharge magnitude when screening was applied. The study also showed that 3D-printed insulation parts are highly sensitive to printing parameters and internal defects, and therefore require careful process control if used in high-voltage environments.

The fourth research question addressed the design of high-current, medium-frequency busbars and parallel windings. Chapter 5 analyzed parasitic parameters and their impact on current sharing in parallel disk windings. The results showed that small inductance mismatches can lead to significant current unbalance, especially at elevated frequencies. A compensation method based on series impedance tuning was developed and experimentally validated. The measurements confirmed that proper compensation can maintain balanced current sharing across a wide frequency range. These findings provide practical guidelines for conductor arrangement and impedance control in high-current MFT designs.

The fifth research question concerned thermal management and suitable cooling methods for MV-MFTs with high power density. Chapter 6 combined analytical loss models, frequency-domain core-loss estimation,

lumped thermal networks, and FEM simulations to evaluate temperature rise under different operating conditions. The results showed that oil-forced and water-forced cooling (OWWF) is required for high-power SST applications to maintain acceptable winding and oil temperatures. Thermal models were validated using prototype measurements, demonstrating good agreement between simulation and experiment. The analysis also highlighted the importance of including auxiliary cooling system size and weight when comparing SST topologies at system level.

Finally, Chapter 7 demonstrated the complete design-to-prototype workflow for a down-scaled three-phase 30 kVA, 1 kHz MV-MFT used in the SST demonstrator. The prototype validated key electromagnetic, insulation, and thermal design assumptions. Frequency response measurements, partial discharge tests, breakdown experiments on scaled samples, and thermal rise tests confirmed that the transformer design meets the selected performance and safety criteria. The experimental results also confirmed that the developed modeling tools provide sufficiently accurate predictions for practical design purposes.

Overall, this thesis provides a structured and experimentally validated design framework for medium-voltage, medium-frequency transformers in SST-based green hydrogen systems. The proposed methods address insulation coordination, parasitic effects, thermal management, and system integration in a unified manner. The presented guidelines can support future industrial development of SST-based power conversion systems for electrolyzers and other medium-voltage DC applications, and contribute to reducing the size and improving the efficiency of power interfaces in large-scale renewable energy systems.

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8.3. RECOMMENDATIONS

The following recommendations are proposed for future research and development of medium-voltage, medium-frequency transformers in solid-state transformer systems:

- Thermal performance is a critical limiting factor in high-power-density MFT designs. More detailed thermal modeling, including coupled electro-thermal simulations and advanced cooling system optimization, is recommended to better capture local hot spots and transient thermal behavior.
- This thesis mainly focused on steady-state operation of the MFT. Transient modeling is recommended to study the interaction between the transformer, power electronic converters, other SST modules, and electrolyzers during short-circuit events, start-up, shut-down, and other fast dynamic phenomena.

- Partial discharge behavior under medium-frequency and high-frequency excitation can differ from that under 50/60 Hz stress. Dedicated partial discharge measurement setups operating at frequencies such as 1 kHz and above are recommended to obtain more representative insulation performance data for MFT applications.
- Existing transformer standards are not fully applicable to SST-based systems and medium-frequency transformers. The development of dedicated testing and qualification procedures, coordinated through national and international standardization committees, is recommended to address insulation, thermal, and dielectric requirements specific to SST and MFT technologies.

Future work shall focus on developing insulation design and testing methodologies tailored to medium-frequency, converter-driven transformers. In particular, experimental investigation of partial discharge behavior under representative non-sinusoidal voltage waveforms is required to assess the validity of conventional 50/60 Hz testing approaches. Additionally, further research is needed on frequency-dependent dielectric properties and their impact on electric field distribution, especially in field grading structures such as stress cones and semiconductive coatings. Finally, a systematic comparison of waveform strategies (e.g., multilevel versus quasi-square operation) from the perspective of insulation stress and reliability would provide valuable insights for the optimal design of SST systems.

APPENDIX

SIMULATION METHOD

The simulations presented in this thesis are based on the finite element method (FEM). Due to limitations in time and computational resources, the FEM simulations are applied selectively to balance accuracy and computational cost. The simulation models are developed in COMSOL Multiphysics using 2D, 2D-axisymmetric, and 3D domains, depending on the complexity of the physical phenomena under investigation.

AC RESISTANCE

The AC resistance of a transformer conductor differs from its DC resistance due to skin and proximity effects. To evaluate the AC resistance in the MFT, the following simulation approach is employed. The magnetic field distribution in the core is evaluated using a two-dimensional axisymmetric finite element model implemented in COMSOL Multiphysics®. The simulation is performed at an operating frequency of 1 kHz and considers a single transformer phase to reduce computational cost. The simulation geometry and mesh are shown in Fig. 1.

The highest field stress is expected at the location with the minimum distance between the core leg and the surrounding structure, corresponding to the 45° angular position. Due to the axisymmetric formulation, the rectangular core cross-section is approximated by an equivalent cylindrical geometry.

The equivalent radius of the cylindrical core is calculated using the core flux utilization factor k_f as

$$R_{eq} = W_c \times \sqrt{2} \times k_f. \quad (1)$$

To maintain an equivalent magnetic path length, the core leg height is adjusted according to

$$H_c = 1.5 \times (H_w + W_c) + W_w + W_c, \quad (2)$$

where H_w and W_w are the window height and width, respectively, and W_c is the core width.

A periodic boundary condition is applied to the surrounding air domain to represent an infinitely repeating structure. The material properties

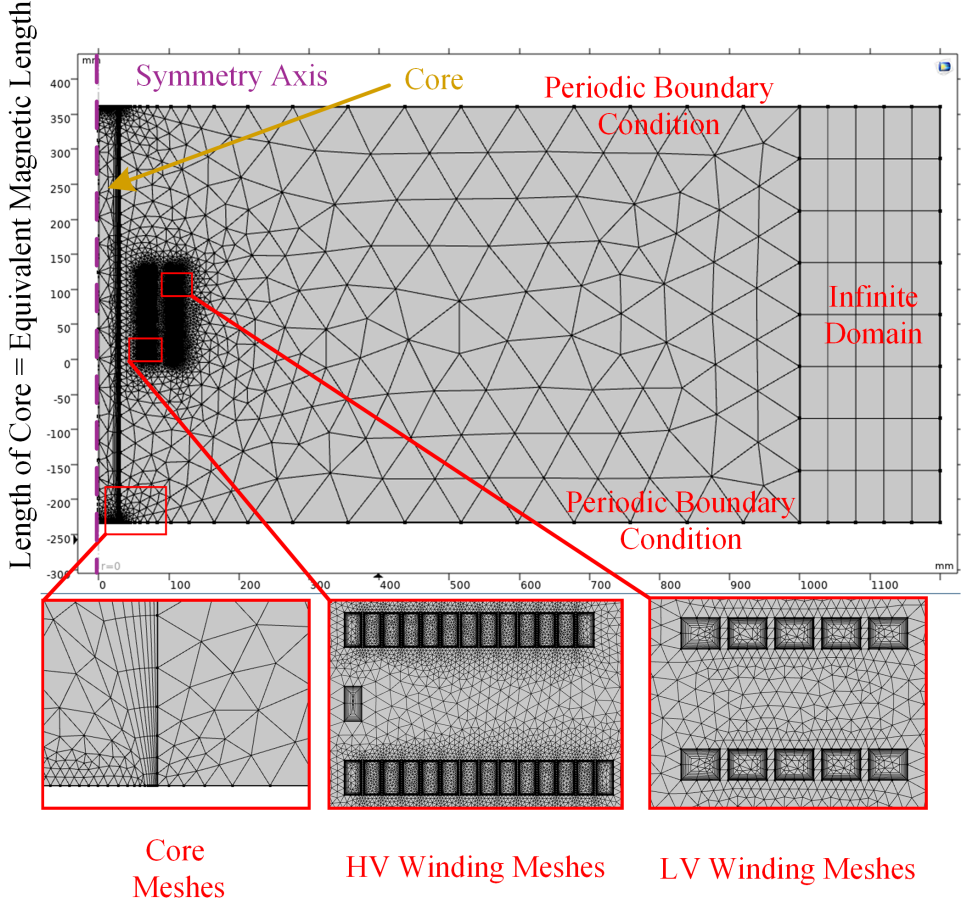


Figure 1: COMSOL geometry and mesh for magnetic fields simulation.

used in the simulation are listed in Table 1. The core is modeled as a solid block with an equivalent electrical conductivity to represent lamination effects. The effective conductivity is calculated as

$$\sigma_c = \frac{44 \times 10^6}{(R_{eq}/w_l)^2}, \quad (3)$$

where R_{eq} can be calculated using Equation 1 and w_l is defined by datasheet of the core material.

Table 1: Material Properties Used in Magnetic Field Simulation

Material	μ_r	σ [S/m]	ϵ_r
Air	1	1×10^{-5}	1
Copper	1	4.2×10^7	∞
Core	4×10^3	29.6	∞

INDUCTANCES

The leakage and magnetizing inductances are calculated using a three-dimensional finite element model. The windings are approximated as a single equivalent cylindrical conductor, which introduces an acceptable level of error. This simplification reduces computational effort and enables efficient solution of the 3D model. A 3D domain is required because, in a core-type three-phase transformer, the leakage and magnetizing inductances of the side phases differ from those of the middle phase. Capturing these differences is necessary for accurate resonant converter design.

To evaluate the leakage inductance, each primary winding is excited individually using a voltage source, while the remaining windings are short-circuited by applying zero voltage. To evaluate the magnetizing inductance, each primary winding is excited individually using a voltage source, while the remaining windings are open-circuited by imposing zero current. In both cases, the calculated inductance corresponds to the total inductance per phase, including both leakage and magnetizing components.

The magnetizing inductance is obtained by subtracting the leakage inductance from the open-circuit inductance as

$$L_m = L_{OC} - L_{LK} \quad (4)$$

where L_{OC} is the inductance calculated under open-circuit conditions and L_{LK} is the leakage inductance.

ELECTRIC FIELD AND CAPACITANCES

The electric field distribution and parasitic capacitances are derived from two-dimensional simulations to avoid unnecessary model complexity and computational cost.

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Every beginning has an end, and hereby I conclude my PhD. Yet, my journey will never truly end, as Saadi beautifully wrote:

"Far better roam the desert road than sit in futile rest; Though I may never reach my aim, I strive with all I possess."

به راه بادیه رفتن به از نشستن باطل که گر مراد نیابم به قدر وسع بکوشم

Reza Mirzadarani
Delft, December 2025

CURRICULUM VITÆ

Reza Mirzadarani

1991 Born in Isfahan, Iran.

EDUCATION

2018 MSc in Power Systems
University of Kashan, Iran

AWARDS

2025 Apple Prize' for outstanding research work in the Department of Electrical Sustainable Energy, TU Delft

2024 International Challenge in Design Methods for Power Electronics, Excellent Innovation Award, 2nd Place

2023 IES Students and Young Professionals Paper Assistance

LIST OF PUBLICATIONS

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13. Symmetrical Transformer for Medium-Voltage Medium-Frequency ISOP Three-Phase LLC SST, IEEE International Power Electronics and Motion Control Conference (PEMC), 2024
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16. Medium Frequency Transformer Optimization and Insulation Testing for Solid State Transformer Applications, IECON (IEEE Industrial Electronics Society Annual Conference), 2025
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18. Flyback-based Input Series Converter as a Wide Input Auxiliary Power Supply, PCIM Conference, 2025
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