

ISTANBUL TECHNICAL UNIVERSITY ★ GRADUATE SCHOOL

**A HOLISTIC DESIGN OPTIMIZATION METHOD
FOR LLC CONVERTERS IN LIGHT ELECTRIC
VEHICLE CHARGERS**

Ph.D. THESIS

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Department of Electrical Engineering

Electrical Engineering Programme

JULY 2024

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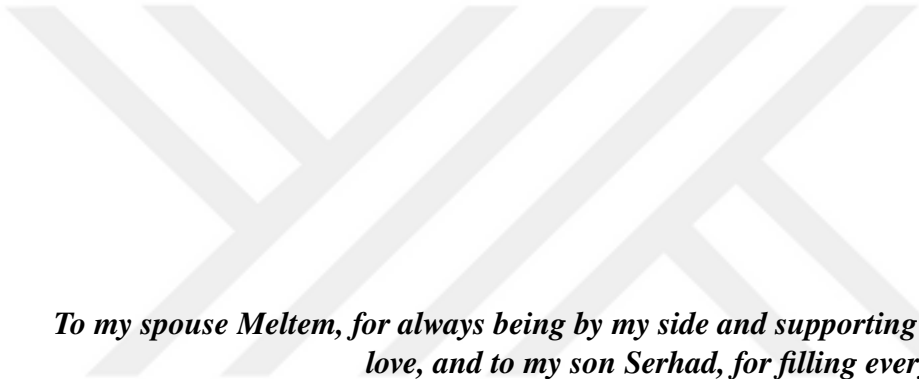
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To my spouse Meltem, for always being by my side and supporting me with endless love, and to my son Serhad, for filling every day with hope,



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ABBREVIATIONS

1D	: One-Dimensional
2D	: Two Dimensional
AC	: Alternating Current
ACO	: Ant Colony Optimization
ACL	: Asymmetrically-Wound Controllable Leakage
CC	: Constant Current
CM	: Common-Mode
CV	: Constant Voltage
DAB	: Dual-Active Bridge
DC	: Direct Current
DAP	: Die-Attached Pad
EMI	: Electromagnetic Interference
EV	: Electric Vehicle
ETD	: Economic Transformer Design
FHA	: First Harmonic Approximation
FEM	: Finite Element Method
FEA	: Finite Element Analysis
GA	: Genetic Algorithm
GHG	: Greenhouse Gas
KCL	: Kirchhoff's Current Law
LDV	: Light-Duty Vehicle
LEV	: Light Electric Vehicle
MF	: Mid-Frequency
MMF	: Magneto-Motive Force
NSGA-II	: Non-dominated Sorting Genetic Algorithm II
PC	: Pot Core
PCB	: Printed Circuit Board
PEIs	: Power Electronic Interfaces
PEV	: Pure Electric Vehicle
PFC	: Power Factor Correction
PHEV	: Plug-in Hybrid Electric Vehicle
PQ	: Power Quality
RM	: Rectangular Modular
RMS	: Root Mean Square
SFLA	: Shuffled Frog Leaping Algorithm
SR	: Synchronous Rectifier
TC	: Toroidal Cores

US	: United States
V2G	: Vehicle-to-Grid
ZCS	: Zero-Current Switching
ZVS	: Zero-Voltage Switching



SYMBOLS

L_r	: Resonance Inductance
L_m	: Magnetizing Inductance
C_r	: Resonance Capacitor
v_{tin}	: Resonant Tank Input Voltage
n	: transformer turn ratio
R_L	: Load
f_0	: Primary Resonant Frequency
f_m	: Secondary Resonant Frequency
V_{DC}	: DC Input Voltage
i_{Lr}	: Resonant Tank Current
φ_{tin}	: Primary Side Phase Shift
i_{in}	: Input Current
i_{sec}	: Tank Output Current
v_{tout}	: Tank Output Voltage
C_f	: Filter Capacitor
V_o	: Output Voltage
I_{out}	: Steady-State Load Current
R_e	: Resistive Load
H_s	: Resonant Tank Transfer Function
L_n	: Inductance Ratio
Q	: Quality Factor
F	: Frequency Ratio
f_s	: Switching Frequency
M	: Voltage Gain
Z_{in}	: Input Impedance
i_{co}	: Rectified Output Current
i_{Lm}	: Magnetizing Current
I_{Rm}	: Primary-Side Coupled Current of the Transformer
i_{D1}	: Diode Current
i_{Rm}	: Secondary Load Current
N_p	: Number of Turns in the Primary Winding
N_{ind}	: Number of Turns in the Inductor Winding
$I_{Lm \max}$	: Peak Value of Magnetizing Current
μ_{eff}	: Effective Relative Permeability
L_c	: Magnetic Path Length of the Core
P_{total}	: Total Loss
P_{cu}	: DC Copper Losses
R_{dc}	: DC Resistance

μ_0	: Magnetic Permeability of Free Space
e	: Natural Logarithm
r	: Radius
p	: Layers of Winding
Δ_{opt}	: Optimum Layer Thickness
A_{ws}	: Secondary Winding Area
ψ	: AC Resistance Coefficient
d_{opt}	: Optimum Foil Thickness
L_{opt}	: Optimum Foil Length
W_c	: Ratio of the Winding Conduction Area
W_a	: Window Winding Area of the Core
k_u	: Window Utilization Factor
ρ_w	: Electrical Resistivity of the Winding
N_p	: Number of Turns of the Primary Winding
N_{s1}	: Number of Turns of the Secondary Winding
P_{cuDC}	: DC Copper Loss
A_c	: Cross-Sectional Area of the Core
L_{lkg}	: Primary-Side Referred Leakage Inductance
a	: Winding Length
b	: Winding Build
c	: Insulation Thickness
V_c	: Core Volume
K_c	: Coefficient Loss Curves
α	: Coefficient Loss Curves
β	: Coefficient Loss Curves
ΔT	: Function of the Temperature Rise
R_θ	: Thermal Resistance for the Core Assembly
R_g	: Internal Gate Resistance
R_{gext}	: External Resistance
C_{iss}	: Effective Input Capacitance
V_{plt}	: Gate Plateau Voltage
V_{th}	: Threshold Voltage
V_{sd}	: Body Diode Voltage Drop,
I_{sd}	: Source to Drain Current
t_{bcd}	: Total Body-diode Conduction Time
L_{I1}	: Magnitudes of the Primary Leakage Inductance
L_{I2}	: Magnitudes of the Secondary Leakage Inductance
L_{ad}	: Sum of the Primary Leakage Inductance and Magnetizing Inductance
L_{kd}	: Inductance Value Measured with an RLC Meter
L_{C1}	: Parasitic Capacitance Measurements for the Primary Side
L_{C2}	: Parasitic Capacitance Measurements for the Secondary Side
C_{oss}	: MOSFET of Output Capacitance
w_{OP}	: Weight of Each Operation
h_c	: Heat Transfer Coefficient
γ	: Fringing Factor

A_L	: Inductance Per Turn
B_{max}	: Maximum Magnetic Field Density
B_{sat}	: Saturation Magnetic Field Density
A_p	: Product of the Core Window Area
k_a, k_w	: Dimensionless Coefficients
A_w	: Cross-Sectional Areas of the Primary and Secondary Windings
r_{pri}, r_{sec}	: Conductor Diameter
g	: Gap Value
L_c	: Core Winding Length
A_L	: Air-Gap Value
V_c	: Volume of Core
C_m	: Coefficients for Core Loss Calculation
$\sigma_{core, x}$	: Specific Costs Per Weight
$\sigma_{wdg, x}$	: Specific Costs Per Weight
$\sigma_{lab, x}$	: Specific Costs Per Weight
C_{eq}	: Switch Node Capacitance
$\sigma_{sink, x}$	: Specific Costs Per Volume
L_k	: Leakage Inductance
T_c	: Transformer Temperature
T_{amb}	: Ambient Temperature
$P_{pri\ loss}$	: Primary Winding Copper Loss
$P_{sec\ loss}$	: Secondary Winding Copper Loss
$R_{t\ pri}$	: Thermal Resistances of the Primary Winding
$R_{t\ sec}$	: Thermal Resistances of the Secondary Winding
$R_{t\ core}$	: Thermal Resistances of the Core
L_n	: Minimum and Maximum Frequencies are Internally Swept
V_{DS}	: Drain-Source Voltage of Primary Device
R_{dson}	: MOSFET Values on the Secondary Side
ω_m	: Modulation Frequency
ω_s	: Switching Frequency
L_e	: Equivalent Inductor
X_{eq}	: Impedance
L_e, C_e	: Beat Frequency Dynamics
T_0	: Resonant Period
L_{eq}	: Equivalent Resonant Inductor
C	: Coefficients for Various Combinations of Primary and Secondary Connections
h	: Thickness of the Conductor
ε	: Effective Thickness of the Skin Depth
δ	: Skin Depth at the Given Frequency
m	: Impact of a Factor on AC Resistance
P	: Primary Conductor
J	: Current Density Distribution
A_e	: Effective Area
$R_{Sec\ dc}$	: Secondary DC Winding Resistance Winding
$R_{I\ dc}$	: Primary DC Winding Resistance for Inner Turn
P_{core}	: Core Loss
R_{ac}	: AC Resistance



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A HOLISTIC DESIGN OPTIMIZATION METHOD FOR LLC CONVERTERS IN LIGHT ELECTRIC VEHICLE CHARGERS

SUMMARY

Since the early 1990s, gasoline prices have been steadily increasing, making traditional transportation more expensive. Meanwhile, advancements in material science and battery manufacturing technology have led to gradually decreasing battery costs and increasing energy storage capacities each year. These changes are encouraging the shift from traditional gasoline engines to modern electric drive systems, leading to a growing focus on developing Electric Vehicle (EV) technologies.

EVs are gaining attention due to their energy-saving capabilities, low carbon emissions, and long-term cost-effectiveness. The power source for EVs is a high-voltage DC rechargeable battery that can be fully charged by connecting to an external electrical source. EVs require an onboard charging system that converts alternating current (AC) from the grid to direct current (DC) to charge the vehicle's battery. This charging system can be connected to a household outlet or a fast-charging station. Therefore, a highly efficient and compact battery charging system plays a crucial role in the electrical system of EVs.

A novel optimization method for the design of gapped and distributed core magnetics in LLC converters, suitable for applications ranging from battery chargers to light electric vehicle charging systems, is presented in this thesis. Through a sequence of integrated studies, a comprehensive framework for the design and optimization that addresses the challenges posed by varying loads and operational conditions is developed.

Initially, an innovative optimization technique utilizing by sweeping the magnetic flux density is introduced. This technique enables the optimal magnetic flux density to be selected, minimizing the power loss, cost, and volume of magnetic components. Unlike traditional design algorithms that focus on singular core configurations, this method considers multiple distributed cores. The validity of this approach is demonstrated through co-simulation and experimental results on a high-power prototype, offering a significant advancement over conventional design methodologies.

Further, the exploration extends to a holistic design optimization method. This method systematically considers the non-linear load profiles characteristic of battery charging applications and incorporates a multi-core configuration for transformer size reduction. The effectiveness of this approach is evidenced by the implementation in a light electric vehicle charger, achieving a peak efficiency of 98.2%.

Additionally, a high-precision simulation model incorporating parasitic elements is developed to accurately represent LLC resonant converters. This model enhances the

understanding of system behaviors under various operational conditions, highlighting the importance of considering transient phenomena.

An extended describing function model is also examined, facilitating the control of LLC converters with variable switching frequencies. This model aids in the compensator design for constant-current and constant-voltage operations, ensuring stable performance across diverse load conditions.

A specific focus is placed on the design and optimization of a three-phase LLC converter with a $\Delta - \Delta$ winding configuration. This study proposes a systematic design framework that optimizes the resonant frequency to minimize power loss, volume, and cost, showcasing a substantial improvement in magnetic losses compared to traditional configurations.

In this thesis, these contributions offer a significant advancement in the field of LLC converter optimization, presenting a robust foundation for future research and development in power conversion systems.

In conclusion, this thesis presents a series of interconnected studies that collectively enhance the understanding and implementation of LLC converter technology for LEV charging and beyond. While each research component is valuable on its own, together they contribute to a broader narrative of innovation and optimization in power electronics design. The findings not only offer practical solutions to current design challenges but also pave the way for future advancements in this field.

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HAFİF ELEKTRİKLİ ARAÇ ŞARJ CİHAZLARINDAKİ LLC DÖNÜŞTÜRÜCÜLER İÇİN BÜTÜNSSEL BİR TASARIM OPTİMİZASYON YÖNTEMİ

ÖZET

1990'ların başından bu yana benzin fiyatı sürekli olarak artmaktadır. Bu artış geleneksel ulaşımı daha pahalı hale getirmektedir. Bu arada, malzeme bilimi ve batarya üretim teknolojisi ilerledikçe, batarya maliyetleri yavaş yavaş düşmekte ve enerji depolama kapasiteleri her yıl artmaktadır. Bu değişiklikler, geleneksel benzinli motorlardan modern elektrikli sürüş sistemlerine geçişi teşvik etmektedir. Bu nedenle Elektrikli Araç (EV) teknolojilerinin geliştirilmesine giderek daha fazla odaklanılmaktadır.

EV'ler enerji tasarrufu, düşük karbon emisyonu ve uzun vadede uygun maliyet avantajları nedeniyle giderek daha fazla dikkat çekmektedir. EV'lerin güç kaynağı, harici bir elektrik kaynağına takılarak tam şarj edilebilen yüksek gerilimli DC şarj edilebilir bir bataryadır. EV'ler, aracın bataryasını şarj etmek üzere şebekeden gelen alternatif akımı (AA) doğru akıma (DA) dönüştürmek için yerleşik bir şarj sistemine ihtiyaç duyar. Bu şarj sistemi bir ev prizine veya bir hızlı şarj istasyonuna bağlanabilir. Bu nedenle, yüksek verimli ve küçük hacimli bir batarya şarj sistemi, EV'lerin elektrik sisteminde çok önemli bir rol oynamaktadır.

Bu doktora tezi, hafif elektrikli araç (LEV) şarj sistemlerindeki uygulamalara odaklanarak, LLC dönüştürücülerinin tasarım en iyileştirmesi üzerine çok sayıda araştırma çalışmalarını tutarlı bir anlatı halinde birleştirmektedir. Tez, manyetik bileşen tasarımı, sistem en iyileştirmesi ve parazitik unsurların yüksek doğruluklu benzetim modellerine dahil edilmesine yönelik kapsamlı bir yaklaşımla sonuçlanan her bir çalışmanın bulgularını sırayla inşa edecek şekilde yapılandırılmıştır.

İlk çalışma, transformatör ve indüktör bileşenlerinin iyileştirmesine vurgu yaparak LLC rezonans dönüştürücülerinin manyetik tasarımına yeni bir yaklaşım getirmektedir. Bu yaklaşım, batarya şarj cihazları ve hafif elektrikli araç şarj sistemleri gibi çeşitli uygulamalar için uygun bir çözüm sunar. Geliştirilen en iyileştirme tekniği, manyetik akı yoğunluğunun belirli aralıklarda değiştirilmesi esasına dayanır. En iyi manyetik akı yoğunluğunun seçilmesiyle güç kaybını, maliyeti ve hacmi en aza indirgeyerek, geleneksel tasarım algoritmalarına kıyasla önemli bir iyileşme sağlar. Isıl ve manyetik sınırlamalar altında bakır ve çekirdek kayıplarını maliyet, hacim ve güç verimliliği kısıtlamalarına karşı sistematik olarak dengeler. Algoritmanın etkinliği, LEV'ler için 3700W 48V LLC şarj sisteminin tasarımıyla gösterilmiş ve geleneksel tek çekirdekli tasarımlara kıyasla dağıtılmış çekirdek yapılarının ve değişen sargı türlerinin faydaları vurgulanmıştır. Bu optimize edilmiş yaklaşım sadece kayıpları azaltmakla kalmaz, aynı zamanda ısıl yönetim zorluklarını da ele alır, böylece genel verimliliği ve performansı artırır.

Manyetik en iyi duruma getirmeyi genişleten sonraki çalışma, LLC dönüştürücüler için sistem düzeyinde tasarım hususlarını içerecek şekilde kapsamı genişletmektedir. Rezonans frekansı belirleme, anahtarlama frekansı limitleri ve kalite faktörü kriterlerini tasarım sürecine dahil eden bu araştırma, kapsamlı bir uygunluk fonksiyonuna dayalı olarak tüm dönüştürücüyü iyileştirmektedir. Sonuç, yalnızca LEV şarj cihazlarının yüksek verimlilik ve küçük boyut gereksinimlerini karşılamakla kalmayan, aynı zamanda deneysel ve benzetim sonuçları aracılığıyla önerilen tasarım çerçevesini doğrulayan 3700W / 48V LLC prototipidir.

Bu tez, batarya şarj sistemlerinin doğrusal olmayan yük profillerini ve çok çekirdekli konfigürasyonları kullanarak trafonun boyutunun küçültülmesini sistematik olarak dikkate alır. Uygulamada, bu yaklaşım bir hafif elektrikli araç şarj cihazında test edilmiş ve %98,2 gibi etkileyici bir zirve verimliliğe ulaşmıştır.

Doğru modellemenin öneminin farkında olarak, tezin bir başka yönü de parazitik unsurları LLC dönüştürücü benzetimlerine dahil etme zorluğunu ele almaktadır. Bu araştırma, parazitik bileşenlerin gerçek dünyadaki etkilerini hesaba katarak dönüştürücü performansının tahmin doğruluğunu artırmakta ve böylece yüksek verimli, yüksek güç yoğunluklu dönüştürücülerin daha güvenilir tasarımını ve en iyi duruma getirilmesini sağlamaktadır.

Tez, alternatif sargı bağlantılarına sahip (üçgen-üçgen, yıldız-yıldız, zigzag-zigzag vs.) üç fazlı bir LLC dönüştürücünün tasarımına ve optimizasyonuna odaklanır. Bu çalışma, rezonans frekansını optimize ederek güç kaybını, hacmi ve maliyeti en aza indirmeyi amaçlayan sistematik bir tasarım çerçevesi önerir.

Bu tez kapsamında batarya şarj uygulamaları için tasarlanmış, çıkış akımı ve çıkış geriliminin her ikisinin de dikkate alındığı bir LLC rezonans dönüştürücü için bir kapalı döngü kontrolü önermektedir. LLC rezonans dönüştürücü için bir sabit akım (CC) ve sabit gerilim (CV) kontrol sistemi batarya şarj uygulamaları için ayrıntılı olarak sunulmuştur. LLC dönüştürücünün küçük sinyal modeli, daha doğru dinamik model elde etmek için hem rezonans üstü hem de rezonans altı koşullar için türetilmiştir. LLC rezonans dönüştürücünün basit bir üçüncü dereceden eşdeğer devre modeli türetilmiş ve dönüştürücünün darbe frekansı dinamikleri ve küçük sinyal davranışı bu devre modeli ile çok iyi bir şekilde tahmin edilmiştir. Her bir transfer fonksiyonuna oransal-integral kontrol uygulanmıştır. Sırasıyla 48V ve 54V çıkış geriliminde CC ve CV işlemleri için tasarlanan kompensatörler farklı çalışma noktalarında simüle edilmiştir. Çıkış akım ve gerilim referanslarının tüm yük profilleri için başarılı bir şekilde takip edildiği gösterilmiştir.

Yüksek anahtarlama hızı, bir anahtarlama döngüsü sırasında depolanacak daha az enerji anlamına gelir. Sonuç olarak, manyetik bileşenin boyutu büyük ölçüde azaltılabilir. Buna ek olarak, azaltılmış sarım sayısı, baskı devre kartını sargı olarak kullanma fırsatı sağlar. Geleneksel litz-tel tabanlı manyetiklerle karşılaştırıldığında, düzlemsel manyetik bileşenler dönüştürücü boyutunu etkili bir şekilde azaltır.

Sonuç olarak, bu tez LEV şarjı ve ötesi için LLC dönüştürücü teknolojisinin anlaşılmasını ve uygulanmasını bütüncül olarak geliştiren bir dizi birbirine bağlı çalışma sunmaktadır. Her bir araştırma bileşeni kendi başına değerli olmakla birlikte, güç elektroniği tasarımında daha geniş bir yenilik ve optimizasyon anlatısına katkıda bulunmaktadır. Bulgular sadece mevcut tasarım zorluklarına pratik çözümler sunmakla kalmamakta, aynı zamanda bu alanda gelecekteki ilerlemelerin de önünü açmaktadır.

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1. INTRODUCTION

Since the early 1990s, the price of gasoline has been rising steadily. This increase makes traditional transport more expensive. Meanwhile, as material science and battery production technology advance, battery costs are slowly falling, and their energy storage capacity improves every year. These changes are encouraging a shift from traditional gasoline engines to modern electric driving systems. As a result, there's a growing focus on creating better Electric Vehicle (EV) and Plug-in Hybrid Electric Vehicle (PHEV) technologies [1].

In the United States (US), transportation relies almost entirely on oil, posing a significant issue since it accounts for a large share of the world's and the US's Greenhouse Gas (GHG) emissions. Specifically, Light-Duty Vehicles (LDVs) contribute to 17.5% of the US's Carbon Dioxide (CO₂) emissions [2, 3]. A study by the Electric Power Research Institute and the Natural Resources Defense Council shows that traditional vehicles emit three times more GHGs than PHEVs when the electricity comes from renewable sources. Even with coal-generated electricity, emissions are cut by about 40% [4]. Hence, the shift towards EVs, including PHEVs and Pure Electric Vehicles (PEVs), is essential for energy conservation and environmental protection. Both PHEVs and PEVs require an onboard charger to convert AC from the grid to DC to recharge the vehicle's high-voltage battery. This charger can be connected to a home socket or a fast-charging station. Therefore, an efficient yet small battery charger plays a crucial role in the electric system of PEVs and PHEVs.

EV charging systems come in two main forms: conductive and inductive. Conductive charging involves a physical connection to transfer power from the AC grid to the EV. On the other hand, inductive charging relies on magnetic fields created by two coils to move energy from the AC grid to the vehicle's onboard battery without a wired connection. Conductive chargers are organized based on the amount of power they provide, details of which are outlined in Table 1.1 [5].

Table 1.1 : Different level of charging.

Charging Level	Setting	Supply Power
AC Level 1	Residential/parking lot 5mi/h @ 1.7kW	120V AC/16A
AC Level 2 (minimum)	Residential/Commercial 10mi/h @ 3.4kW	208/240V AC/16A
AC Level 2 (maximum)	Residential/Commercial 60mi/h @ 19.2kW	208/240V AC/100A
DC Level 1	Commercial 120mi/h @ 40kW	208V AC/480V AC (Three Phase)
DC Level 2	Commercial 300mi/h @ 100kW	208V AC/ 480V AC (Three Phase)

The Level 1 charger, typically found in homes, works with 120V AC outlets and has a maximum output of 1.9 kW. It can add about 5 miles of range for each hour of charging. Level 2 chargers use either 208V AC or 240V AC, common in U.S. households, and can provide up to 19.2 kW depending on the electrical circuit's capacity. At full power, it can deliver around 120 miles of range per hour of charging. In contrast to AC chargers, DC fast chargers provide power directly to the battery, bypassing the car's onboard charger due to their much higher power capacity. These are typically installed at commercial charging stations. A DC level 1 charger can offer 40 kW, while a DC level 2 charger can go up to 100 kW, enabling much faster charging times.

Over the past ten years, the e-mobility sector has seen significant progress, with both commercial and academic entities making strides. Specifically, there has been a concentrated effort to study and enhance the performance of onboard chargers for Light Electric Vehicles (LEVs).

Li-ion batteries have a much higher energy density compared to other types like lead-acid, nickel-cadmium, and nickel-metal hydride batteries [6]. For EVs and PHEVs, the battery's energy density and weight are crucial as they directly affect the vehicle's driving range. As a result, Li-ion batteries are the preferred choice in the EV and PHEV market. It's important to recognize that despite the advancements in battery technology that have led to longer life cycles, higher energy densities, and some cost reductions, the Li-ion battery pack remains the most expensive and heaviest part of EVs and PHEVs [7]. Furthermore, since the onboard charger's weight contributes to

the overall weight of the vehicle and its cooling system can add to the cost. There is a market drive for chargers that are not only smaller but also highly efficient [8].

Different approaches can be employed to charge Li-ion batteries safely. The power level at which a battery can be charged is influenced by its chemistry and the charging technique applied. The process for charging Li-ion batteries usually involves two main stages: Constant Current (CC) and Constant Voltage (CV). Within these stages, there are four critical operation points: the beginning, nominal, turning, and end points. Each point marks a specific phase in the charging cycle and greatly affects the charger's overall performance.

Figure 1.1 likely displays the typical charging profile of a Li-ion battery cell, highlighting the four essential operation points and showing the transition from the CC phase to the CV phase. This profile is fundamental to grasping how Li-ion batteries charge and is vital for designing and refining the charger to ensure optimal performance and battery health.

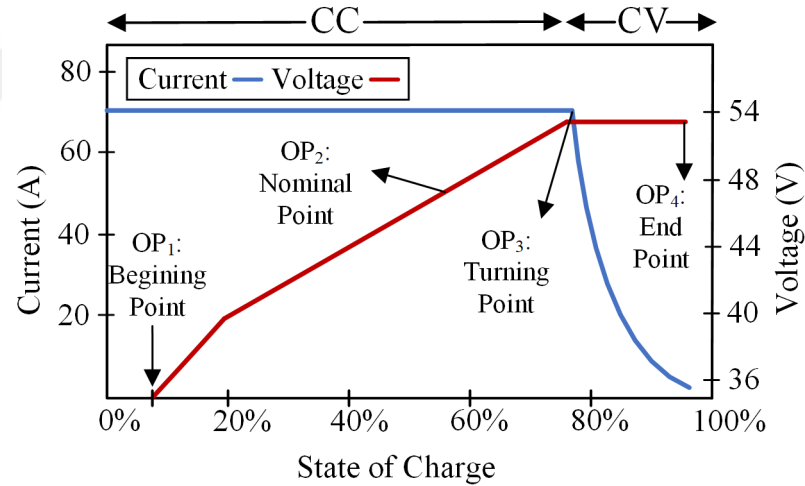


Figure 1.1 : Typical charging characteristics of a Li-ion battery cell and key operation points.

The standard power setup of an EV is depicted in Figure 1.2 [9]. This setup includes a high-voltage (300 V to 400 V) battery pack with a high energy capacity (tens of kWh) that serves as the onboard energy storage. Additionally, there are three main Power Electronic Interfaces (PEIs) responsible for different power conversion processes. These are: a) the PEI for electric propulsion, which powers the vehicle's movement; b) the PEI for onboard appliances, which supports the functioning of internal devices; and c) the PEI for onboard charging, which manages the battery's charging process [10].

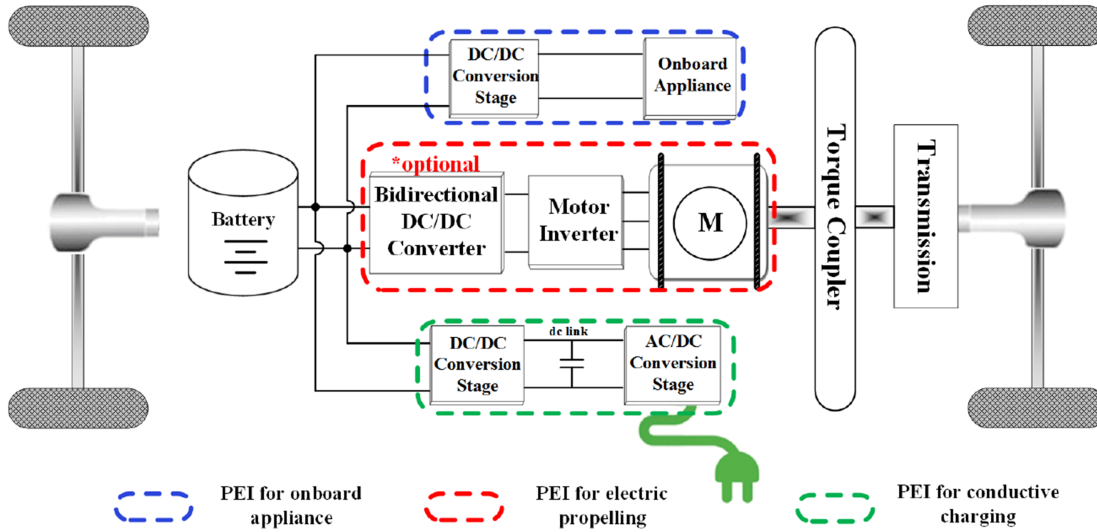


Figure 1.2 : General power architecture of an EV [9].

The PEI for electric propulsion in an EV mainly includes a motor inverter and a bidirectional DC/DC converter. This interface works in two modes: the propelling mode and the regenerative braking mode. During propelling mode, energy flows from the battery pack to power the electric motor, which propels the vehicle. In regenerative braking mode, the motor reverses its role and acts as a generator, converting the kinetic energy from braking back into electrical energy, which is then fed into the battery pack. The motor inverter functions as a rectifier in this mode. If present, the bidirectional DC/DC converter allows for better management of the battery's charge and discharge processes as part of the PEI for electric propulsion [11].

The PEI for onboard appliances in an EV typically consists of a DC/DC converter. This converter's role is to reduce the high voltage from the battery pack, which ranges from 300 to 400 volts, down to 12 volts. This lower voltage is used to power the vehicle's electrical appliances, including the air conditioning, headlights, stereo systems, among others. It is essential for the converter to have galvanic isolation to safeguard the low voltage electronics from the high voltage system, which could be dangerous [12].

In the PEI for onboard appliances, the energy flow is only in one direction. For the onboard charger PEI, its primary function is to manage the transfer of power from the electrical grid to recharge the battery pack. This onboard charger typically includes two main stages: the first stage is responsible for converting AC power from the grid to DC, along with improving the power factor, which enhances the efficiency of the

power transfer. The second stage involves converting DC power to the appropriate level needed by the battery, while also providing galvanic isolation to ensure safety by preventing any direct electrical connection between the high-voltage and low-voltage systems [13].

At present, onboard chargers in commercial vehicles are designed for power to flow in one direction only, from the grid to the vehicle. Given that vehicles are parked about 95% of the time on average, it's suggested that their batteries could serve a dual purpose by allowing electricity to flow both ways—from the vehicle back to the power grid and vice versa. This concept is central to the developing Vehicle-to-Grid (V2G) technology, which necessitates onboard chargers to support bidirectional power flow. This means that when the vehicle is not in use, its battery could supply power back to the grid, potentially contributing to energy distribution and storage solutions [14–17].

Traditional LEV chargers usually feature a two-stage power architecture, which includes stages for Power Factor Correction (PFC) and isolated DC/DC power conversion as shown in Figure 1.3(a) [18]. The first stage, the AC/DC PFC converter, is composed of an Electromagnetic Interference (EMI) filter, a rectifier, a PFC converter and a DC link capacitor. The PFC converter is operated with a high-frequency signal to ensure that the AC line current aligns with the AC line voltage and frequency. Ideally, this AC/DC PFC stage behaves similarly to a resistive load to eliminate total harmonic distortion and enhance power transfer efficiency as shown in Figure 1.3(b).

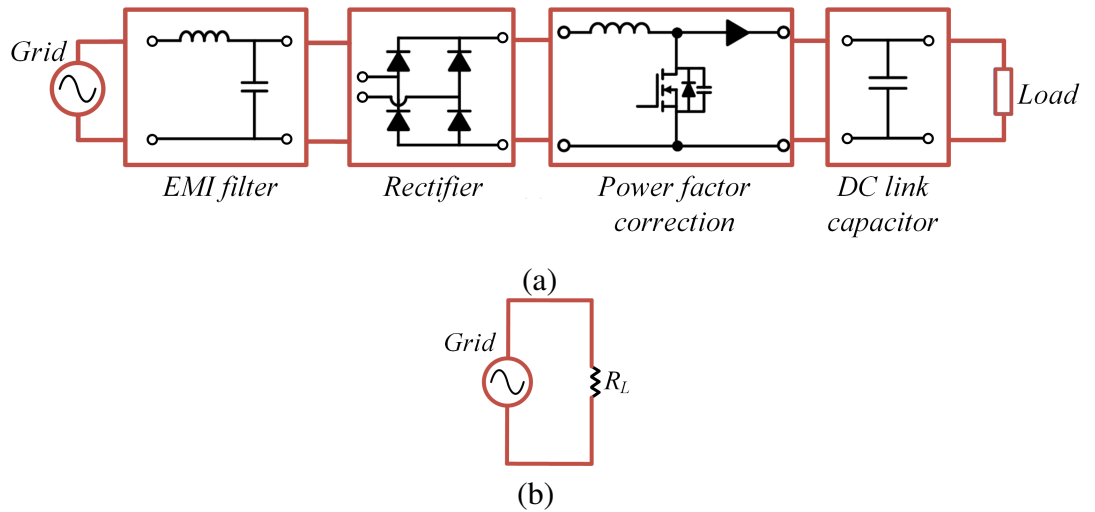


Figure 1.3 : AC/DC PFC stage (a) Typical block diagram, (b) equivalent circuit model.

The typical second stage involves an isolated DC/DC converter, which includes a switching network, a high-frequency transformer, a rectifier, and a low-pass filter as shown in Figure 1.4. For frequency modulated resonant converters, a resonant tank is also needed between the switching network and the high-frequency transformer to facilitate the conversion process. In the DC/DC conversion stage of LEV battery chargers, resonant topologies are often employed to convert high voltage into an isolated and controlled low voltage [19,20]. Resonant converters enable lower switching losses and higher switching frequencies through soft switching. This improvement increases the energy density of power electronic circuits used in various industrial fields such as power supplies for communication systems, battery charging systems, and power provisioning for data centers. Therefore, due to the growing need for power electronics in recent years, resonant converters have become increasingly prominent for reasons such as high frequency, high power density and high efficiency [21].

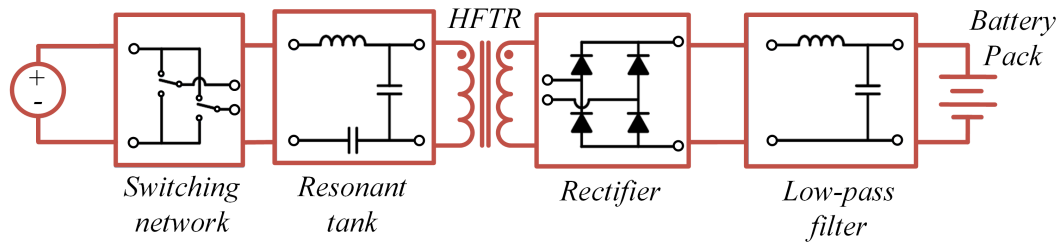


Figure 1.4 : Typical configuration of isolated DC/DC topology.

In resonant converters, series and parallel resonant converters are primarily distinguished. However, in series resonant converters, voltage control can not be maintained when operating in no-load conditions, resulting in a voltage lower than the source voltage. Therefore, they can only be used as step-down converters. On the other hand, the efficiency of parallel resonant converters is quite low under light load conditions, and they operate inefficiently in applications with variable load currents [22]. For these reasons, series and parallel converters are not preferred in EV applications or in applications requiring a wide output voltage range. Instead, LLC resonant converters, which combine the strong advantages of both series and parallel converters and offer high efficiency along with high power density, are increasingly favored in such applications [23]. For high-power applications, such as a 3700W / 48V LEV charger, the LLC resonant converter has become the standard DC/DC

isolating converter. Its ability to operate under Zero-Voltage Switching (ZVS) across the full range of frequencies and loads makes it particularly appealing. When designed correctly, LLC resonant converters can achieve high power density and efficiency, making them ideal for such demanding applications [24–28]. LLC converters are capable of achieving ZVS from zero to the rated load, maintaining a low turn-off current for the primary switches. Additionally, they can achieve Zero Current Switching (ZCS) for the Synchronous Rectifiers (SRs). These soft turn-on and turn-off characteristics help in significantly reducing EMI, making LLC converters efficient and suitable for applications where minimizing noise and disruption is critical [25]. Controlling an LLC resonant converter presents a challenging task due to its variable switching frequency operation, especially under varying load conditions like those encountered with batteries in EV. This variability in load demands precise and adaptable control strategies to maintain optimal performance and efficiency across the entire range of operation [29]. To achieve a regulated output in both PWM converters and resonant converters, a feedback loop is incorporated into the control system. This feedback loop is essential for stabilizing the system and regulating the output voltage or current, ensuring that the output remains consistent despite variations in input or load. Typically, small signal equivalent circuit models are employed in the design of a robust control system for all types of converter structures, allowing for precise control and stability across a wide range of operating conditions.

1.1 Purpose of Thesis

In this thesis context, a LLC resonant converter design for LEV applications is aimed to be achieved, which employs a single and 3-phase with different connection topologies, conventional magnetics and integrated planar transformer, where input-output voltages can vary over a wide range. The common approach to design an LLC converter is to determine the switching frequency range and resonance frequency, and then select the magnetics, power switches and heat-sinks, accordingly [30]. The challenge with this approach is that the switching frequency is chosen based on designer's experience or availability of components without conducting a systematic design framework [31]. The downside of this design procedure is clear; the resonances elements such as resonance inductance (L_r), magnetizing inductance (L_m), and resonance capacitor (C_r) values,

and corresponding core material and size for magnetics are not optimally selected. This approach hinders the power density as the total power losses and power density are determined based on the initially selected switching frequency range, and resonance frequency [32].

In this thesis, a systematic design framework for LLC converter is proposed to find the optimal resonance frequency, minimum and maximum switching frequency limits that satisfy the required voltage gains at corresponding quality factors by minimizing the system level fitness function. For each resonance frequency, satisfying L_r , L_m and C_r values are calculated. For each calculated combination, a transformer and an inductor are designed based on the analytical models for every possible core in the database. Next, the transformer and the inductor are selected among the designed candidates, which minimizes the defined local fitness function. A key strength of this approach lies in its capacity to apply diverse fitness functions to local and system-level designs. In this approach, multi-core transformer configuration with primary windings in series and secondary windings in parallel is also considered instead of a single-core transformer as in the traditional LLC resonance converter. On the power stage side, the primary and secondary power switch losses for the devices in the database and the required heatsink sizes considering thermal resistances of the packages are calculated. After obtaining the design space, a system level fitness function consisting of power loss, system volume and cost is applied to find the most optimal resonance frequency together with the corresponding locally optimized L_r - L_m - C_r combinations and all other non-trivial components. An integrated magnetic structure based on Printed Circuit Board (PCB) winding, requiring a simple EI core for the integration of both the transformer and the inductor, is proposed. The proposed structure allows to control the leakage inductance through the adjustment of the center leg air gap according to the requirements of different applications. Moreover, the leakage flux is confined within the core instead of flowing through the air, which could potentially benefit radiated EMI and avoid additional eddy current loss in nearby metals.

Furthermore, in this study, parasitic effects in an LLC resonant converter, designed for LEV charging system with 400V/48V input/output voltage and 3.7kW rated power, are added to the simulation model. The proposed simulation model includes the primary and secondary side power switch parasitic capacitances, the resistance and parasitic

capacitance of the resonant inductance, the primary, secondary and inter-winding parasitic capacitances of the transformer, and leakage inductance of the transformer. By adding the parasitic components to the simulation model, a more detailed model showing the oscillation waveforms during transients is presented in this paper. In addition, this thesis proposes a closed-loop control for an LLC resonant converter, designed for battery charging applications, in which the output current and output voltage are both considered. Extended describing function method considering the states of the resonance elements relative to the switching frequency for equivalent circuit approximation has been proposed for modeling LLC converter. Thus, a simple third order equivalent circuit model of the LLC resonant converter was obtained and the beat frequency dynamics and small signal behavior of the converter were very well predicted with this circuit model. A proportional-integral control has been applied to the each transfer function.

1.2 Literature Review

In comparison to soft-switching PWM-controlled converters, LLC resonance converters can work at a higher switching frequency with higher efficiency particularly at light load, resulting in a higher power density [19, 20, 33]. However, the selection and design of components remain a difficult task, particularly for varying loads such as batteries in LEVs. These vehicles are defined by the industry as EVs with power range from 1 kW to 10 kW and voltage classes between 24 V and 72 V [34]. Within these voltage classes, the 48 V category is commonly utilized.

For high power, i.e. 3700 W / 48 V LEV charger, LLC resonant converter has been the mainstream DC/DC isolating converter as it can operate under ZVS in the full frequency and load ranges offering high power density and efficiency when properly designed. So far in literature, mainly single-phase LLC resonant converters have been preferred in low output voltage-high current LEV charging applications [26–28, 35]. Different design techniques have been proposed to achieve high power density and a wide voltage gain range of the single-phase LLC resonant converter [36].

These studies cover changes in resonance tank structure, modifications on the impedance network and modulations of primary and secondary side switches [37–39]. Moreover, integrated planar structures [24], variable DC link voltage [8, 40], a

configurable dual LLC converter [41], and a two-level turn ratio transformer [42] are also proposed to boost the efficiency and increase the power density of single-phase LEV chargers. Alternative to the single-phase LLC converter, three-phase LLC resonant converters have been proposed in some studies in high power charging DC/DC conversion stage to further boost the efficiency, shrink the size of the magnetic cores, and provide better thermal performance [43]. One of the first studies on three-phase LLC converters has been proposed in [44], where the primary sides are connected in wye-configuration. Due to the poor transient performance of the wye-connection of primary side, the winding configuration in which the resonant tank capacitors are connected in Δ is suggested in [45].

Different three-phase interleaved LLC converter topologies and state-of-the-art magnetic integration are investigated in [46]. Efforts have also been made to improve the voltage gain over a wide range for three-phase LLC converters [47]. In none of these studies, the optimization of the three-phase LLC resonant converter has been investigated. Furthermore, the detailed expressions of the resonant current depending on the output current and magnetizing current have not been provided in any of the literature.

To increase the power density of an LLC converter, the switching frequency is pushed higher to shrink the size of magnetics. Even though appropriate ferrite cores exhibiting low core loss at high frequencies are preferred, there may still be considerable power losses on the core and the windings due to skin and proximity effects resulting in thermal management issues. In the literature, some researchers proposed topologies with distributed cores instead of using a single core [35, 48–53]. With the distributed core structure, the copper losses and generated heat on the transformer are distributed equally among multiple cores. Likewise, the thermal burden on the secondary-side rectifiers are reduced. The details of the distributed core structure connected in series-parallel are provided in detail in [48] and [49]. In [50] and [51], it was shown that using two small transformers instead of a large one increased the power density of the converter by 24.28% [52, 53]. Substantial research and development efforts have been performed on the optimization of high frequency transformers [54–60]. The high frequency performance of different core materials and loss models are analyzed in [54]. In [55], the authors recommended a formula for the optimized foil or layer dimensions to estimate

winding copper losses under any given arbitrary current waveform. In [56] and [57], the optimum number of strands of Litz wire is obtained by Ferreira and Dowell for slackly and closely packed wires. These studies provide in-depth analyses to estimate power losses and allow developing detailed loss models that are used in the magnetic design models. In [61], it is shown that using both analytical and finite elements aided design procedure is more effective for magnetic design optimization. The flux density and core losses are calculated for a single operating point in analytical methods, while they can be calculated for different operating points in Finite Elements Method (FEM) calculations. Although detailed studies on various aspects of coil and core design have been provided in the literature, there are very few studies that provides a multi-objective meta-heuristic optimization of high-frequency air-gapped magnetics [55]. The selection of optimal frequency range in LLC considering the transformer volume and weight has been studied in [62], and multi-objective optimization of high frequency transformers using GAs has been proposed in [63], where the core cross-section, window area product, and the power loss have been chosen as fitness functions while the frequency and maximum flux density have been selected as optimization variables. However, the impact of the stray components have not been considered. In [64], high frequency power transformer has been optimized using Non-dominated Sorting Genetic Algorithm II (NSGA-II) by taking the constraints, such as the core saturation flux density and winding structure, into account. However, only the power loss and leakage inductance have been considered as the fitness function.

In [65], the optimization of the planar transformer of LLC converter using a metaheuristic method such as the Genetic Algorithm (GA), Ant Colony Optimization method (ACO), Shuffled Frog Leaping Algorithm (SFLA). According to the analytical design tool proposed in [58], more efficient planar transformers with lower power loss and core volume can be designed. This study, on the other hand, does not consider multi-core structure, different core materials and winding types, cost and thermal handling of the core.

Multiple types of two-phase interleaved LLC converters, referenced in [66–68], have their advantages such as improved thermal performance and better efficiency at light loads but come with the downsides of requiring extra switches, components, and complex control algorithms for current sharing. This results in increased system costs

and reduced efficiency. On the other hand, three-phase LLC resonant converters stand out as a more efficient and cost-effective alternative for high-power DC/DC conversion, as outlined in studies [45,46]. These converters benefit from the perks of a three-phase system, including better magnetic integration, decreased conduction losses, decreased EMI, and smaller output voltage ripples.

In addition, the adoption of high-frequency operation has been shown to downsize the passive components. Planar trans-formers, as discussed in references [69–73], are particularly effective at high-frequency operation. They are characterized by their low-profile, predictable parasitic components, ease of manufacturing, and consistent construction quality. Unlike conventional wire-wound transformers, planar transformers typically feature windings made of copper sheets etched onto a PCB in a spiral pattern. These transformers allow for a more efficient, heavily interleaved design of primary and secondary windings, especially in a manufacturing or automated setting. This integration of three-phase LLC resonant converters with planar transformers at high-frequency operation provides a promising path for more efficient, compact, and reliable high-power charging systems.

The optimization of LLC converters is more difficult than conventional PWM controlled converters due to their variety of operating modes and non-linear dynamics [74]. Many designs and optimization methodologies have been proposed for this type of converter in the past decades. Optimal design methods are developed based on the operation mode analysis in [74, 75], but these approaches require utilizing sophisticated calculation tools. A simple and highly accurate design procedure is presented in [76], but the wide output voltage range and load profile has not been discussed. Based on the characteristics of the battery charging profiles, time-weighted average efficiency index is proposed in [77], which represents the average weight of conversion efficiency during battery charging period. However, the presented design methodology only considers the power losses without taking cost and volume into account. The wide adjustable output voltage range for lithium-ion battery charger systems has been analyzed in [78–80] considering the charging profiles and corresponding worst-case scenarios for the primary side ZVS operation. On the other hand, all these research efforts are only towards finding optimum resonance tank design under limited design criteria. To obtain optimal solution across multiple objectives with varying design frequency, the designers

still require a multi-objective optimization method considering the design frequency, power losses, system cost and volume. A comprehensive optimization design procedure considering magnetic design, power switches and associated heat-sinks has not been proposed for LLC resonance converter operating under wide load range.

Several small signal equivalent circuit models are employed through various approaches for pulse width modulation (PWM) converters; these include voltage mode control [81, 82], current mode control that encompasses inductor current sideband [83–86], and V^2 control which integrates both inductor current sideband and capacitor voltage sideband [87–89]. For resonant converters, the state-space averaging method is not preferred since some state variables lack direct current (DC) components and contain switching frequency harmonics. Due to oscillations arising in resonance situations, the switching frequency interacts with the resonance frequency. This interaction is elucidated through beat frequency dynamics, wherein a pair of double poles located at the beat frequency determines the high-frequency response [90]. Consequently, an extended describing function method that evaluates the states of the resonance frequency in relation to the switching frequency offers the most effective approach for a robust LLC controller design [91].

1.3 Hypothesis

This thesis hypothesizes that a more energy-efficient, cost-effective, and thermally optimized LLC resonant converter design for LEV applications can be realized through a novel systematic design framework. This framework diverges from conventional design practices by incorporating a multi-faceted approach that accounts for user-specific charging profiles, thermal resistances, and a high-level system fitness function encompassing power loss, system volume, and cost.

The central tenets of the hypothesis are as follows:

- **Magnetic Design Optimization:** The magnetic design and switch selection, when carried out by considering total energy optimization reflective of user charging profiles, will result in a system that operates more efficiently across varying conditions as opposed to designs optimized for worst-case scenarios or nominal points.

- **Thermal Management Efficiency:** The inclusion of thermal resistances of power switches and corresponding heat-sink volumes in the design process will significantly improve the thermal management of the system, ensuring sustained performance under a range of operational temperatures.
- **Material Selection:** An optimal effective relative permeability for gapped magnetic cores will be determined, enhancing the magnetic efficiency and thereby reducing the size and cost of the magnetic components.
- **Transformer Structure Advancement:** A primary-series secondary-parallel connected multi-core transformer structure will offer better performance in terms of power density and efficiency than traditional single-core solutions.
- **Switching Frequency Accuracy:** New derivations of current equations relative to switching frequency will yield a more accurate prediction of converter behavior, allowing for improved control and performance.
- **System Fitness Maximization:** A high-level system fitness function that evaluates power losses, system volume, and cost will enable the identification of the most optimal design parameters, leading to an LLC resonant converter that is superior in terms of overall system performance and cost-efficiency.

Moreover, this thesis asserts that by incorporating parasitic effects and employing a closed-loop control mechanism within the design, the converter will demonstrate enhanced stability and accuracy in operation, particularly during transient conditions. The application of an extended describing function method for modeling the LLC converter is expected to accurately predict the beat frequency dynamics and small signal behavior, which are crucial for the effective control of the converter.

2. ANALYTICAL MODEL OF LLC RESONANT CONVERTER

Various resonant-converter designs exist, but they fundamentally work alike. They use a square pulse of voltage or current, created by power switches, to activate a resonant circuit. Within this circuit, energy flows and a portion or all of this energy is subsequently drawn out to power the output [31].

The Figure 2.1 illustrates a common LLC resonant converter, which consists of three key components:

1. The first part includes a DC voltage source and four complementary switches that function as a square wave generator. This generator creates a square-wave voltage, is resonant tank input voltage (v_{tin}), by alternating the activation of switch pairs S_1 - S_4 and S_2 - S_3 with 50% duty cycles for each pair. A brief pause between switches is necessary to prevent cross-conduction and to allow for ZVS to occur.
2. The resonant circuit, or resonant network, is made up of the C_r , and two inductors: the series L_r and the transformer's L_m , with a transformer turns ratio of n . This network circulates electrical current, transferring energy to the load via the transformer. The transformer's primary winding receives a V_{tin} , which is then transferred to the secondary side. The transformer provides electrical isolation and adjusts the voltage level to the output requirements through its turns ratio.
3. On the secondary side of the converter, two diodes form a full-wave rectifier to change Alternating Current (AC) input into Direct Current (DC) output for the load R_L . The output capacitor helps to smooth out the rectified voltage and current. This rectifier network can be configured either as a full-wave bridge or a center-tapped arrangement with a capacitive output filter. Alternatively, MOSFETs can be used for synchronous rectification, reducing conduction losses, which is particularly advantageous in applications involving low voltage and high current.

This resonant network features two distinct resonant frequencies. The primary resonant frequency (f_0) is set by the interaction between the L_r and C_r . On the other hand, the

secondary resonant frequency (f_m) is defined by the combined inductances of L_r and L_m along with C_r . The formulas to calculate f_0 and f_m are provided in equations (2.1) and (2.2), respectively.

$$f_0 = \frac{1}{2\pi\sqrt{L_r C_r}} \quad (2.1)$$

$$f_m = \frac{1}{2\pi\sqrt{(L_r + L_m)C_r}} \quad (2.2)$$

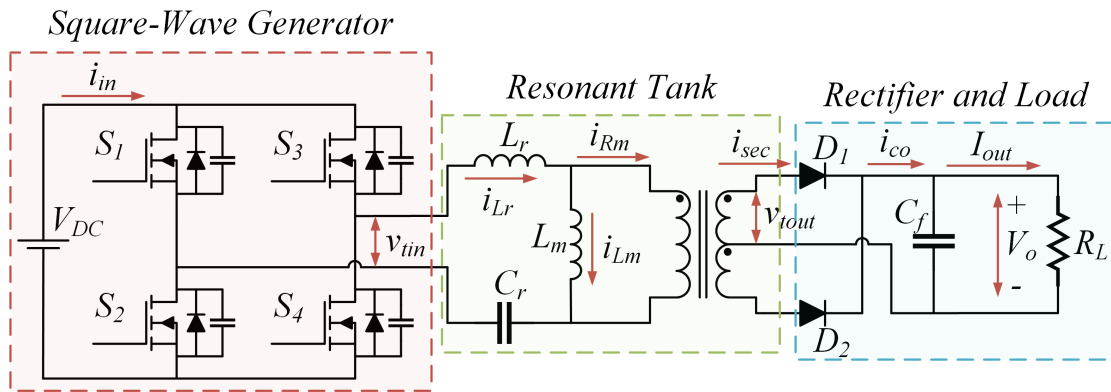


Figure 2.1 : Three stage configuration of LLC resonant converter.

2.1 Sinusoidal Analysis of LLC Resonant Converter

To design a converter capable of varying energy transfer and regulating output voltage, it's essential to understand the mathematical relation between input and output voltages, known as the conversion ratio. The First Harmonic Approximation (FHA) method is useful here, as it allows only the first harmonic to pass through the resonant network, helping to establish the gain, or the input-to-output voltage transfer function. For ease of analysis, the leakage flux and parasitic effects on the transformer's secondary side are usually disregarded. The operation of the LLC converter is summarized in Figure 2.2. When switches S_1 and S_4 are ON, v_{tin} is equal to the DC input voltage V_{DC} . Conversely, when switches S_2 and S_3 are ON, v_{tin} equals the negative of the V_{DC} . The resonant tank primarily responds to the fundamental component of v_{tin} , so v_{tin} is represented by this component, with higher-order harmonics being ignored. The Fourier Series representation of v_{tin} is detailed in equations (2.3)-(2.7) [92].

$$v_{tin_k} = D_0 + \sum_{k=1}^{\infty} [C_k \cos(k\omega t) + S_k \sin(k\omega t)] \quad (2.3)$$

$$D_0 = 0 \quad (2.4)$$

$$C_k = 0 \quad (2.5)$$

$$S_k = \frac{2}{T} \int_0^{\frac{T}{2}} V_{DC} \sin(k\omega t) dt = \frac{4V_{DC}}{k\pi} \sin(k\omega t) \quad (2.6)$$

$$v_{tin_1} = \frac{4V_{DC}}{\pi} \sin(\omega t) \quad (2.7)$$

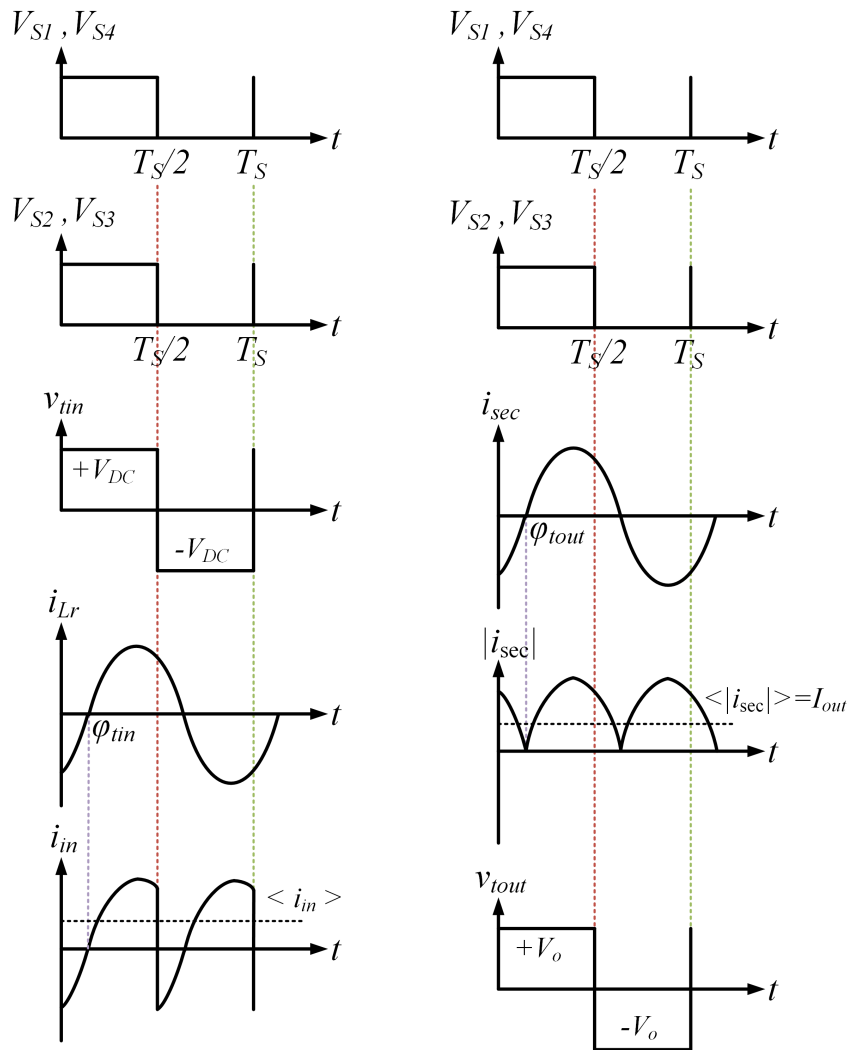


Figure 2.2 : The operation of LLC resonant converter.

Resonant tank current (i_{Lr}) is a sinusoidal current in a phase shift φ_{tin} with respect to the fundamental component of the v_{tin} as shown in equation (2.8).

$$i_{Lr} = I_{Lr} \sin(\omega t - \varphi_{tin}) \quad (2.8)$$

When S_1 and S_4 are conducted, input current of switch network (i_{in}) equals to the i_{Lr} and when S_2 and S_3 are conducted, i_{in} equals to $-i_{Lr}$. Then the average of i_{in} is obtained as equation (2.9).

$$\langle i_{in} \rangle = \frac{2}{T} \int_0^{\frac{T}{2}} I_{Lr} \sin(\omega t - \varphi_{tin}) dt = \frac{2I_{Lr}}{\pi} \cos(\varphi_{tin}) \quad (2.9)$$

The output rectifier of the circuit is controlled by the tank output current (i_{sec}), which is almost sinusoidal in nature, and the tank output voltage (v_{tout}). Importantly, v_{tout} is in phase with i_{sec} . Furthermore, a filter capacitor (C_f) is installed at the DC output. This setup ensures that the output voltage (V_o) has minimal harmonics related to the switching frequency, leading to a smoother and more stable output. While i_{sec} is positive, D_1 diode is conducted and v_{tout} equals to $+V_o$. While i_{sec} is negative, D_2 diode is conducted and v_{tout} equals to $-V_o$. v_{tout} is expressed for the first harmonic in the equation (2.10).

$$v_{tout} = \frac{4V_o}{\pi} \sin(\omega t - \varphi_{tout}) \quad (2.10)$$

The average value of the rectified tank output current ($\langle |i_{sec}| \rangle$) needs to match the steady-state load current (I_{out}), as indicated in equation (2.11). This requirement arises because the C_f does not allow for the passage of any DC current.

$$\langle |i_{sec}| \rangle = I_{out} = \frac{2}{T} \int_0^{\frac{T}{2}} I_{sec} \sin(\omega t - \varphi_{tout}) dt = \frac{2I_{sec}}{\pi} \quad (2.11)$$

The center tapped rectifier is like a resistive load (R_e) to the tank circuit as v_{tout} and i_{sec} are in phase with each other. Therefore, R_e can be expressed as in equation (2.12).

$$R_e = \frac{v_{tout}}{i_{sec}} = \frac{8R_L}{\pi^2} \quad (2.12)$$

The Figure 2.3 displays the circuit of the resonant tank along with its reduced secondary circuit. The magnitude of resonant tank transfer function (H_s) is derived and presented in equation (2.13). This function is expressed in terms of the inductance ratio ($L_n = L_m/L_r$), quality factor (Q), and frequency ratio ($F = f_s/f_0$) where f_s is the switching frequency. Q is given in equation (2.14). Furthermore, the Figure 2.4 illustrates the equivalent circuit of the converter as derived using the FHA method. The voltage gain (M) achieved by this configuration is detailed in equation (2.15).

$$H_s = \frac{1}{\sqrt{\left(1 + \frac{1}{L_n} \left(1 - \frac{1}{F^2}\right)\right)^2 + \left(\left(F - \frac{1}{F}\right) Q\right)^2}} \quad (2.13)$$

$$Q = \frac{\sqrt{\frac{L_r}{C_r}}}{R_e} \quad (2.14)$$

$$M = \frac{nV_o}{V_{DC}} = H_s \quad (2.15)$$

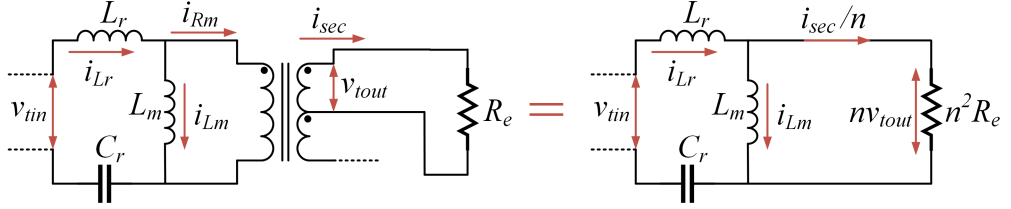


Figure 2.3 : The FHA model of LLC resonant converter.

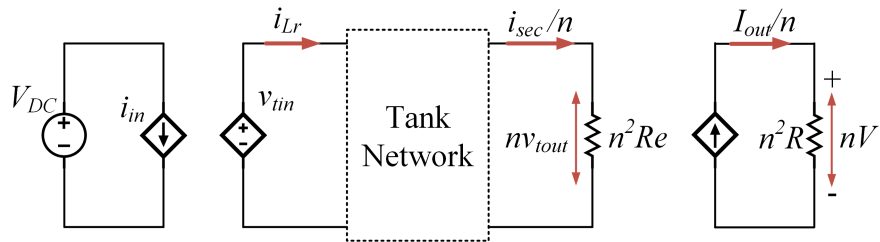


Figure 2.4 : The resonant tank equivalent circuit of LLC resonant converter.

Understanding the behavior of H_s in relation to three key factors - F , L_n , and Q - is crucial. In the gain function, F serves as the control variable. L_n and Q are considered dummy variables because their values are determined by their physical parameters and remain fixed thereafter. Once the design is finalized, H_s is adjusted by varying F .

A practical approach to illustrate how the gain function responds is to plot H_s against F for a set of predetermined values for L_n and Q . For any specific values of L_n and Q , H_s typically displays a convex curve shape near the circuit's resonant frequency, indicative of a resonant converter's gain characteristic. Altering L_n and Q modifies the shape of the H_s curve in relation to F .

Q , being a function of the load as per equation (2.12) and (2.14), means that H_s represents a series of curves that link frequency modulation to changes in the load. When L_n is constant, an increase in Q narrows the curve, leading to a tighter frequency-control range. This is expected since Q reflects the quality factor of the series resonant circuit. Conversely, for a constant Q , reducing L_n compresses the curve, causing the entire curve to tighten and the f_m to shift closer to the f_0 . This adjustment results in an improved frequency-control band with a higher peak gain.

The resonant network's nature, whether inductive or capacitive, is determined by its input impedance (Z_{in}). The phase of the Z_{in} is the key indicator: if it exceeds 180° , the converter operates inductively. Conversely, if the phase is less than 180° , the converter operates capacitively. The critical juncture between these two modes is marked by the 180° phase line.

The Figure 2.5 illustrates that all gain curves peak at points defining the boundary between the inductive and capacitive impedances of the resonant tank. These peaks allow for the demarcation of inductive and capacitive operational regions, as indicated by shaded areas in the plot. The goal is to maintain inductive operation across all input voltage and load current ranges, avoiding capacitive operation. This is crucial because ZVS is only achievable in the inductive region. In contrast, capacitive operation, where current leads voltage, can lead to undesirable effects. In such a scenario, the current in the MOSFET would reverse direction before the MOSFET turns off. Subsequently, once the MOSFET is off, the reverse current would flow through the MOSFET's body diode. This leads to hard commutation on the body diode when another MOSFET in the bridge turns on, causing reverse recovery losses and noise. It could also result in high current spikes and potential device failure.

The functioning of an LLC resonant converter is primarily defined by the relationship between the f_s and f_0 . The gain of the LLC network is modulated by frequency,

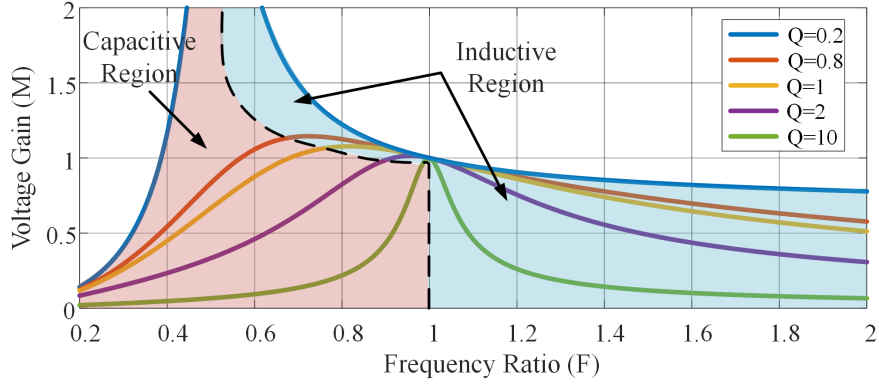


Figure 2.5 : Gain curves of LLC resonant converter.

allowing the converter to operate in three distinct modes based on the input voltage and load current conditions: resonant operation ($f_s = f_0$), below-resonant operation ($f_s < f_0$), and above-resonant operation ($f_s > f_0$).

The Figure 2.6 shows typical waveforms for an LLC resonant converter operating at, below, or above the f_0 . The waveforms, displayed sequentially from top to bottom, include the gate voltage of S_1 and S_4 , the gate voltage of S_2 and S_3 , the i_{Lr} , i_{Lm} and the rectified output current i_{co} .

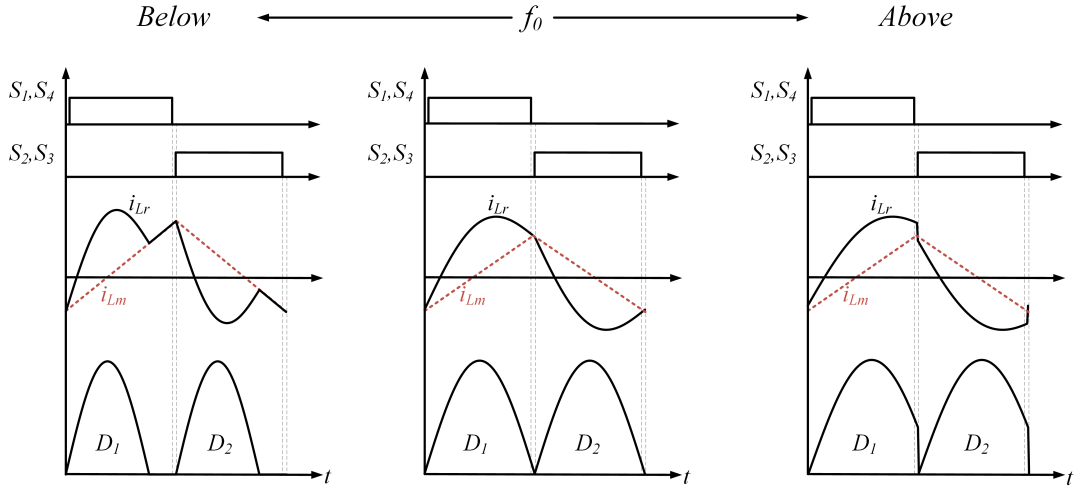


Figure 2.6 : Operation modes of LLC resonant converter.

It's important to note that the primary-side current is the sum of the magnetizing current (i_{Lm}) and the secondary-side current, as referred back to the primary side. However, the magnetizing current, which only flows on the primary side, does not contribute to the power transfer from the primary-side source to the secondary-side load. This distinction is crucial for understanding the converter's operation and its efficiency in power transfer.

In the operation at resonance mode, the switching frequency of the LLC resonant converter is set to match the series resonant frequency. In this mode, each half of the switching cycle encompasses a full power delivery process, where the resonant half-cycle is completed within the duration of the switching half-cycle. At the end of each switching half-cycle, the current in the resonant inductor aligns with the magnetizing current, and the current through the rectifier drops to zero. This synchronization ensures that the resonant tank operates with unity gain, achieving the most optimized performance and efficiency. Consequently, the transformer's turns ratio is specifically designed to enable the converter to operate at this resonance point under nominal input and output voltage conditions. This design consideration is key to maximizing the efficiency and effectiveness of the LLC resonant converter when operating at resonance.

In the operation below resonance mode, the resonant current decreases to match the magnetizing current before the driving pulse width ends. This results in the cessation of power transfer, although the magnetizing current continues. Operating below the series resonant frequency can still achieve ZVS on the primary side and ensure soft commutation of the rectifier diodes on the secondary side. However, in this mode, the secondary-side diodes operate in a discontinuous current mode, necessitating more circulating current in the resonant circuit to deliver the same energy to the load. This increased current leads to higher conduction losses on both the primary and secondary sides. A notable point is the potential loss of primary ZVS if the switching frequency is too low, which would result in increased switching losses and related problems [93].

In the operation above resonance mode, the primary side has a lower circulating current in the resonant circuit, reducing conduction losses as the resonant circuit's current remains in continuous-current mode. This results in less Root Mean Square (RMS) current for the same load. However, the rectifier diodes in this mode do not benefit from soft commutation and suffer from reverse recovery losses. Despite this, operating above the resonant frequency can still achieve primary ZVS. One issue with this mode is the significant increase in frequency under light-load conditions [93].

From this discussion, it's clear that the converter can be designed for operation at either $f_s > f_0$ or $f_s \leq f_0$, or by varying f_s around f_0 . Optimal operation is typically found near the f_0 , where the advantages of the LLC converter are most pronounced. In

regions where $f_m < f_s < f_0$ (with f_s close to f_0), and $f_s > f_0$, the freewheeling diodes (or body diodes) turn off at zero current and zero voltage, minimizing the need for fast reverse-recovery characteristics. In such scenarios, MOSFETs can be used as switches in applications requiring high switching frequencies.

2.2 Frequency Dependent Current Model of LLC Converter

The power losses in an LLC resonant converter are primarily influenced by the design of the resonant tank. An efficiency-oriented optimal design hinges on understanding the relationships between the inductance-capacitance parameters and power losses. However, the operational modes of resonant converters vary depending on the switching frequency and load conditions, making it challenging to accurately and simply estimate conduction losses.

If we assume a purely sinusoidal current flows through the resonant circuit and the circuit operates exactly at the f_0 , the design results are quite accurate. However, a design that operates solely at f_0 is impractical; deviating from f_0 introduces discrepancies compared to bench-test measurements. The FHA design approach is commonly understood to provide an initial design framework, maintaining clear physical concepts and meanings of design variables, which is crucial for designers to grasp the behavior of the LLC converter.

For precise magnetic design, it's necessary to know the RMS and peak values of the resonant current, the peak value of the magnetizing current, and the RMS value of the secondary-side currents. These operating currents vary with the LLC's "below," "at," and "above" resonance operating conditions. Thus, all current equations should be expressed in terms of the f_s .

When the f_s is below the f_0 , the conduction time of each rectifier diode is nearly equal to the half period of the resonance. Based on this, Certain assumptions can be made. For instance, assume that the conduction time of a rectifier diode equals the half period of the resonance and the diode current waveform is a half-cycle sinusoid. The Figure 2.7 illustrates simplified shapes of current waveforms in the LLC resonant converter based on these assumptions.

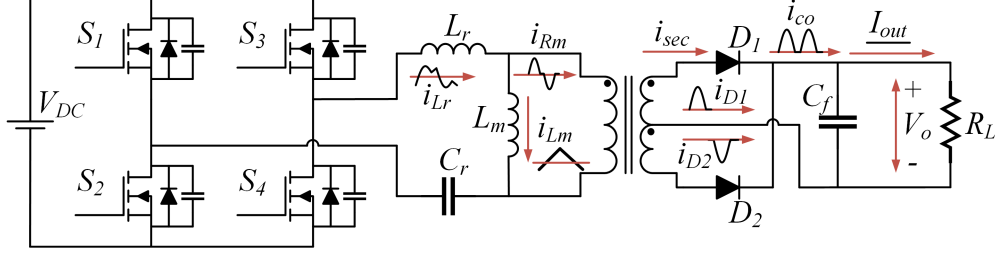


Figure 2.7 : Simplified shapes of current waveforms of LLC resonant converter.

The current equations for the LLC resonant converter operating below the f_0 are derived from these assumptions. Interestingly, these frequency-dependent equations are also valid for operations at and above the resonant frequency. This universality underscores the importance of understanding the fundamental principles of LLC resonant converter operation across different frequencies and conditions.

The currents of D_1 and D_2 is given by

$$i_{D1}(t) = \begin{cases} nI_{Rm} \sin(\omega_0 t) & 0 \leq t < \frac{T_0}{2} \\ 0 & \frac{T_0}{2} \leq t < T_s \end{cases} \quad (2.16)$$

$$i_{D2}(t) = \begin{cases} 0 & 0 \leq t < \frac{T_s}{2} \\ nI_{Rm} \sin(\omega_0 t) & \frac{T_s}{2} \leq t < \frac{T_s + T_0}{2} \\ 0 & \frac{T_s + T_0}{2} \leq t < T_s \end{cases} \quad (2.17)$$

where ω_0 is $2\pi f_0$ and I_{Rm} is the primary-side coupled current of the transformer. According to ampere-sec balance and Kirchhoff's Current Law (KCL), I_{Rm} can be found as

$$I_{Rm} = \frac{\pi I_{out}}{2n(\frac{f_s}{f_0})} \quad (2.18)$$

Under identical load conditions, the influence of high-order harmonic components becomes more significant as the f_s diverges from the f_0 . This increase in the prominence of high-order harmonics results in higher peak currents in the rectifier diodes. The ratio f_s/f_0 is indicative of the weight or contribution of these high-order harmonics in the overall current profile.

The peak current of the rectifier diode shows an inverse relationship to the ratio f_s/f_0 . This implies that, for the same load, the current through the diode increases as f_s moves further away from f_0 . Essentially, the further the f_s strays from the f_0 , the

greater the impact of high-order harmonics, leading to a corresponding increase in the peak currents experienced by the rectifier diodes. This relationship highlights the importance of carefully considering the frequency operation in relation to the resonant frequency to manage and mitigate potential issues with diode currents in LLC resonant converters [94].

The average value of diode current $\langle i_{D1} \rangle$ is

$$\langle i_{D1} \rangle = \frac{I_{out}}{2} \quad (2.19)$$

The RMS value of diode current i_{D1} is

$$I_{D1_{RMS}} = \frac{\pi I_{out}}{4} \sqrt{\frac{f_0}{f_s}} \quad (2.20)$$

C_f is given by

$$i_{Cf}(t) = \begin{cases} nI_{Rm} \sin(2\omega_0 t) - I_{out} & 0 \leq t < \frac{T_0}{2} \\ -I_{out} & \frac{T_0}{2} \leq t < \frac{T_s}{2} \end{cases} \quad (2.21)$$

The magnetizing inductance in an LLC resonant converter plays a role in the resonance process specifically when the resonant inductor current equals the magnetizing current. This specific condition is when L_m actively contributes to the resonance phenomenon within the circuit.

To simplify the analysis, the current flowing through L_m is often approximated using a triangular function. This approximation is useful for analytical purposes, as it simplifies the complex behaviors in the circuit, making it easier to understand and predict the performance of the converter under various operating conditions. The triangular approximation represents the current waveform through L_m , which is a practical approach for modeling and analyzing the dynamics of the LLC resonant converter. The expression of i_{Lr} is given in equation 2.22.

$$i_{Lr}(t) = \begin{cases} \frac{nV_o}{L_m} t - \frac{nV_o}{L_m} \frac{T_s}{4} + \frac{1}{n} i_{D1}(t) & 0 \leq t < \frac{T_s}{2} \\ \frac{nV_o}{L_m} \frac{T_s}{4} - \frac{nV_o}{L_m} \left(t - \frac{T_s}{2} \right) - \frac{1}{n} i_{D2}(t) & \frac{T_s}{2} \leq t < T_s \end{cases} \quad (2.22)$$

The RMS value of i_{Lr} is derived from equation (2.22)

$$I_{Lr_{RMS}} = \sqrt{\frac{1}{48} \left(\frac{nV_o}{f_s L_m} \right)^2 + \frac{\pi^2}{8} \left(\frac{I_o}{n} \sqrt{\frac{f_0}{f_s}} \right)^2 - I_{out} \frac{V_o}{L_m} \frac{1}{2} \left(\frac{1}{f_s} - \frac{1}{f_0} \right)} \quad (2.23)$$

The RMS value of secondary current is derived from equation (2.20)

$$I_{sec_{RMS}} = \sqrt{2} I_{D1_{RMS}} = \frac{\sqrt{2} \pi I_{out}}{4} \sqrt{\frac{f_0}{f_s}} \quad (2.24)$$

2.3 Design and Loss Model of Magnetics

The design of the resonant elements in LLC is vital to achieve soft-switching. The most widely used method to design the resonant tank is the FHA. L_m , L_r and C_r values are expressed by equations (2.25)-(2.27) using FHA method.

$$C_r = \frac{1}{2\pi f_0 R_e Q} \quad (2.25)$$

$$L_r = \frac{1}{(2\pi f_0)^2 C_r} \quad (2.26)$$

$$L_m = L_n L_r \quad (2.27)$$

In the design of LLC resonant converters, the transformer and inductor play pivotal roles and their design is closely tied to specific values of inductances. The transformer is designed based on the magnetizing inductance value, while the design of the inductor is dependent on the resonant inductance value. One significant challenge in this design process is dealing with the leakage inductance and the risk of saturation in the magnetic material of the transformer. Leakage inductance can be difficult to control precisely, and saturation of the magnetic core material can lead to a decrease in inductance and inefficient operation. To mitigate these issues, it's common practice to design a separate resonance inductor instead of integrating it within the transformer. This approach allows for more precise control over the inductance values and helps avoid the complications associated with leakage inductance and core saturation, ensuring more reliable and efficient operation of the LLC resonant converter.

In a conventional transformer design, the magnetizing inductance is typically quite large, meaning that a small current through the winding can easily build up flux in the magnetic core. This is a key aspect of traditional transformer operation, where the design focuses primarily on efficient power transfer.

However, in the context of an LLC resonant converter, the design considerations for the transformer are quite different. Here, the magnetizing inductance is specifically defined to ensure the proper performance of the circuit. To achieve the desired L_m value and control the flux within the core, a gap is introduced into the core of the transformer. This introduction of a gap significantly alters the behavior of the magnetic core, reducing its permeability and thereby the magnetizing inductance.

As a result of this modification, the design of the transformer in an LLC resonant converter shifts from a focus on power transfer (as in conventional transformers) to a focus on stored energy, akin to inductor design. This approach is necessary to tailor the transformer's characteristics to the unique requirements of the LLC resonant converter, which includes precise control over the resonant behavior of the circuit.

The magnetic model for both the transformer and the inductor in an LLC resonant converter is thus developed based on the given values of L_m and L_r . This modeling is essential to ensure that both components effectively fulfill their roles in the resonant circuit, contributing to the overall efficiency and performance of the converter.

The circuit model of Gapped transformer in the LLC resonant converter as shown in Figure 2.8(a). Three windings are wound around the core, they are the primary winding on the left side, the secondary 1 and secondary 2 windings on the right side as shown in Figure 2.8(b). In an LLC resonant converter, the current dynamics within the transformer are quite distinct compared to a conventional transformer. The current through the primary winding, referred to as the resonant current i_{Lr} , is a combination of the magnetizing current and the primary current through the ideal transformer. This primary current is what balances the secondary load current i_{Rm} . On the secondary side, there are two currents, i_{s1} and i_{s2} , which are complementary to each other.

The magnetizing current flowing through the magnetizing inductance is responsible for establishing the magnetic flux within the transformer. In contrast, the primary current is mainly associated with the losses in the transformer. This distinction is important

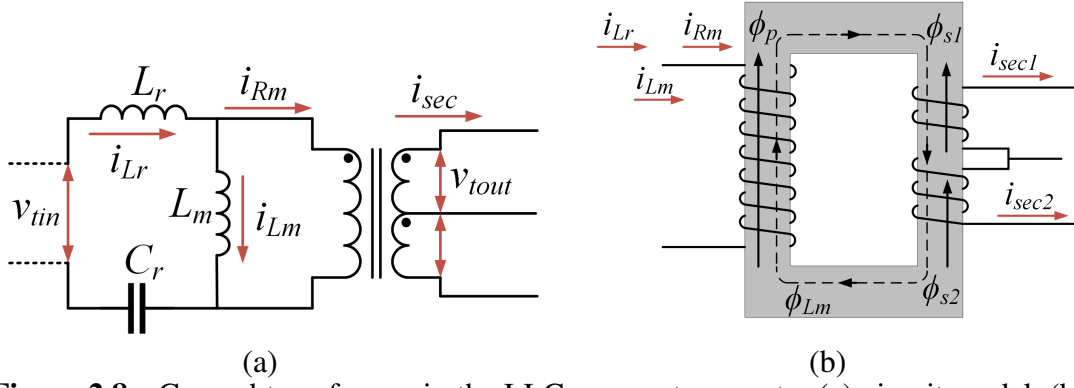


Figure 2.8 : Gapped transformer in the LLC resonant converter (a) circuit model, (b) simplified fabrication sketch of the gapped transformer.

because, in the LLC resonant converter's transformer design, both the magnetizing current and the primary current need to be considered for the overall sizing. This differs from a traditional inductor, where the winding current also serves as the magnetizing current in a single winding setup [95].

The magnetizing current not only sets up the common flux linking all the windings but also directly influences the flux density within the core. The flux density is directly proportional to i_{Lm} . To quantify this relationship, the maximum flux density in the core for both the transformer and the inductor can be calculated using specific formulas, designated as equation (2.28) for the transformer and equation (2.29) for the inductor. These formulas take into account the core material properties and the design parameters of the transformer or inductor, ensuring that the core operates within its optimal magnetic limits and does not reach saturation under normal operating conditions [96].

$$B_{\max_tr} = \frac{\mu_0 \mu_{eff} N_p I_{Lm_{\max}}}{l_c}, \quad (2.28)$$

$$B_{\max_ind} = \frac{\mu_0 \mu_{eff} N_{ind} I_{Lr_{\max}}}{l_c}, \quad (2.29)$$

where N_p is the number of turns in the primary winding, N_{ind} is the number of turns in the inductor winding, $I_{Lm_{\max}}$ is the peak value of magnetizing current, μ_{eff} and l_c are the effective relative permeability and the magnetic path length of the core, respectively.

During the design process, the total loss (P_{total}) is defined as a factor of the DC copper losses (P_{cu}). The definition is

$$P_{total} = (1 + \gamma)P_{cu} \quad (2.30)$$

where γ is used to represent the total losses except the DC copper loss. The total winding losses for transformer is given equation (2.31), where R_{DC_p} and R_{DC_s} are the DC resistance of the primary and secondary winding, respectively.

$$P_{cuDC} = R_{DC_p} I_{LrRMS}^2 + R_{DC_s} I_{secRMS}^2 \quad (2.31)$$

When an alternating current flows through an isolated conductor, it induces an alternating magnetic field. This field, in turn, generates eddy currents within the conductor. These eddy currents are created to counteract the field produced by the original current. The interaction between the original current, the magnetic field it creates, and the induced eddy currents leads to a notable phenomenon known as the skin effect. The skin effect is characterized by a non-uniform distribution of current density across the cross-section of the conductor. Specifically, the current density is higher near the surface of the conductor compared to its center. This uneven distribution occurs because the induced eddy currents oppose and reduce the original current in the center of the conductor, pushing the majority of the current towards the surface. Due to the skin effect, the effective area through which the current flows is reduced. This smaller conduction area leads to an increase in resistance and, consequently, greater losses in the conductor. The concept of skin depth is integral to understanding the skin effect. Skin depth in equation (2.32) is defined as the distance from the outer surface of the conductor to the depth at which the electromagnetic wave, traveling within the lossy conductor, decreases to $1/e$ (approximately 37%) of its original value at the surface where μ_0 is the magnetic permeability of free space with the value of $4\pi 10^{-7}$ H/m for a conductor with conductivity σ and operating at frequency f_s . Here, "e" represents the base of the natural logarithm. This measure gives an indication of how deeply into the conductor the AC current effectively penetrates, and it varies depending on factors such as the frequency of the AC current and the electrical properties of the conductor material.

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}} \quad (2.32)$$

In the context of inductors or transformers, where there are multiple winding turns, the interaction between conductors becomes an important consideration. The current flowing through one conductor generates a magnetic field that influences the current in adjacent conductors. This interaction is particularly pronounced when dealing with the skin and proximity effects in high-frequency applications, such as in LLC resonant converters.

The proximity effect complements the skin effect, especially at high frequencies. It refers to the tendency of the alternating current in one conductor to redistribute the current flow in a neighboring conductor, further contributing to winding losses. Both the skin and proximity effects lead to what is collectively known as eddy current losses in the windings.

Given the high switching frequency of LLC resonant converters, these high-frequency effects on winding loss are significant and cannot be ignored, particularly in applications involving high current and high frequency. Consequently, different types of winding and configurations, such as foil or round wire, should be carefully considered, especially for the secondary side of the transformer or inductor.

To mitigate these effects, two common solutions are employed: Litz wire and interleaving windings. Litz wire is made up of numerous small, insulated strands woven together. This design helps in reducing both skin and proximity effects by evenly distributing the current among the strands, effectively increasing the surface area for current flow. Interleaving windings, on the other hand, involves a specific arrangement of the windings to minimize the mutual magnetic coupling between them, thereby reducing the proximity effect.

The equation labeled as (2.33), where r is the radius of the copper, can be used to calculate the AC resistance of the round wire considering the skin effect. These calculations and considerations are vital in designing efficient, high-performance LLC resonant converters, particularly for managing winding losses at high frequencies.

$$R_{AC_{round}} = R_{DC} \left[1 + \frac{\left(\frac{r}{\delta}\right)^4}{48 + 0.8\left(\frac{r}{\delta}\right)^4} \right] \quad (2.33)$$

The investigation of eddy current losses, especially in transformers with foil windings, is a complex topic that has been extensively studied. One notable contribution to

this field was made by Dowell, as presented in his work referenced in [97]. Dowell's work provides a one-dimensional (1D) solution to the relevant equations, incorporating appropriate assumptions for simplifying the complex phenomena of eddy currents in foil windings.

Further research, as cited in references [98] and [99], expanded on this topic by considering eddy current losses under conditions of non-sinusoidal excitation currents. This is particularly relevant in applications where the current waveform deviates from a simple sinusoidal shape, as is often the case in power electronics.

Building upon these studies, Hurley, in references [55], proposed a method for optimizing conductor thickness in transformer windings. This methodology is notable for not requiring the calculation of Fourier coefficients, which simplifies the optimization process. Hurley's approach is valuable in designing transformers with arbitrary current waveforms, where standard sinusoidal assumptions may not hold.

Utilizing equations (2.34)-(2.38), it's possible to calculate the AC resistance of foil windings and determine optimal foil thicknesses and lengths for various applications. Key parameters in these calculations include the optimum layer thickness (Δ_{opt}) which varies based on the current waveform, A_{ws} secondary winding area and ψ the AC resistance coefficient for p layers of winding. Additionally, d_{opt} and l_{opt} represent the optimum foil thickness and length, respectively.

$$\psi = \frac{5p^2 - 1}{15} \quad (2.34)$$

$$\Delta_{opt} = \sqrt[4]{\left(\frac{4 \left(\frac{f_s}{2f_0} \right)^2}{\psi} \right)} \quad (2.35)$$

$$d_{opt} = \Delta_{opt} \delta \quad (2.36)$$

$$l_{opt} = \frac{A_{ws}}{d_{opt}} \quad (2.37)$$

$$R_{AC_{foil}} = R_{DC} \left(1 + \frac{1}{3} \left(\frac{\Delta}{\Delta_{opt}} \right) \right) \quad (2.38)$$

These studies and methods provide crucial insights for engineers and designers in optimizing transformer designs, especially in high-frequency applications where eddy current losses can significantly impact performance and efficiency. By applying these principles, it's possible to tailor transformer and inductor designs to specific application needs, balancing factors such as size, weight, efficiency, and performance.

The total window utilization factor of all windings is defined as the ratio of the winding conduction area (W_c) to the window winding area of the core (W_a) and represented as k_u . For multiple windings transformers, the optimum window utilization factor for the primary winding and secondary windings may be investigated by the DC winding loss. If the primary winding has the window utilization factor of k_{up} , the total winding loss of the transformer can be obtained by

$$P_{cuDC} = \rho_w \left[\frac{N_p MLT}{\frac{k_{up} W_a}{N_p}} I_{LrRMS}^2 + 2 \frac{N_{s1} MLT}{\frac{(k_u - k_{up}) W_a}{2N_{s1}}} I_{secRMS}^2 \right] \quad (2.39)$$

where ρ_w is the electrical resistivity of the winding, MLT is the mean length of the turn, N_p and N_{s1} are the number of turns of the primary winding and the secondary winding 1. To find the best value of P_{cuDC} (the DC copper loss) in relation to k_{up} (the window utilization factor of the primary winding), you calculate the partial derivative of P_{cuDC} with respect to k_{up} and then set this derivative to zero. This approach helps to determine the relationship between k_{up} and the total window utilization factor, as explained by

$$k_{up} = k_u \frac{1}{1 + \frac{2I_{secRMS}}{nI_{LrRMS}}} \quad (2.40)$$

With the equivalent relative permeability of the gapped core μ_{eff} , the magnetizing inductance can be calculated by the basic inductance equation

$$L_m = \frac{\mu_0 \mu_{eff} N_p^2 A_c}{l_c}, \quad (2.41)$$

where A_c is the cross-sectional area of the core.

In magnetic components like transformers, the magnetic flux produced by one winding doesn't fully connect with all other windings, leading to some flux leakage through

the conductors or air space. This phenomenon is represented by leakage inductance, which signifies the imperfect coupling between windings. The leakage inductance in transformers can be calculated from the energy stored in the magnetic core's window area, where the excitations from the primary and secondary windings neutralize each other. The Magneto-Motive Force (MMF) and magnetic intensity are typically involved in determining this leakage energy [100]. Accurately estimating leakage inductance is crucial for LLC resonance converters, as a well-matched resonance frequency is essential, highlighted in [101]. The primary-side referred leakage inductance (L_{lk}) can be derived from the energy stored in the magnetic field, which involves factors like winding length (a), winding build (b), and insulation thickness (c).

$$L_{lk} = \frac{\pi(MLT)N_p^2}{a} \left(\sum c + \frac{\sum b}{3} \right) 10^{-6} \quad (2.42)$$

Another main power loss of the magnetic component comes from the magnetic core. The maximum flux density can be calculated by equation (2.29) and the core loss may be obtained with the Steinmetz equation

$$P_{fe} = V_c K_c f^\alpha B_{\max}^\beta, \quad (2.43)$$

where V_c is the core volume. Typical values of K_c , α , β are usually given by the core manufacturers or obtained from the loss curves.

P_{total} , as a function of the temperature rise, ΔT is

$$P_{total} \approx \frac{\Delta T}{R_\theta}, \quad (2.44)$$

where R_θ is the thermal resistance for the core assembly. R_θ can be calculated for each core by the empirical formula given in equation (2.45) for natural convection cooling.

$$R_\theta = \frac{0.06}{\sqrt{V_c}}, \quad (2.45)$$

To increase the power density of an LLC converter, designers often boost the switching frequency. This allows for smaller magnetics, but even with low-loss ferrite cores suitable for high frequencies, significant power losses can occur in the core and windings

due to skin and proximity effects, leading to thermal challenges. To address this, some researchers suggest using distributed cores rather than a single core, as shown in various studies [35, 48–53]. This approach spreads the copper losses and heat across multiple cores, easing the thermal load on the transformer and the secondary-side rectifiers. The specifics of this series-parallel connected distributed core structure are detailed in [48] and [49]. Research in [50] and [51] demonstrated that using two smaller transformers instead of one large one can increase the converter's power density by 24.28%, as further discussed in [52] and [53].

2.4 Loss Model of Power Switches and Heat-sink Size Determination

The total loss of the switches are including the switches in the full bridge and the switches in the or diodes in the rectifier. Since the duty cycle of the diagonal power switches are identical to that of the remaining diagonal power switches with 180° phase shift, the total power dissipation of the full-bridge leg can be found by

$$P_{cond_{pri,FB}} = 2R_{dson_{pri}} I_{Lr_{RMS}}^2 \quad (2.46)$$

In an LLC converter operating in the inductive region, the switching processes for MOSFETs and free-wheeling diodes are ZVS and lossless. This means when MOSFETs turn on and free-wheeling diodes turn off, they do so without incurring significant losses. Additionally, the turn-on losses for power diodes are minimal. Furthermore, the diodes used for rectification on the secondary side switch on and off at zero current, which eliminates additional switching losses. As a result, the dominant source of switching losses in the LLC converter comes from the turning-off of the MOSFETs [9]. Even though the turn-off power losses are not as high as the capacitive losses of a hard-switched converter, they can be considerably high when the switching frequency is low and magnetizing current is high. The total turn-off power losses are expressed as

$$P_{turn_{off,FB}} = 2V_{DC} I_{Lr} t_{fall} f_s \quad (2.47)$$

where,

$$t_{fall} = (R_g + R_{gext})C_{iss} \log\left(\frac{V_{plt}}{V_{th}}\right) \quad (2.48)$$

Here, R_g is the internal gate resistance, R_{gext} is the external resistance, C_{iss} is the effective input capacitance, V_{plt} is the gate plateau voltage, V_{th} is the threshold voltage. The other frequency dependent power loss component is the gate drive loss, which is expressed as

$$P_{gate,FB} = 4Q_g V_{gs} f_s \quad (2.49)$$

Another loss component, which may become significant in high frequency designs is the dead-time losses. Depending on the switching period, it is either inserted as a constant time or a percentage of the soft-switching time, and can be expressed as

$$P_{bodydiode,FB} = 4V_{sd} I_{sd} t_{bdc} f_s \quad (2.50)$$

Here, V_{sd} is the body diode voltage drop, I_{sd} is the average source to drain current during total body-diode conduction time (t_{bdc}) both during turn-on and turn-off. The total conduction losses on the secondary side synchronous switches of center-tapped configuration are expressed as

$$P_{condsec} = R_{dson,sec} I_{secRMS}^2 \quad (2.51)$$

In the case of using SiC diodes, the total conduction losses per center-tap configuration are expressed as,

$$P_{condiode} = V_f I_{out} \quad (2.52)$$

When increasing the number of transformers and power devices in parallel in each secondary side output stage of an LLC converter, both the output current and the RMS current in the secondary circuit decrease proportionally. This parallel arrangement effectively distributes the load among multiple transformers and power devices, reducing the current each has to handle. On the secondary side, the synchronous devices operate with ZCS during below-resonance operation. This means they turn

on and off without carrying current, reducing switching losses. In contrast, during above-resonance operation, the secondary side power devices turn on under current. However, the rate of change of current di/dt during this process is limited by the loop and leakage inductances in the circuit. This limitation helps manage the stress on the devices during the turn-on process, contributing to the overall efficiency and longevity of the converter.

To determine the appropriate size for a heat-sink in a power device, it's crucial to estimate the thermal resistance from the device's junction to the heat-sink. While Finite Element Analysis (FEA) tools offer a precise way to calculate total thermal resistance, they require extensive computational resources. For a more straightforward approach, the theory of iso-flux rectangles and strips on compound flux channels, as described in [102], can be applied to each layer. The thickness of the Die-Attached Pad (DAP) varies based on the die's construction and the package design. The entire thermal pad area is utilized for cooling and is filled with vias, each 8 mils in diameter. This extensive use of vias for cooling is particularly suitable for soft-switching converters, where the power loop isn't as critical as in hard-switching converters. In terms of construction specifics, it's assumed that both the top and bottom copper (Cu) layers of the PCB are $70\text{ }\mu\text{m}$ thick. The total thickness of the PCB is set at 1.6 mm. The vias are plated with $25\text{ }\mu\text{m}$ of copper and filled with conducting epoxy to enhance thermal conductivity and cooling efficiency. This design effectively supports the dissipation of heat generated during the operation of the power device.

3. SYSTEM LEVEL OPTIMIZATION BASED ON THE 3.7 kW LLC CONVERTER

LLC converters can obtain ZVS for zero to rated load with a low turn-off current for the primary switches, while achieving ZCS for the SRs. Soft turn-on and turn-off features allow minimizing the EMI [106]. In comparison to a soft switching PWM converters, LLC converters can work at a higher switching frequency with higher efficiency across the full load range, resulting in a higher power density and lower overall cost [25, 107]. However, as it is controlled with a variable switching frequency, the selection and design of the components remains as a difficult task, particularly for varying loads such as batteries in EVs [19]. The common approach to design an LLC converter is to determine the switching frequency range and the resonance frequency first, and then select the magnetics, power switches and heatsinks, accordingly [33]. The challenge with this approach is that the switching frequency is chosen based on designer's experience without conducting a systematic design framework [20]. The downside of this design procedure is clear; the switching frequency range and the resonance frequency of the converter, thereby, the resonance inductance, magnetizing inductance, and resonance capacitor values and the core material and size for magnetics are not optimally selected. This approach prohibits obtaining high power density as the total power losses and power density are determined based on the initially selected switching frequency range, and resonance frequency [34].

Optimizing LLC converters is more challenging than traditional PWM controlled converters due to their various operating modes and non-linear behavior [74]. Over the years, many design and optimization methods have been suggested for these converters. There are optimal design techniques based on operational mode analysis [74, 75], but they require advanced calculation tools. A simpler and more precise design process is described in [76], yet it doesn't cover wide output voltage ranges and load profiles. A time-weighted average efficiency index, considering battery charging profiles, has been introduced in [77]. This index focuses on the average efficiency during the battery charging period, but it overlooks factors like cost and volume.

The broad adjustable output voltage range for lithium-ion battery chargers and the related worst-case scenarios for primary side ZVS operation are analyzed in [78–80]. However, these studies mainly focus on optimizing the resonance tank design under limited criteria. For an optimal solution considering multiple objectives and design frequencies, designers still need a multi-objective optimization method that accounts for design frequency, power losses, system cost, and volume. Despite these efforts, a comprehensive optimization design method that includes magnetic design, power switches, and heat-sinks for LLC resonance converters operating across a wide load range has not been developed [21].

In this study, a systematic design framework for LLC converter is proposed to find the optimal resonance frequency, minimum and maximum switching frequency limits that satisfy the required voltage gains at corresponding quality factors by minimizing the system level fitness function. The high-level block diagram summarizing the proposed optimization tools is shown in Figure 3.1. For each resonance frequency, satisfying L_r , L_m , and C_r values are calculated. For each calculated combination, a transformer and an inductor are designed based on the analytical models for every possible core in the database. Next, the transformer and the inductor are selected among the designed candidates, which minimizes the defined local fitness function. A key strength of this approach lies in its capacity to apply diverse fitness functions to local and system-level designs. In this approach, multi-core transformer configuration with primary windings in series and secondary windings in parallel is also considered instead of a single-core transformer as in the traditional LLC resonance converter. On the power stage side, the primary and secondary power switch losses for the devices in the database and the required heatsink sizes considering thermal resistances of the packages are calculated. After obtaining the design space, a system level fitness function consisting of power loss, system volume and cost is applied to find the most optimal resonance frequency together with the corresponding locally optimized L_r - L_m - C_r combinations and all other non-trivial components. Different than the previous studies, the algorithm provides;

1. the magnetic design and switch selection has been carried out by considering total energy optimization based on the user charging profile as opposed to the worst-case scenarios or nominal point designs. All losses have been computed for these four operating points and weighted based on the different operating conditions,

contributing to 50%, 20%, 20%, and 10% of the overall consideration, respectively [108, 109],

2. inclusion of thermal resistances of power switches and corresponding heat-sink volumes,
3. determination of optimal effective relative permeability for gapped magnetic cores in the database that is found by sweeping the operating flux density,
4. primary-series secondary-parallel connected multi-core transformer structure,
5. new derivations of current equations in relation to switching frequency,
6. a high level system fitness function consisting of power losses, system volume, and cost [110].

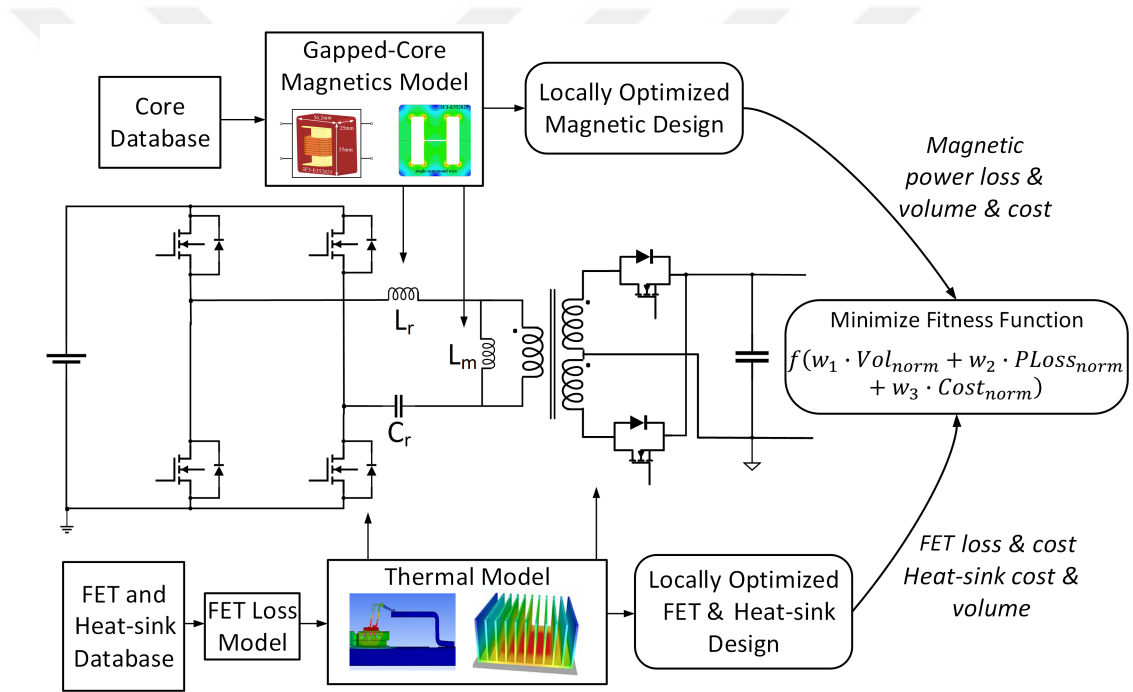


Figure 3.1 : The high-level block diagram of the optimization of LLC resonant converter .

3.1 Proposed Optimization Algorithm

The design of the resonance tank in LLC converters, which is based on the resonance frequency, is crucial for achieving soft switching. A lower L_n ratio leads to higher gain for a given Q and shifts the operating point away from the capacitive region, ensuring ZVS turn-on for the primary switches. However, as L_n decreases, the magnetization current rises, leading to increased circulating current and associated

conduction losses. A smaller Q value results in higher gain, but a higher Q might not fulfill the design requirements and could shift the operation point into the capacitive region. Consequently, optimizing the design of LLC converters is essential, especially for applications involving a wide range of input and output voltages and dynamic load profiles [110].

Designing an LLC resonance converter for a lithium-ion battery charger is more complex than for regular passive load applications due to the non-linear current-voltage (I-V) characteristics of the battery [80]. The charging process of a lithium-ion battery typically includes CC and CV stages, characterized by four critical points as illustrated in Figure 3.2. Initially, in the CC stage, the input current from the high voltage DC bus is minimal, but the switching frequency is at its highest to maintain a low voltage conversion ratio. The transition from CC to CV charging, the turning point, is where maximum power transfer occurs. During the CV stage, the charging current is significantly lower than in the CC stage, making efficiency under light load conditions more crucial. Due to these factors, designs optimized for a specific operating point may not perform well at others. For example, designs tailored for the beginning point might suffer from overheating power switches due to circulating conduction losses or magnetic core overheating from high switching frequencies. Similarly, a design based on the initial point could fail at the turning point due to increased conduction losses in the secondary-side power FETs and transformer. Moreover, a design that is efficient at one operating point may not be effective at others.

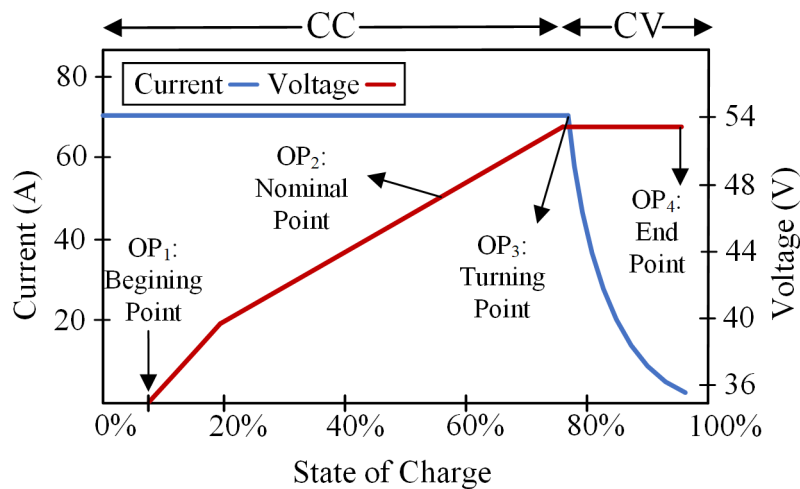


Figure 3.2 : Typical charging characteristics of a Li-ion battery cell and key operation points.

In the presented design framework, all four key operation points are considered in the optimization process. Firstly, for a given resonance frequency, the switching range that covers the voltage conversion ratio of wide input and output voltage ranges is found. The power stage and magnetics are designed and evaluated at other operating points as well. The designs that do not meet the design criteria such as thermal limits and saturation flux density are eliminated. Furthermore, the designs are evaluated at all four operation points with different weights assigned to the fitness function for each operation point as

$$\text{minimize } \sum_{i=1}^4 w_{OP,i} \cdot f(V_{bat_{OP,i}}, I_{bat_{OP,i}}) \quad (3.1)$$

where, $w_{OP,i}$ denotes the weights assigned to each operation point, and $f(V_{bat_{OP,i}}, I_{bat_{OP,i}})$ is the fitness function of each operation point in the local optimization loop. As a result, the proposed algorithm offers optimizing the design for a given operation point as well as across multiple operation points. In this study, the weight distributions have been defined as 50%, 20%, 20%, and 10% respectively. Significantly, our algorithm is inherently adaptable, allowing for the alteration of these weight distributions to accommodate specific user needs or distinct operating conditions.

3.1.1 Determination of switching frequency range

The parametric analyzes are conducted by sweeping the L_n , Q , and f_0 , which inherently determine the switching frequency range. Using the FHA method, L_r , L_m , and C_r values are expressed.

The voltage gain of the converter can be expressed in terms of F , L_n , and Q with the relationship given in equation (2.15). By rearranging equation (2.15) and solving for F , both the minimum and maximum switching frequency that would satisfy the minimum and maximum voltage conversion ratio in can be calculated as given in equation (3.2).

$$\begin{aligned} Q^2 F^6 + \left(\left(1 + \frac{1}{L_n} \right)^2 - 2Q^2 - \frac{1}{M^2} \right) F^4 \\ + \left(\frac{-2}{L_n} \left(1 + \frac{1}{L_n} \right) + Q^2 \right) F^2 + \frac{1}{L_n^2} = 0 \end{aligned} \quad (3.2)$$

Some L_n and Q pairs may require a large switching frequency range to meet the required M of the application. To avoid inefficient results, the maximum frequency is limited to 1.8 times of the resonance frequency.

3.1.2 Transformer and resonance inductor design

The backbone of this systematic approach undoubtedly requires a comprehensive magnetic model that

1. selects the right core material and geometry of the core considering operating flux density, saturation flux and thermal handling of the core from a given database,
2. designs windings with optimized wire thickness considering AC losses,
3. modelling of the magnetic components in terms of power loss as the core and copper losses of the magnetic components contribute to a large portion of LLC converter losses [20].

Unlike conventional transformer design, the magnetizing inductance in LLC converter has to be precisely adjusted for the resonant tank to operate as expected. In order to obtain the required magnetizing inductance for ZVS, an air-gap is inserted into the transformer body on the flux-path [34]. The aim of a gapped-core magnetics design optimization procedure is to determine the core size, core window width, magnetic flux density, current density, number of turns, wire size by means of a previously defined fitness function including power losses, volume, cost, etc. This can be regarded as a multi-objective optimization problem.

The magnetic design for an LLC resonant converter-based lithium-ion battery charger differs notably from designs for regular passive load applications. In such chargers, the process starts with high-frequency switching, gradually reducing the frequency as it approaches the turning point. The resonant current waveform is dependent on the switching frequency, which varies throughout the charging profile. Therefore, it's essential to define the current as a function of both the switching and resonance frequencies. As the frequency lowers, the magnetization current increases. This rise in current can lead to saturation in the magnetic material, necessitating a design that accommodates the lowest frequency and highest transmitted power point. In the design process, power losses are calculated for each critical operation point mentioned.

The final design choice takes into account the importance (or weights) assigned to each of these critical points. In this particular study, the magnetic components are optimized at the turning point, where the charging power reaches its maximum. This approach ensures that the design is most efficient where the power demand is highest, addressing the challenges posed by the varying frequencies and power levels throughout the charging cycle [111].

The magnetic model developed in this study designs locally optimized transformer and inductor according to the given L_m and L_r values, respectively, and provides these locally optimized magnetic designs to the system level optimization tool. Since high power LEV battery chargers generate high output currents, i.e. 68.5 A at 54 V battery voltage for a charger rated at 3700 W, it is a good practice to parallel transformers to decrease the current density on the secondary windings. On the other hand, the current density on the primary windings is not as high due to being fed from high bus voltage. This allows connecting the primary windings of transformers in series to lower the number of turns. Due to the difficulty in controlling the leakage inductance and the possibility of saturation of the magnetic material, a separate resonance inductor is designed instead of integrating it into the transformer core [110].

In order for proper magnetic design, the resonance RMS and peak currents, magnetization peak current, and the secondary-side RMS currents are required. The operating currents are changing with respect to the operation points of LLC namely, ‘below’, ‘at’ and ‘above’ resonance operating conditions. Therefore, the current quantities need to be expressed in relation with the switching frequency. The necessary current expressions of LLC converter that are needed for the magnetic model are extracted in Chapter 2 and maximum value of resonant current and magnetization current are given in equations (3.3) and (3.4), respectively.

$$I_{Lm_{\max}} = \frac{nV_o}{4L_m f_s} \quad (3.3)$$

$$I_{Lr_{\max}} = \sqrt{\left(\frac{\pi I_{out} f_0}{2n f_s}\right)^2 + (I_{Lm_{\max}})^2} \quad (3.4)$$

The transformer and inductor are designed for each L_n , Q and f_0 combination using the core database. This database comprises EE type cores associated with four distinct

materials N87, 3C94, 3C95, and 3F36. Furthermore, these cores present in a spectrum of 16 different volumes, varying from 1490 mm³ (20/10/06) to 102000 mm³ (65/32/27), amounting to a total of 64 different core variants [110].

The flowchart of the proposed optimization algorithm for magnetic components of LLC is given in Figure 3.3.

Step 1: The gapped magnetic design starts off with getting input parameters such as required inductance, allowed temperature rise, ambient temperature, heat transfer coefficient (h_c), window utilization factor and fringing factor (γ). For a given resonance frequency and load conditions, resonance frequency and load conditions, the switching frequency range that will provide ZVS can be determined by FHA. The required current values for magnetic design are calculated. It must be noted that the proposed framework also considers the multi-core structure, where the magnetizing inductance is distributed equally among the cores in an attempt to use smaller cores to reduce the penalty function.

Step 2: A core database is created that stores the information of the core sizes and material information such as the operation switching frequency range, the saturation flux density, and the inductance per turn (A_L) information based on the air gap, etc. For given cores in the database, the first step is to check if the core is suitable for the given switching frequency range that satisfies ZVS and voltage gain across all operating conditions.

Step 3: Unlike classical methods where a pre-determined B_{max} is selected, the operating flux density is swept up to 65% of the B_{sat} in this study [112]. In this way, the optimal operating flux density for each core in the database can be found based on the given penalty function consisting of power loss, cost and volume with low computation effort [113].

Step 4: The current density is calculated for the cores that are provided in the population as given in equation (3.5), where A_p is defined as the product of the core window area and the crosssection area, k_a and k_w are the dimensionless coefficients. According to [95], the typical values of k_a , k_w , k_u are 40, 10, 0.6, respectively, which vary in value for different core material types.

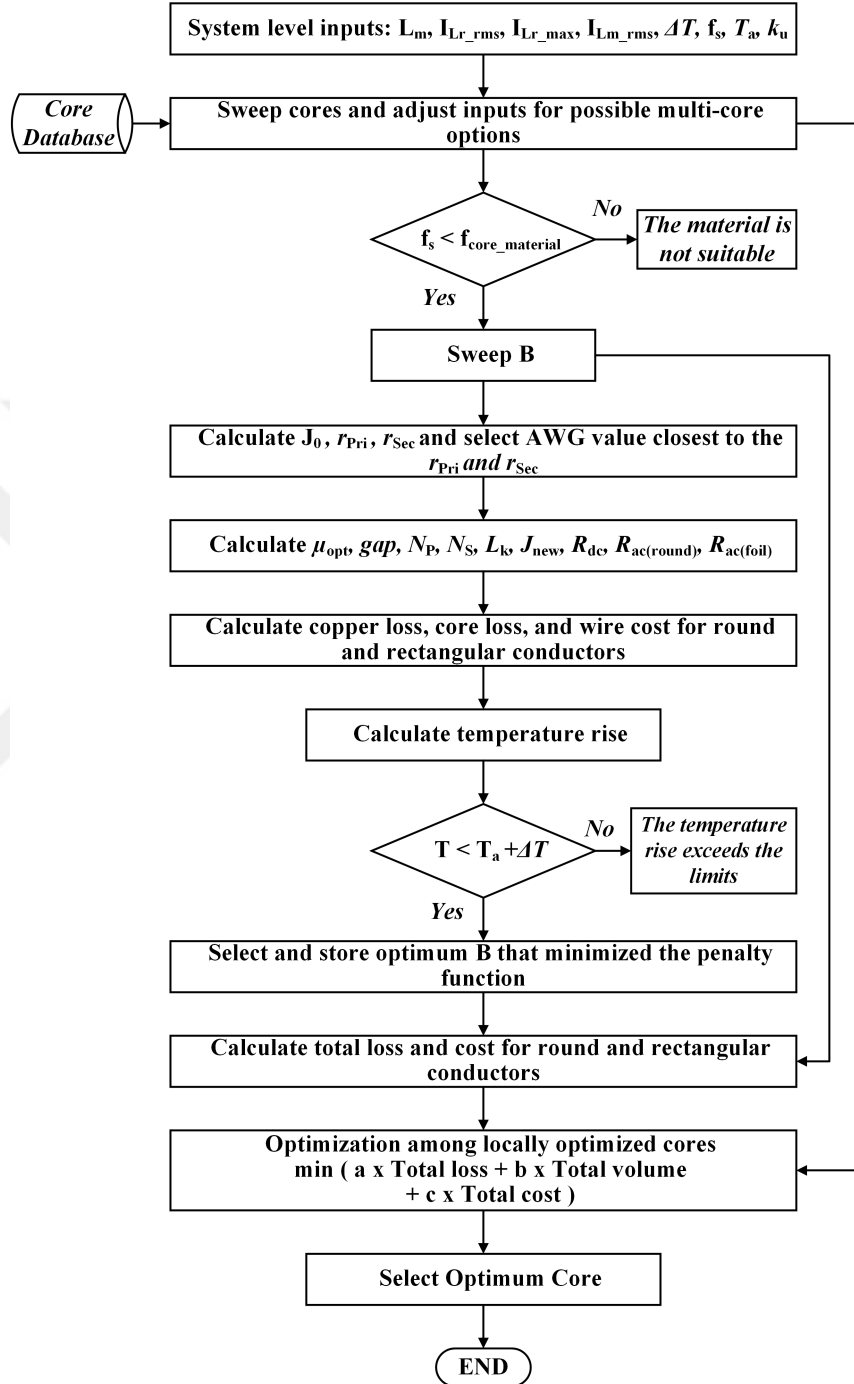


Figure 3.3 : Design flow chart for the gapped magnetics.

$$J_0 = \sqrt{\frac{h_c k_a}{\rho_w k_w}} \sqrt{\frac{\Delta T}{k_u (1 + \gamma)}} \frac{1}{\sqrt[8]{A_p}} \quad (3.5)$$

The cross-sectional areas of the primary and secondary (A_w) windings are calculated in equations (3.6) and (3.7), respectively. The AWG value closest to the calculated conductor dimensions is selected as the conductor diameter (r_{pri} , r_{sec}), and the new current density values for the primary and secondary windings are updated for the cross-sectional areas of the selected conductors with standard AWGs.

$$A_{wp} = \frac{I_{L_{RMS}}}{J_0} \quad (3.6)$$

$$A_{ws} = \frac{I_{sec_{RMS}}}{J_0} \quad (3.7)$$

Step 5: The selection of the operating flux density directly affects the air gap and the number of turns. The optimal relative permeability for each flux density is calculated as in equation (3.8). The optimal relative permeability is derived from the relationship between core loss and dissipation. Afterwards, the air gap value (g) is calculated using the core winding length (l_c) as in equation (3.9). The empirical formula for $A_L - g$ is provided in the datasheet of the manufacturers, and the A_L value corresponding to the calculated air-gap value is determined for each core from this empirical formula. The number of primary turns is calculated as in equation (3.10) depending on the desired magnetization inductance, while the secondary turns are determined using the voltage conversion ratio.

$$\mu_{opt} = \frac{B_{max} l_c}{\mu_0 \sqrt{\frac{P_d k_{up} W_a}{\rho_w M L T}}} \frac{I_{L_{RMS}}}{I_{L_{m_{max}}}} \quad (3.8)$$

$$g = \frac{l_c}{\mu_{opt}} \quad (3.9)$$

$$N_p = \sqrt{\frac{L_m}{A_L}} \quad (3.10)$$

The core window length and width are defined as parameters in the core database in order to check whether the windings fit in the window area. The calculated winding diameters and number of turns that do not fit into the window area are wound in multi-layers with an equal number of turns. The windings are checked one more

time to see if they are fitting inside the window area with this arrangement. By using calculated number of winding and conductor diameters, DC resistance can be calculated as in equation (3.11) where MLT is the mean length turn of the winding.

$$R_{DC} = \frac{N\rho_w MLT}{A_w} \quad (3.11)$$

The skin effect, which causes an increase in the resistivity of the copper as the effective conductive area is reduced at higher frequencies. It contributes to the AC resistance of the round wire as equation given in equation (2.33).

In high current and high frequency LLC converter applications, different winding types (such as foil, round etc...) should be considered particularly at the secondary side. The AC resistance of foil windings, optimal foil thicknesses and lengths are calculated as equations given in equation (2.38).

Step 6: After finding the AC resistances and current values, the copper and core losses are calculated by equations (3.12)-(3.13), where V_c is the volume of core. The Steinmetz equation is used for core loss calculation and the C_m , α and β coefficients were derived by using the P (W/mm³)- B (mT) graphs for different frequency values given by the manufacturer data.

$$P_{cu} = R_{AC} I_{Lr_{RMS}}^2 \quad (3.12)$$

$$P_{core} = C_m f_s^\alpha B_{sweep}^\beta V_c \quad (3.13)$$

Step 7: In order to examine the effect of the losses of the magnetic model on the temperature rise, thermal resistance is calculated for each core by the empirical formula given in equation (2.45) for natural convection cooling, and temperature rise is checked to see if the core temperature rise stays within the given limits.

Step 8: A penalty function consisting of volume, cost and power losses are applied to each designed transformer and inductor. In order to include the cost in the penalty function, the winding Cu cost, material cost based on volume, and labor costs are extracted from different suppliers. Total core, winding and labor costs can be calculated as in equations (3.14)-(3.15), where $\sigma_{core,x}$, $\sigma_{wdg,x}$ and $\sigma_{lab,x}$ are the specific costs per weight considering different magnetic core and the winding types. $\Sigma_{core,x}^{fc}$, $\Sigma_{wdg,x}^{fc}$

and Σ_{lab}^{fc} indicate the fixed costs for the magnetic cores including the coil formers, connectors, and labor, where N_{stack} represents the stacking factor. All numerical data retrieved from manufacturer is given in Table 3.1 [114].

$$\Sigma_{core} = N_{stack} \Sigma_{core,x}^{fc} + \sigma_{core,x} W_{core} \quad (3.14)$$

$$\Sigma_{wdg} = \Sigma_{wdg,x}^{fc} + \sigma_{wdg,x} W_{wdg} \quad (3.15)$$

$$\Sigma_{lab} = \Sigma_{lab}^{fc} + \sigma_{lab,x} W_{wdg} \quad (3.16)$$

Step 9: After obtaining the penalty function, the processes are repeated for each core and flux density, where the cost function is applied at each sweep. The most optimal core and design of the magnetics minimizing the cost function is selected at the end of the proposed framework [113].

This study provides a systematic design method for air-gapped magnetics in an LLC converter based on a given core database by optimizing cost, volume and power losses. The presented method can be used as a magnetic optimization tool in a higher level system optimization framework for given L_r and L_m combinations that satisfy the required voltage conversion gain across whole load range. Different than the previous studies, the algorithm provides;

1. an optimization according to the assigned weights of cost, weight, and volume. The cost function includes core, winding and labour costs. The power loss function includes core loss, temperature dependent copper loss, and AC winding loss due to skin effect.

Table 3.1 : Retrieved numerical cost model datas from several winding, core and inductor manufacturers.

Specific Cost	Value
$\Sigma_{core,x}^{fc}$ (€/unit)	0.08
$\sigma_{core,x}$ (€/kg)	7.5
$\Sigma_{wdg,x}^{fc}$ (€/unit)	0.25
$\sigma_{wdg,x}$ (€/kg)	16.5
Σ_{lab}^{fc} (€/unit)	0.75
$\sigma_{lab,x}$ (€/kg)	7

2. Unlike conventional design method where optimal permeability is assigned, it is found by sweeping the operating flux density for the cores in the given database that includes various cores with different material each one resulting in different loss coefficients for frequency and flux density.
3. Both foil and Litz wires with interleaved structure have been considered.
4. Distributed multiple cores at both primary and secondary sides have been considered.
5. This systematic design can be easily integrated with the Metaheuristic Algorithm. By doing so, the optimal flux density value can be identified without sweeping through all flux density values, thereby reducing the computation time [111].

3.1.3 Power switch and heat-sink selections

The first criteria in choosing a power device for the primary and secondary sides is to satisfy the ZVS operation. For the given design variables of L_m and f_0 , the energy stored in the magnetizing inductance must be sufficient to soft-switch the primary side FETs. Besides, in the dead-time period, the energy from the magnetic field, associated with the I_{Lm} , needs to charge and discharge the equivalent switch node capacitance (C_{eq}). The minimum energy and dead-time required to ensure ZVS operation is are given in equations (3.17) and (3.18).

$$\frac{1}{2} (L_m + L_r) (I_{Lm_{\max}})^2 \geq \frac{1}{2} (C_{eq}) \cdot V_{dc}^2 \quad (3.17)$$

$$t_{dead} \geq 4 (C_{eq}) f_s L_m \quad (3.18)$$

where, C_{eq} is the equivalent capacitance referred to the switch node, which includes the total parasitic output capacitance of the primary side power devices, the total parasitic output capacitance of the secondary side power devices reflected to the primary side, the parasitic capacitance of the switch node traces as well as the transformer and inductor winding capacitances. For simplicity, C_{eq} has been taken equal to $4 \times C_{oss_{pri}}$ in literature [33]. Since the ZVS operation of the primary power switches are guaranteed by choosing appropriate power devices in accordance with (3.17), there are no power losses during turn-on. However, there are still conduction and turn-off losses occurring

in the primary power switches. In high-frequency circuit designs, dead-time losses are another type of loss that can be important. These losses depend on the switching period. They can either be added as a constant amount of time or as a percentage of the time when soft-switching occurs. All relevant losses can be calculated with the given equations (2.46)-(2.52).

Please note that I_{out} , and thereby, $I_{sec_{RMS}}$ decrease proportionally with parallel number of transformers and power devices per each secondary side output stage. The secondary side synchronous devices are ZCS turned on and off at below resonance operation. In the above resonance operation, the secondary side power devices are turned on under current, where di/dt is limited by the loop and leakage inductances.

To determine the necessary heat-sink size for each combination of L_n , Q , and f_0 , and thus select the best power device and package, estimating the thermal resistance from the power device's junction to the heat-sink is crucial. A comprehensive MOSFET/SiC diode database from various manufacturers, including package information, has been assembled for this purpose. This database includes 18 different switches for the primary side and 16 for the secondary side. While FEA tools could be used for total thermal resistance calculations, this requires intensive computation. To simplify, the theory of iso-flux rectangles and strips on compound flux channels, as detailed in [102], has been applied to each layer. The thickness of the DAP varies depending on the die and package construction. The entire thermal pad area is utilized for cooling and is filled with vias 8 mils in diameter, acknowledging that the power loop is less critical in soft-switching converters compared to hard-switching ones. It is assumed that the top and bottom copper layers are 70 μm thick, the total PCB thickness is 1.6 mm, the vias are plated with 25 μm copper, and filled with conducting epoxy. For each power device in the database, the required heat-sink is calculated based on the product of power losses and thermal resistances. A correlation is established between the heat-sink surface area and its thermal resistance. Then, the heat-sink from the database, which includes 72 different heat-sinks of various sizes, is chosen based on the closest surface area to this value.

For each power device in the database, the required heat-sink is calculated based on the product of power losses and thermal resistances. A correlation between the heat-sink

surface area and corresponding thermal resistance is obtained. Then, the heat-sink in the database that has closest surface area to this value is selected from the database.

The costs of heat-sinks generally depend on the base material, volume and weight as well as the manufacturing and engineering costs. Heat-sink costs can be calculated in equation (3.19), where $\sigma_{sink,x}$ is the specific costs per volume depending on heat-sink type and $\Sigma_{sink,x}^{fc}$ is fixed costs which can result from additional engineering work and processing steps which are independent from heat-sink volume [114].

$$\Sigma_{sink} = \Sigma_{sink}^{fc} + \sigma_{sink,x} W_{sink} \quad (3.19)$$

3.1.4 Optimization flowchart

After finding the volume, cost, power losses of the local best primary and secondary power devices as well as the transformer and inductor, the values are stored for the corresponding L_n and Q combination. In this way, a design space is created with locally optimized designs. The final selection of the design parameters is achieved by assigning weights to the volume, size and cost of the locally optimized designs and finding a global solution that minimizes the fitness function. The flowchart of the proposed framework is shown in Figure 3.4.

3.2 Optimization Algorithm and Experimental Results

3.2.1 Results of gapped and distrubeted core magnetics optimization algorithm

In this study, the goal is to optimize the magnetic components of a 400 V/48 V, 3700 W LLC design for a LEV charger, which operates at a resonance frequency of 370 kHz. The specified values for the magnetization inductance, resonant inductance, and resonant capacitor are 37.52 μ H, 9.38 μ H, and 19.73 nF, respectively. These parameters are input into the proposed magnetics optimization model, which is part of a broader design space evaluated by a higher-level converter optimization tool. The evaluation of the magnetic components occurs at a switching frequency of 299.8 kHz, as determined by the sweeping action of the higher-level converter optimization loop. The design of the magnetic components is based on the provided L_m and L_r values. It is important to note that the higher-level converter optimization algorithm for the LLC converter is

not covered in this study case. Additional details and specifications of the design are presented in Table 3.2.

Table 3.2 : Design specification of LLC.

Symbol	Description	Value
P_o	Output power	3700 W
f_s	Switching frequency	300 kHz
f_0	Resonant frequency	370 kHz
L_m	Magnetizing inductance	37.52 μ H
L_r	Resonant inductance	9.38 μ H
C_r	Resonant capacitance	19.73 nF
ΔT	Core maximum temperature rise	90 °C
I_{LrRMS}	Resonant RMS current	10.55 A
I_{Lmmax}	Magnetizing peak current	9.6 A
I_{secRMS}	Secondary side RMS current	84.55 A

The gapped transformer design steps are applied to a number of cores in the database. The presented framework eliminates the cores that do not provide either the necessary magnetization inductance, thermal limit, B_{sat} limit or frequency limits. This essentially eliminates too-small cores. In addition, too-large cores are also eliminated thanks to the fitness function. Afterwards, among the satisfying transformers, the design that provides the minimum value for the penalty function with assigned weights of 60%, 20%, and 20% to volume, power loss, and cost, respectively, is selected. The chosen design involves two parallel 3F36-E422120 cores with a combined volume of 45400 mm³ and a 1.272 mm air gap. Their primary windings are connected in series, and secondary windings in parallel. The total core loss for this design is calculated to be 9.74 W. Depending on the type of winding used on the secondary side, total copper loss is calculated at 2.84 W for round wire and 2.7 W for copper foil.

For comparison, the optimization algorithm was also run under the constraint of using only single cores. The best single core identified was the 3F36-E552825 with 8 turns and a 1.027 mm air gap for the required inductance. The core loss for this single-core design is 15.72 W, with copper losses at 1.674 W and 1.59 W for round wire and copper foil, respectively. The results show that the algorithm opts for two 3F36-E422120 cores in a series-parallel configuration because the total volume and loss are lower than that of the single 3F36-E552825 core, particularly when the assigned weight to volume is

60%. This demonstrates the effectiveness of the optimization algorithm in selecting the most efficient design based on the given criteria.

To validate the optimization framework's model, the designed transformers listed in Table 3.3, along with the LLC circuit model specified in Table 3.2, are co-simulated using Ansys Electronics Desktop and Simplorer. For FEA and Simplorer co-simulations, two transformers are selected: the 3F36-E552825 with a single parallel configuration and the 3F36-E422120 with a double parallel configuration. The 3D models of these transformers are depicted in Figure 3.5. A two-dimensional (2D) transient magnetic model is employed in the FEA to compare results and assess operational parameters like magnetic flux density, current density, and losses. These simulations are carried out in Ansys Electronics Desktop, utilizing the actual B-H curve of the magnetic material, which accounts for high-frequency saturation effects. One important consideration is the fringing flux in the transformer's air gap, which can induce eddy currents in nearby conductors. This fringing flux is dependent on the air-gap geometry and can only be accurately evaluated using FEA. Additionally, the electric field generated between the primary and secondary windings can lead to common mode noise through the transformer. Literature suggests that power loss due to the proximity effect and primary to secondary winding capacitance can be mitigated by interleaving the windings [112]. Interleaving not only helps in reducing power loss but also minimizes leakage inductance. Consequently, the selected transformers are designed and analyzed with interleaved winding configurations using FEM. The resulting magnetic flux distributions are shown in Figure 3.6 for each transformer, using both round wires and copper foils. The figures indicate significant high-frequency effects, such as fringing near the air-gap and proximity effects between intra-windings and inter-windings. As observed in Figure 3.6, the flux is denser near the core section close to the winding window area, where the flux path length is minimal, and it becomes more concentrated on the inner edge of the core, as expected.

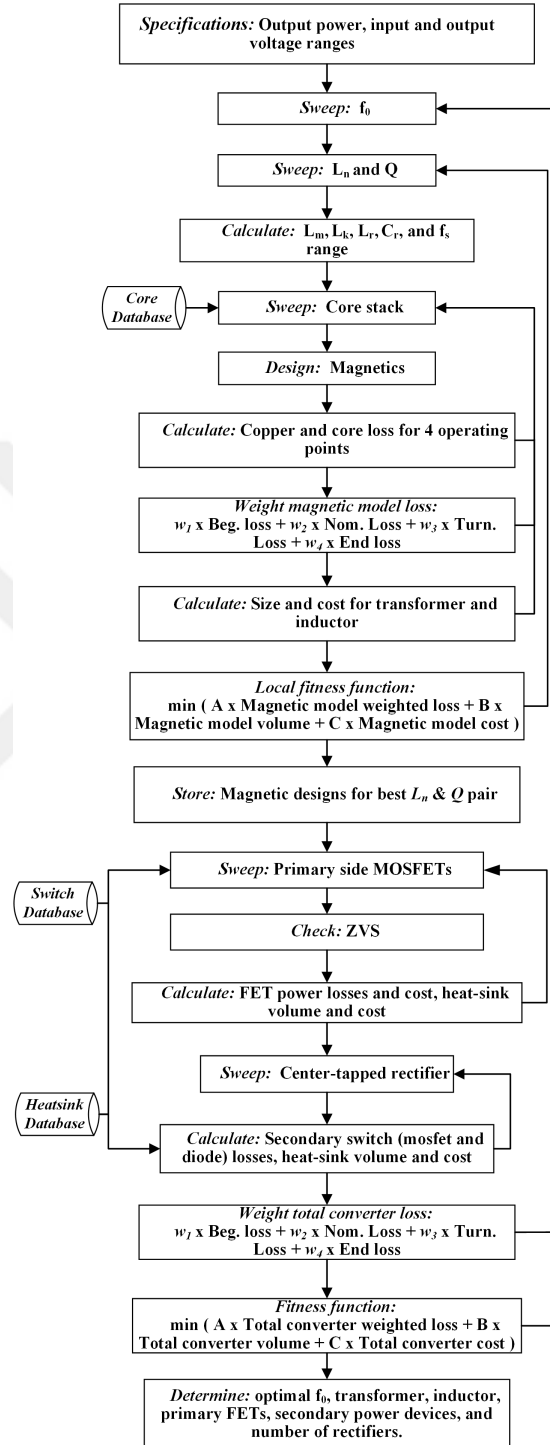


Figure 3.4 : Detailed flow chart of the system-level design procedure for LLC converter.

Table 3.3 : Results provided by design and optimization algorithm.

Selected Transformer	Nu. of Series /Parallel Config.	Sec. Winding		Total Volume (mm ³)	Cost (\$)	ΔB_{max} (mT)	Core Loss (W)	Copper Loss (W)	Temp. Rise (ΔT)	Air-gap (mm)		
		Sec. Turn	Winding Type								Dimensions Width/Thickness (mm) or AWG	
3F36-E552825	1	8	1	Round	AWG4	52000	3.01	107.21	15.72	1.674	48.23	1.027
				Foil	23.73/0.4		2.94			1.59	48	
3F36-E422120	2	8	2	Round	AWG7	45400	3.15	96.63	9.74	2.84	26.42	1.272
				Foil	25/0.2		3.08			2.7	26.1	

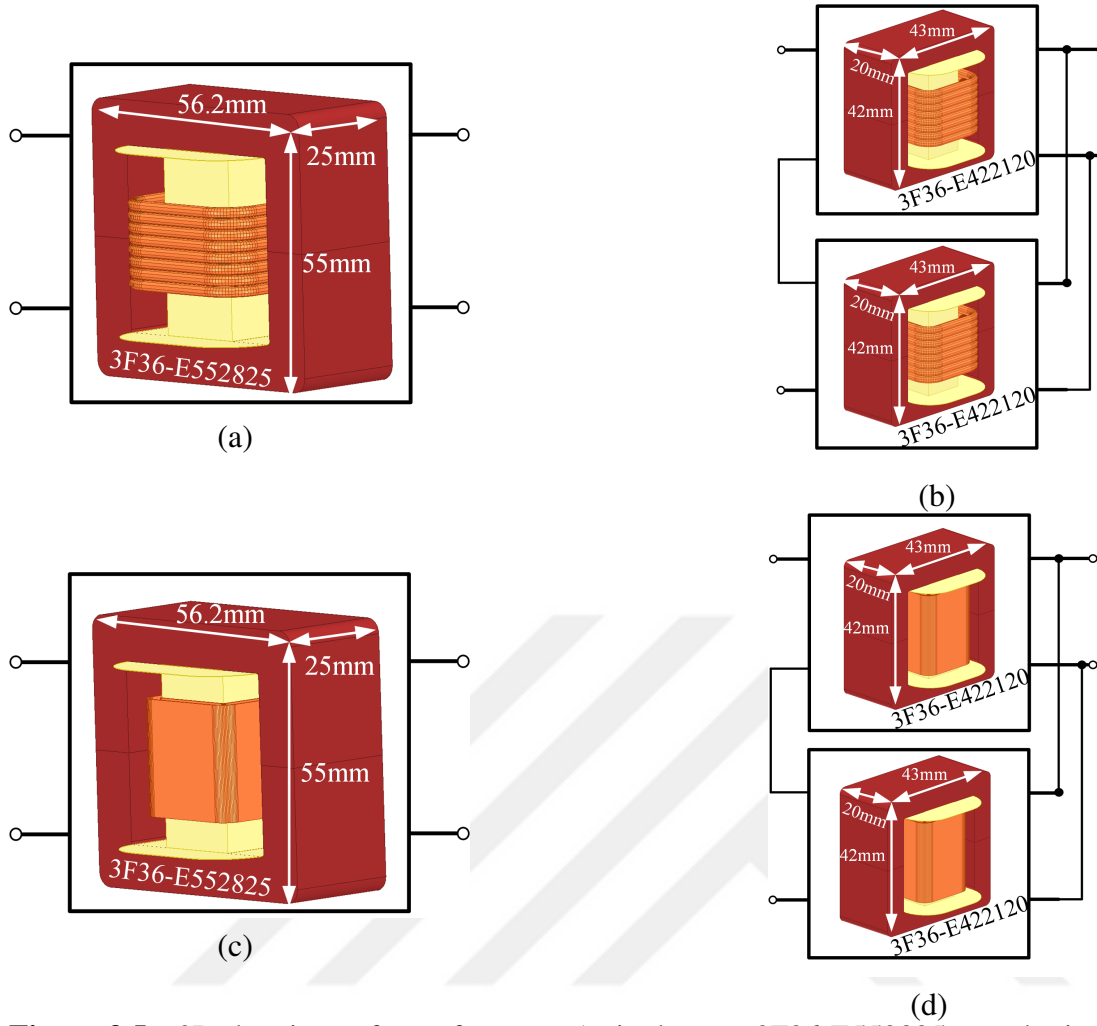


Figure 3.5 : 3D drawings of transformers, a) single-core 3F36-E552825 round wire, b) two-core 3F36-E422120 round wire, c) single-core 3F36-E552825 Cu foil, d) two-core 3F36-E422120 Cu foil.

Figure 3.7 shows the magnetic flux lines and the current density distributions in the round and foil windings. As seen from the figure, the current gets crowded on the conductors near the air gap, and also on the surfaces, where primary and secondary windings are facing each other. Obviously, the eddy currents crowding these surfaces contribute to copper losses. FEA tools are used to calculate the copper losses due to the fringing effect and proximity effect.

The resultant magnetizing inductance in Ansys Electronic Desktop and Simplorer co-simulation is found as $35.644 \mu\text{m}$, which matches closely with the target specification of $37.52 \mu\text{m}$. The analytical and FEA calculation of copper and core losses are given in Figure 3.8. According to Figure 3.8, there is a slight difference between the FEA

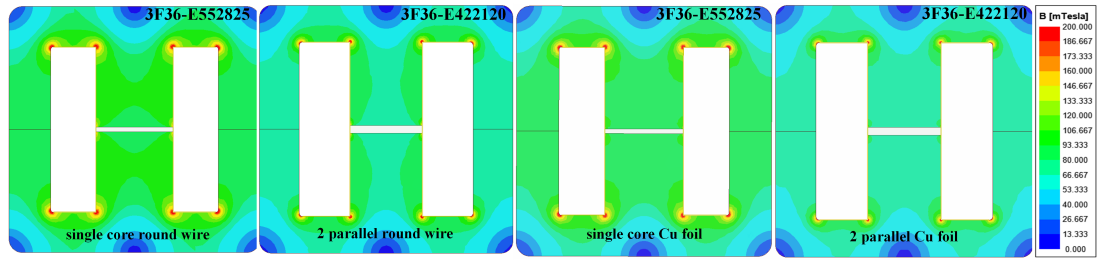


Figure 3.6 : Magnetic flux density distribution of each transformers.

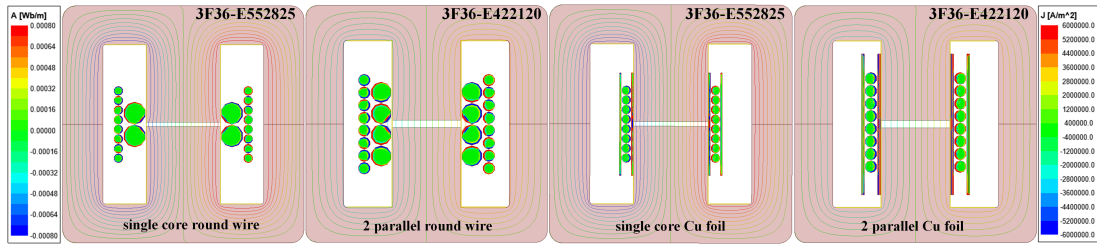


Figure 3.7 : Magnetic flux lines and current density distribution of each transformers.

measurements and the values calculated from the optimization algorithm. The main reasons of this error can be summarized as follows;

1. the algorithm does not consider a dynamic BH curve,
2. the fringing effects are more accurately modelled with the FEA tool,
3. there are frequency dependent uncertainties in core loss coefficients in the FEA tool.

This error can be further reduced by using higher number of meshes and different mesh types; however, this would require an extensive computation time and processing power.

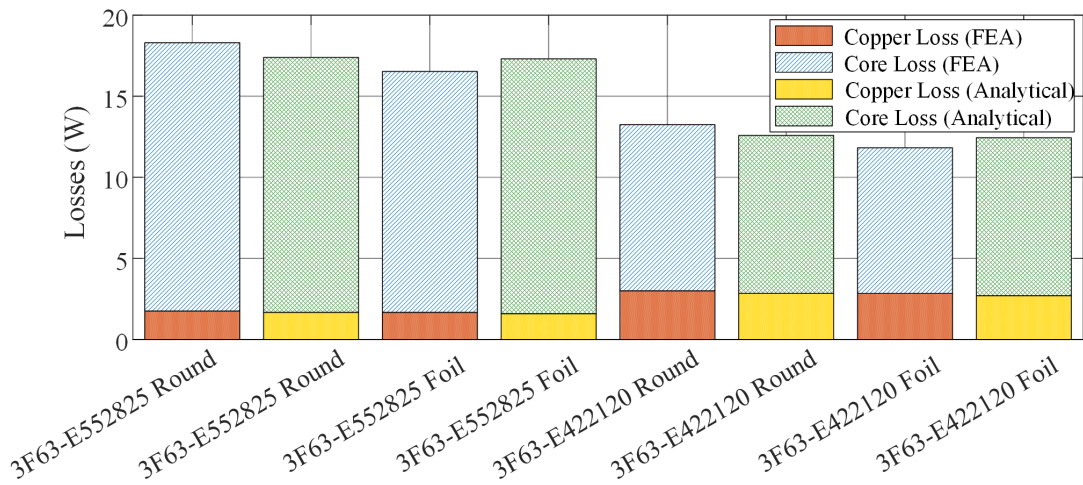


Figure 3.8 : Copper and core losses of selected transformers for round wire and Cu foil windings.

3.2.2 Experimental results of optimum gapped core magnetics

A gapped transformer based on the design results has been built to verify the presented methodology. The transformer is placed in the 3700 W LLC resonant converter. The prototype of LLC resonant converter is shown in Figure 3.9.

According to the values obtained from the design algorithm, the L_m , L_k and L_r values should be $37.52 \mu\text{H}$, $1.3 \mu\text{H}$, and $7.78 \mu\text{H}$, respectively. Here, the loop inductance of the path was calculated approximately as $0.5 \mu\text{H}$. However, there is an error margin of 5-10% in the designs of the manufacturers and the L_m , L_k and L_r values have been measured as $35.245 \mu\text{H}$, $1.4 \mu\text{H}$ and $8.2 \mu\text{H}$, respectively. The resonance frequency has been experimentally obtained approximately around 350 kHz.

It is not easy to separate the magnetic losses from the overall converter losses during operation. To estimate the transformer power losses, the thermal resistances of the transformer windings have been extracted by passing currents through the primary and secondary windings from 0 A to 20 A at 1 A resolution. Then, the power loss and corresponding temperature data are recorded experimentally. In this way, the transformer power losses are estimated through observing the temperature rise. The curve fitted data for the thermal resistances of primary and secondary windings are given in Figure 3.10.

In order to verify the power loss results generated by the proposed algorithm, the calculated core and winding losses are multiplied with the thermal resistances of the core, primary and secondary windings, respectively, as given in (3.20). Here, T_c is

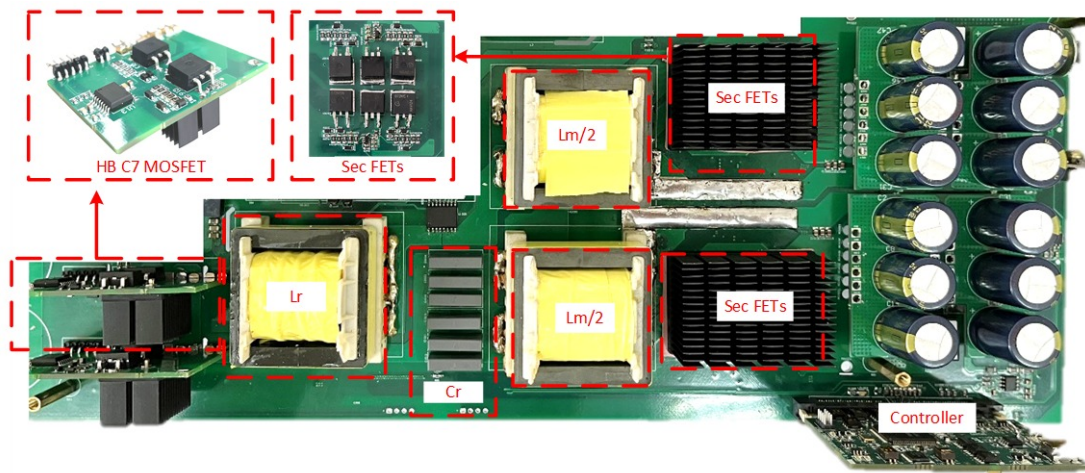


Figure 3.9 : The designed LLC prototype rated at 3.7 kW, 400 V/48 V.

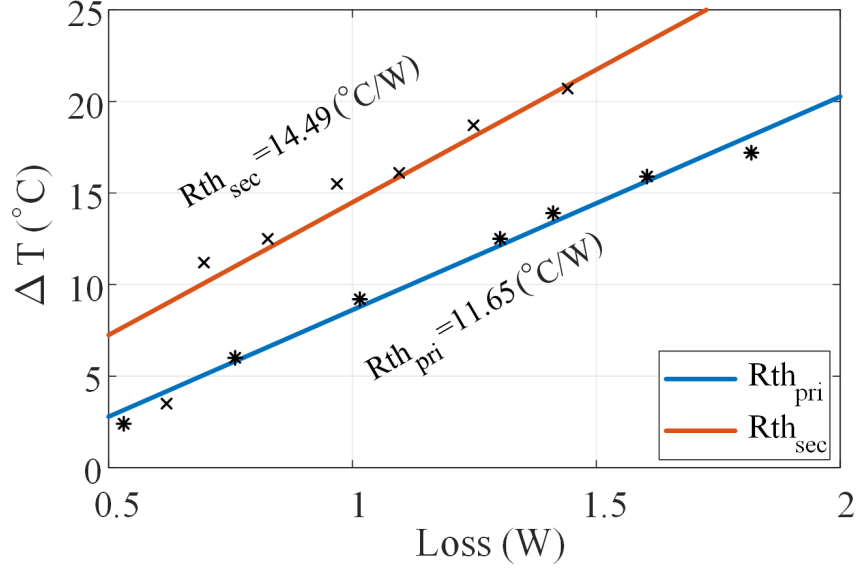


Figure 3.10 : The power loss vs temperature increments of primary and secondary windings.

the transformer temperature, T_{amb} is the ambient temperature, $P_{pri_{loss}}$ is the primary winding copper loss, $P_{sec_{loss}}$ is the secondary winding copper loss, $P_{core_{loss}}$ is the core loss, $R_{t_{pri}}$, $R_{t_{sec}}$ and $R_{t_{core}}$ are the thermal resistances of the primary winding, secondary winding and core, respectively.

$$T_c = T_{amb} + (P_{pri_{loss}} R_{t_{pri}} + P_{sec_{loss}} R_{t_{sec}} + P_{core_{loss}} R_{t_{core}}) \quad (3.20)$$

The parameters obtained from the algorithm at different frequencies and load conditions are given in Table 3.4. At 340 kHz and 2.19 kW operation, the temperature rise of the primary winding, secondary windings and core are estimated as 6.23 °C, 9.74 °C and 17.6 °C, respectively, when the ambient temperature is 25 °C. Consequently, the temperature of the transformer is estimated to be 58.5 °C, which is quite close to the experimentally obtained temperature value of 58.7 °C as shown in the Figure 3.11(a). Similarly, the temperature of the transformer is estimated to be 51.1 °C for 1.8 kW operation at 280 kHz as shown in Table 3.4. The experimentally recorded temperature for this case is shown in Figure 3.11(b). The results suggest that the output of the proposed algorithm are consistent with the experimental readings, which verifies the power loss equations and analysis incorporated in the proposed magnetic design algorithm.

Table 3.4 : Thermal results of the proposed magnetic design algorithm and the experiments.

V_{in} (V)	f_0 (kHz)	I_o (A)	V_o (V)	$P_{pricu_{loss}}$ (W)	$P_{seccu_{loss}}$ (W)	P_{core} (W)	Load (Ω)	T_c ($^{\circ}\text{C}$)	T_c ($^{\circ}\text{C}$) (Expr.)
390	340	40.51	54.06	0.5351	0.6724	4.5	1.4	58.5	58.7
390	280	30.72	55.72	0.4473	0.4695	3.66	1.7	51.1	50.1

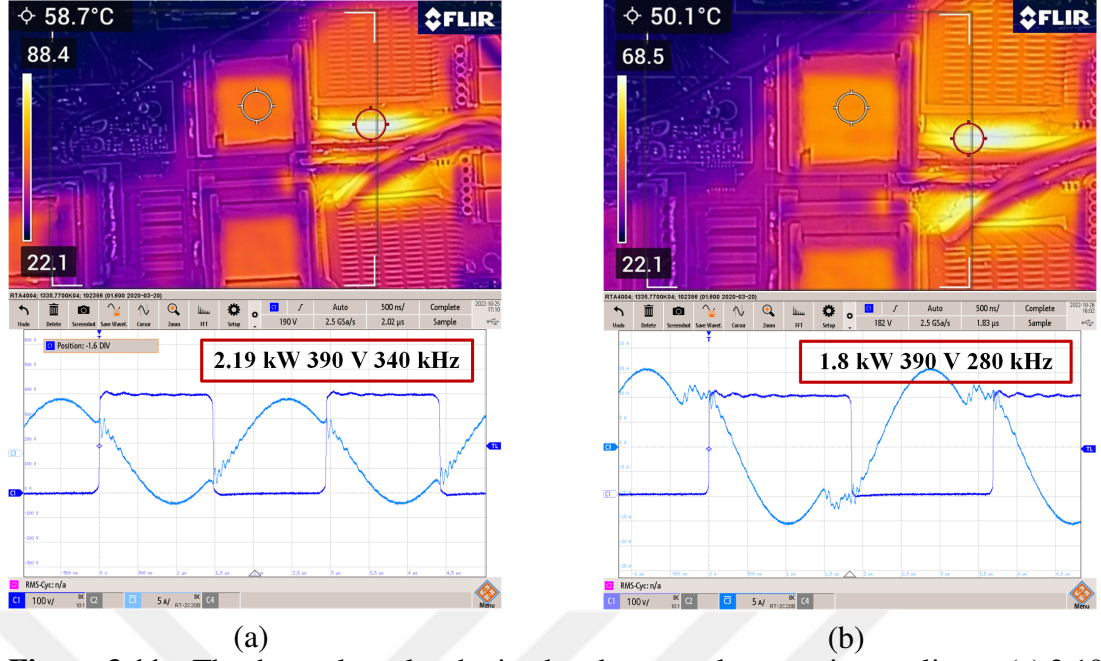


Figure 3.11 : The thermal results obtained under natural convection cooling at (a) 2.19 kW 390 V 340 kHz, (b) 1.8 kW 390 V 280 kHz.

In order to show the advantage of the proposed algorithm, the proposed optimization algorithm has been run with the design specifications of some of the reference designs and LLC prototypes published in journals [20, 115–118] and magnetics volume results of the optimization algorithm have been compared in Figure 3.12. As it can be seen from the comparison results, magnetics can be designed in smaller volume by using the proposed algorithm.

This study proposed a comprehensive magnetics design approach for LLC resonant converter transformers and inductors. For given L_r and L_m pairs that satisfy the required voltage conversion gain over the entire load range, the ratio between the copper losses and core losses of magnetic components is optimized by minimizing the fitness function consisting of cost, power losses and volume under the thermal and magnetic restrictions such as saturation flux density, current density on the wires, etc. The algorithm then chooses the optimal core, air gap, and the winding design. Unlike conventional approaches, the proposed methodology sweeps the magnetic flux density through a metaheuristic optimization method named as Big-bang Big-Crunch, and considers distributed core structure and different winding types such as Litz wire, round wire, and Cu foils.

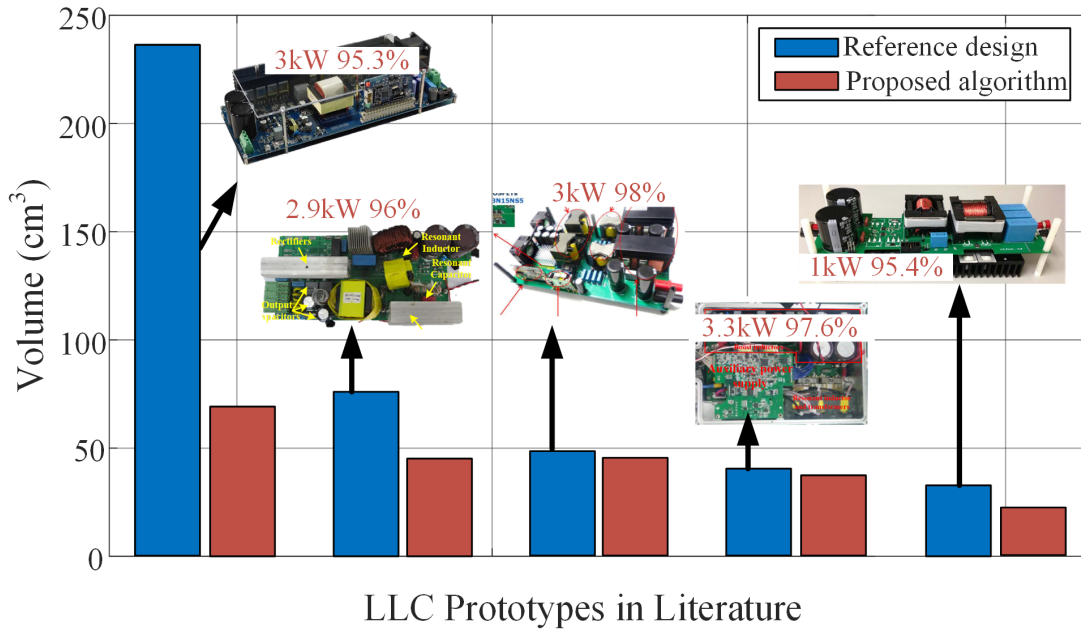


Figure 3.12 : Comparison of magnetics volumes in published LLC prototypes and the results of the proposed algorithm.

The proposed algorithm is used to design the transformer of a 3700 W 48 V LLC resonant charger in a LEV for a given magnetizing inductance of $37.52 \mu\text{H}$. The penalty function has weights of 60%, 20%, and 20% assigned to volume, cost and power losses, respectively. The algorithm designed a transformer with distributed cores, in which the primary and secondary windings are connected in series and parallel, respectively. The results show that when the weight of the volume in the fitness function is higher, designs with smaller cores are chosen over single-core designs, as the thermal stress is smaller per core in comparison to a single core design. For this specific design, the algorithm chose two 3F36-E422120 cores and wound 8 turns with round wire on the primary side and 2 turns with 0.2 mm Cu foils on the secondary. Interleaved winding structure reduced the copper and core losses by up to 15% and 5%, respectively, as the proximity and fringing effect diminishes with interleaved winding structure. To verify the analytical equations, a 3-D model has been constructed and co-simulated by Ansys Electronic Desktop and Simplorer. The results closely match with the analytical model. Furthermore, a 3700 W 48 V LLC prototype has been built with the magnetics suggested by the developed algorithm, and the estimated power losses have been verified experimentally through monitoring the temperature rise of the transformer.

The algorithm has been forced to design a transformer with a single core. The model has designed 3700 W, 400 V/48 V 370 kHz resonant frequency. The designed transformer

with a single core resulted in 39% higher power losses and 14.53% more volume in comparison to the optimized distributed core design. All in all, the proposed gapped core optimization method can be used as a part of a higher system level LLC converter optimization, where each viable L_r and L_m combination can be evaluated in order to find the most optimum LLC design.

3.2.3 Results of system level optimization algorithm

A 3700 W LLC converter with an input-output voltage range of 430V-370V and 54V-48V is designed using the proposed optimization framework under a 97% efficiency constraint. The algorithm sweeps the resonance frequency between 250-500 kHz at the outer loop, while the minimum and maximum frequencies are internally swept through L_n and Q sweeps. Both the weights of the local and system level fitness functions are chosen as 60%, 20%, and 20% for volume, power loss, and cost, respectively.

Figure 3.13(a) shows the design space considering magnetic designs achieved using every feasible core in the database for each resonance frequency and L_n - Q combinations. As it can be seen in Figure 3.13(a), increase in f_0 decreases the total volume and cost of magnetic component; however, the power losses also increase. By applying a local fitness function to the magnetic model, the design space is further reduced.

The resultant design spaces in the case of using SR and SiC diodes on the secondary side are shown in Figure 3.13(b) and Figure 3.13(c), respectively. These points show the total volume, cost and efficiency of the converter for the local best designs at each swept resonance frequency. For high power design specifications, use of SR on the secondary side brings an advantage in efficiency and heat-sink volume, even though it increases the cost. The details of the design space with SR is provided in Figure 3.14.

For detailed in-depth analysis, the volume, loss and cost for the primary and secondary FETs, the transformer and the inductor are provided with the radar chart given in Figure 3.15(a), 3.15(b), 3.15(c). In order to observe the distribution of primary and secondary MOSFET power losses in more detail, the conduction and switching losses are shown in Figure 3.15(d), 3.15(e). The results of the system level optimization tool are summarized with the radar chart given in Figure 3.15(f).

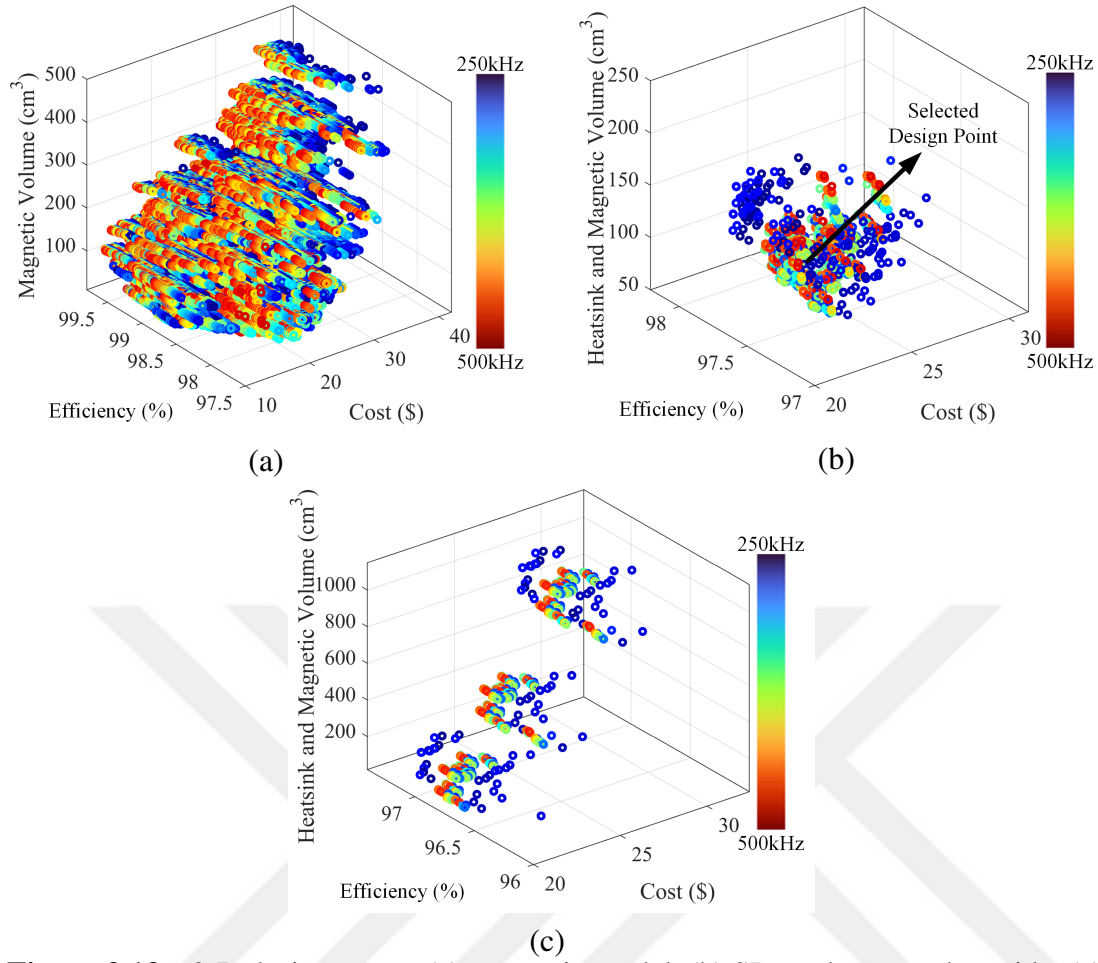
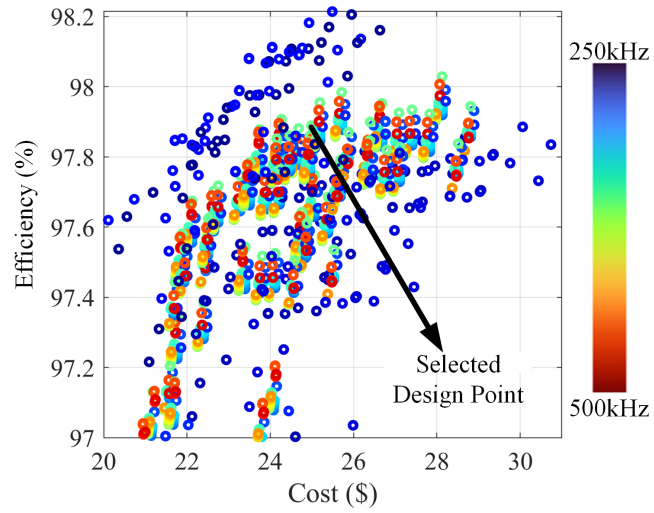
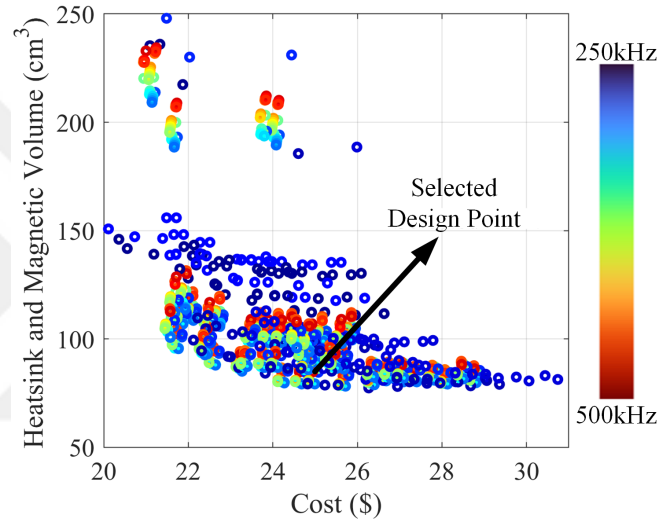


Figure 3.13 : 3-D design space; (a) magnetic model, (b) SR on the secondary side, (c) use of diodes on the secondary side.

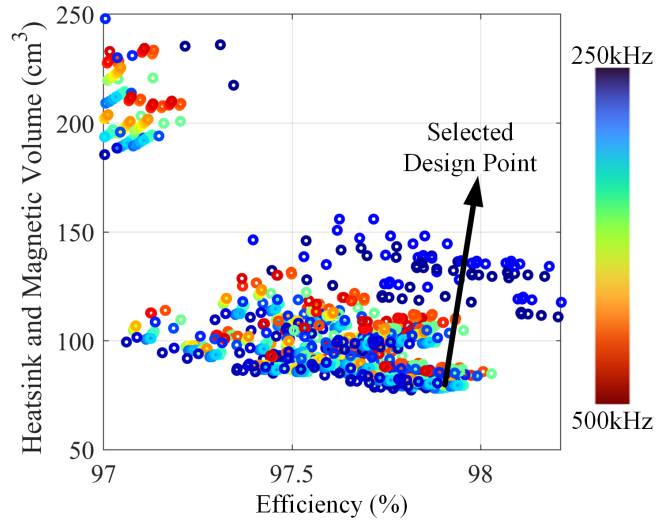
By applying the same fitness function applied locally to the whole design space, the optimum design parameters are found as $37.52 \mu\text{H}$, $1.3 \mu\text{H}$, $9.38 \mu\text{H}$ and 19.73 nF for L_m , L_k , L_r and C_r , respectively. The algorithm suggests using $40 \text{ m}\Omega$ / 650V and 7.2Ω / 150 V MOSFETs on the primary and secondary side, respectively. The optimal f_0 , f_s range and efficiency values for this optimized system are found as 370 kHz - 610 kHz and 97.83% , respectively. For higher efficiency and smaller heat-sink size, two MOSFETs are paralleled; therefore, there are a total of eight devices on the secondary side. The core material is 3F36. The design parameters of the transformer and inductor are provided in Table 3.5.



(a)

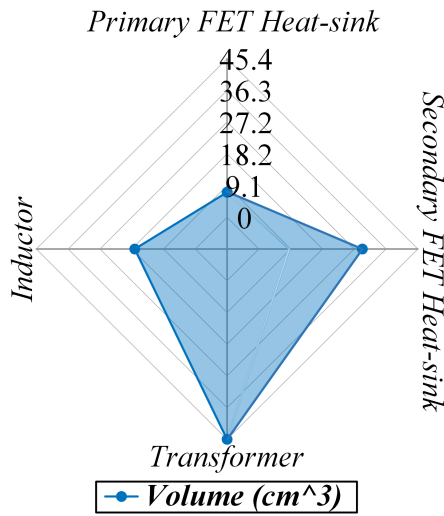


(b)

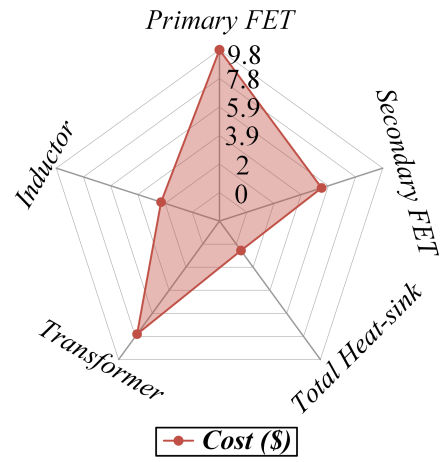


(c)

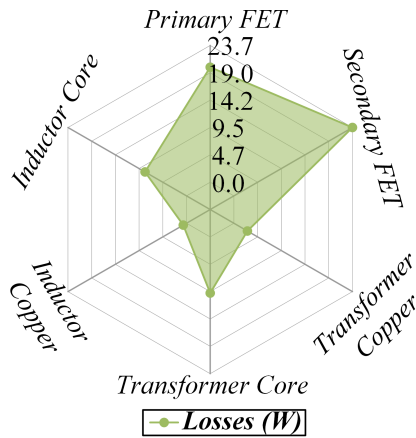
Figure 3.14 : 2-D design space with SR on the secondary side; (a) efficiency-cost, (b) volume-cost, (c) volume-efficiency.



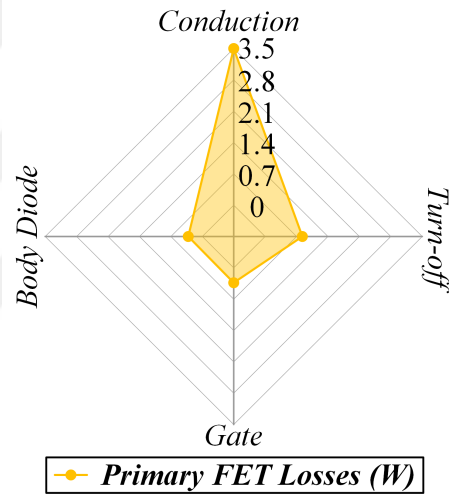
(a)



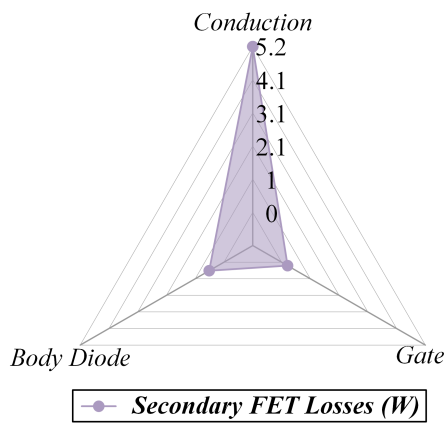
(b)



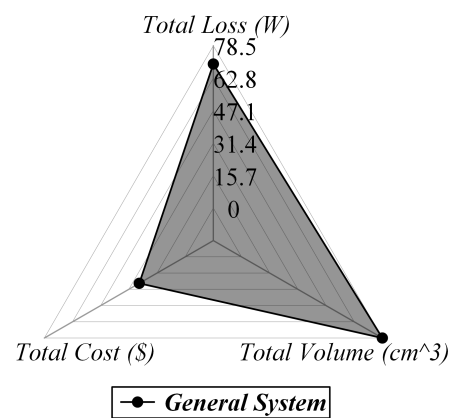
(c)



(d)



(e)



(f)

Figure 3.15 : Radar chart for the optimal design point; (a) total volume, (b) total cost, (c) total power losses, (d) power losses on primary FETs, (e) power losses on secondary FETs, (f) system level results.

Table 3.5 : The optimum transformer and inductor design parameters for $L_m=37.52 \mu\text{H}$ and $L_r=9.38 \mu\text{H}$.

Magnetics	Transformer Core	Number of Cores	Primary Turn	Secondary Turn	ΔB_{max} (mT)	Temperature		Air-gap (mm)	Primary J (A/(mm ²))	Secondary J (A/(mm ²))
						Rise (ΔT)	Rise (ΔT)			
Transformer	3F36-E422120	2	8	2	96.63	26.1	26.1	1.272	2.95	2.83
Inductor	3F36-E422115	1		10	101.06	43	43	3.942	2.95	

3.2.4 Experimental results of optimum design

A 3700 W/48 V LLC prototype has been designed to verify the theoretical calculations used in the optimization tool. The photo of the designed converter is shown in Figure 3.16. Using the proposed optimization algorithm, the L_m , L_k , and L_r values were found as 37.52 μH , 1.3 μH , and 7.78 μH , respectively. However, there is typically an error margin of 5-10% in manufacturing of wound magnetics. The designed L_m , L_k , and L_r values have been measured as 35.245 μH , 1.4 μH and 8.2 μH , respectively. Here, the loop inductance on the PCB enclosing the impedance network from the half-bridge is calculated approximately as 0.5 μH . Therefore, the resonance frequency has been experimentally obtained around 350 kHz.

Figure 3.17 shows the operation of LLC converter under different loads up to 4300 W. Here, V_{DS} is the drain-source voltage of primary device Q_2 , i_{L_r} is the current through resonance inductor L_r . As seen in Figure 3.17, the primary side devices operate under ZVS in both full-load and light-load conditions over the entire switching

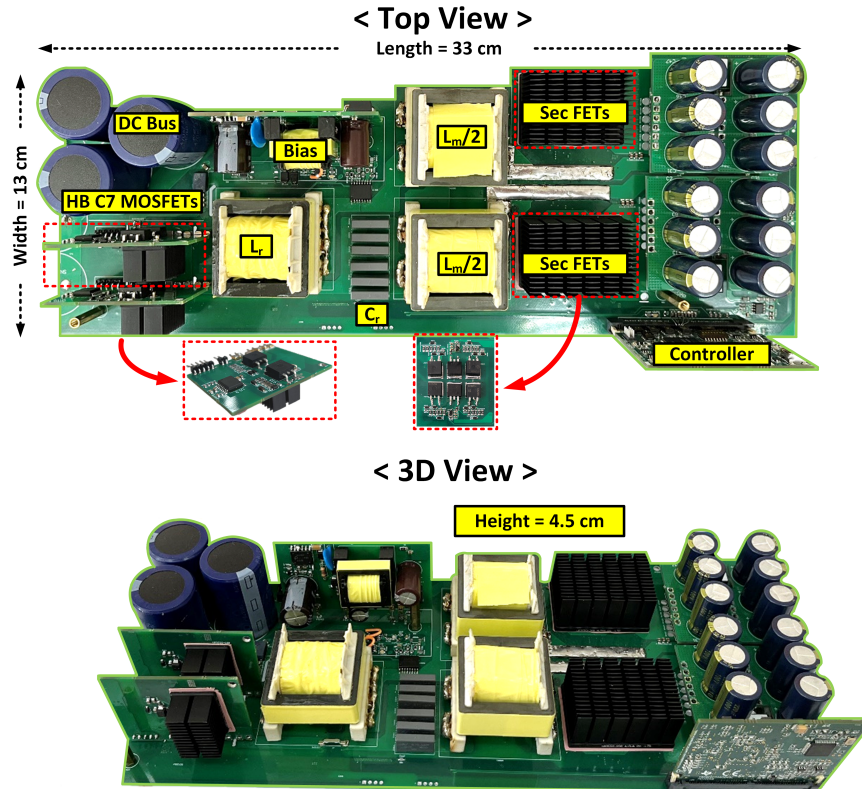


Figure 3.16 : Prototype of the designed 400 V/48 V LLC resonance converter rated at 3700 W.

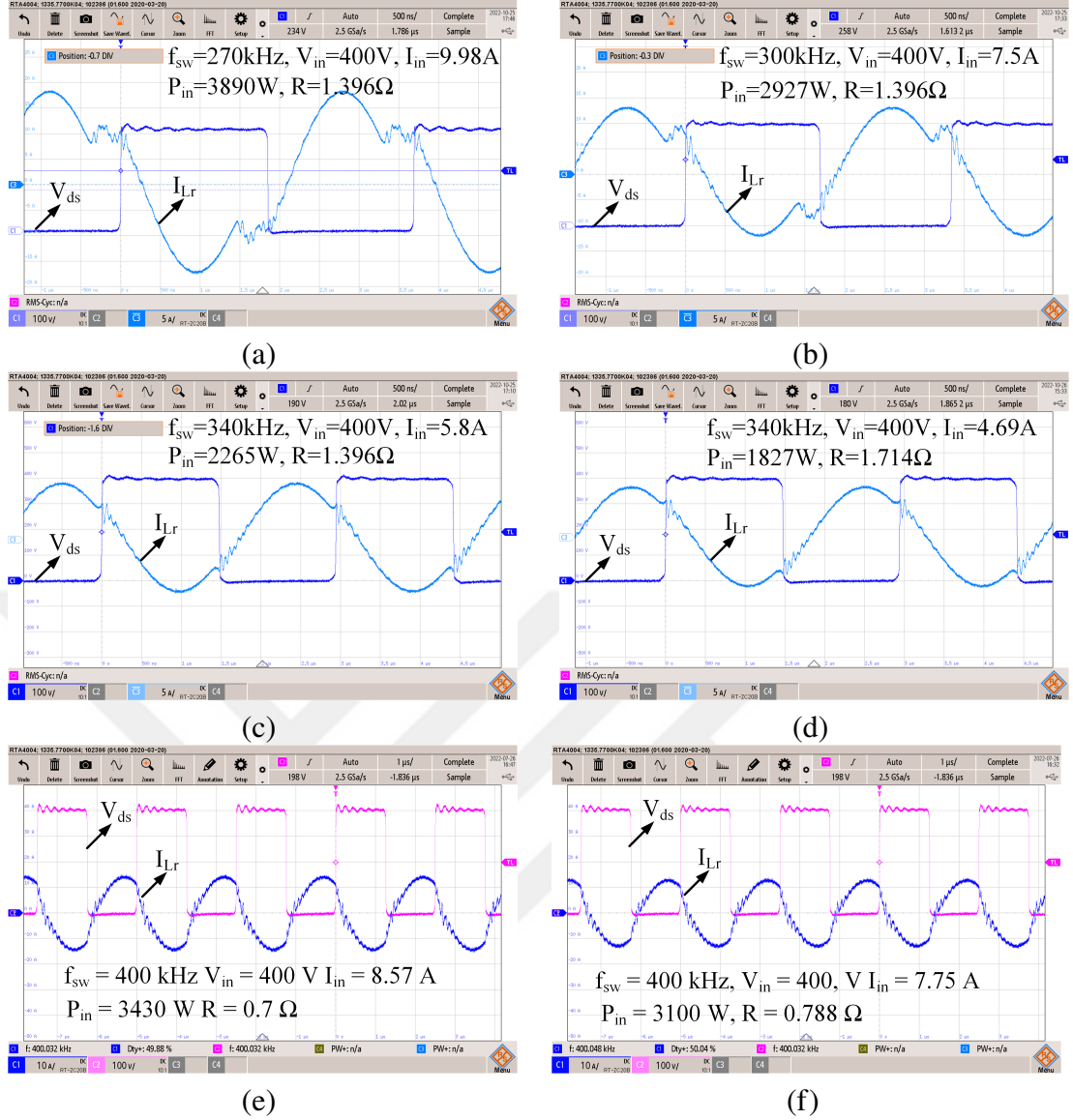


Figure 3.17 : Experimental waveforms for; (a) $f_s=270\text{ kHz}$, $P_{in}=3890\text{ W}$, (b) $f_s=300\text{ kHz}$, $P_{in}=2927\text{ W}$, (c) $f_s=340\text{ kHz}$, $P_{in}=2265\text{ W}$, (d) $f_s=340\text{ kHz}$, $P_{in}=1827\text{ W}$.

frequency range. The oscillation that starts after $i_{Lr}=i_{Lm}$ is due to the resonance between the leakage inductance and secondary rectifier's junction capacitance in the below resonance region.

The gain curve obtained by the FHA analysis, simulation and experimental data at the nominal operation point are shown in Figure 3.18(a). As seen from Figure 3.18(a), the gain obtained from the simulations is very similar to that obtained by the experiments. Since the effects of harmonics other than the fundamental harmonics increase in the below resonance region, the gain obtained in this region from FHA is lower than that

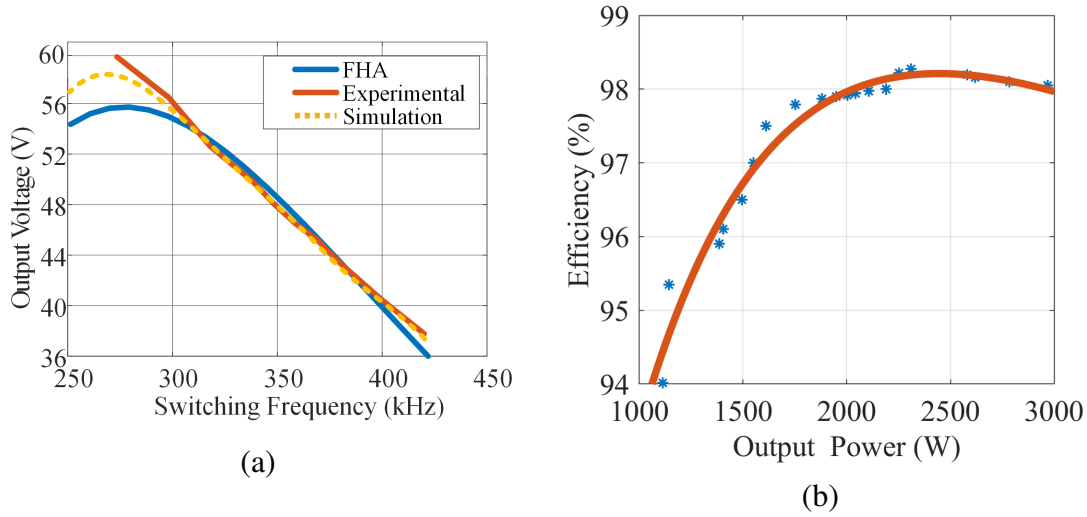


Figure 3.18 : Converter characteristics; (a) gain curve comparison at nominal operation point, (b) measured efficiency.

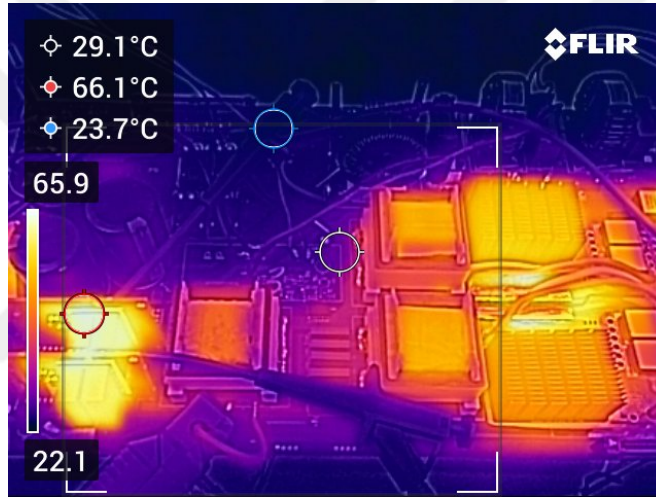


Figure 3.19 : Thermal test results at 2200 W under natural convection cooling.

in simulation and experiment. Moreover, the lumped interwinding and intra-winding parasitic capacitances of the transformer impacts the voltage gain curve slightly.

The expected efficiency at the turning operation point, where the transferred power is maximum, had been provided by the proposed tool as 98%. The designed converter exhibits a peak efficiency of 98.2% at around 60% load and 98% efficiency at 80% load as shown in Figure 3.18(b). Figure 3.19 provides the thermal test results of the proposed LLC converter under natural convection cooling switched at 340 kHz transmitting a power of 2200 W at 25 °C ambient temperature. The highest temperature is 66.1 °C and occurs on the primary devices. As predicted by the optimization tool, 3700 W is achieved by 200 LFM cooling providing a maximum of 60 °C temperature rise on the FETs.

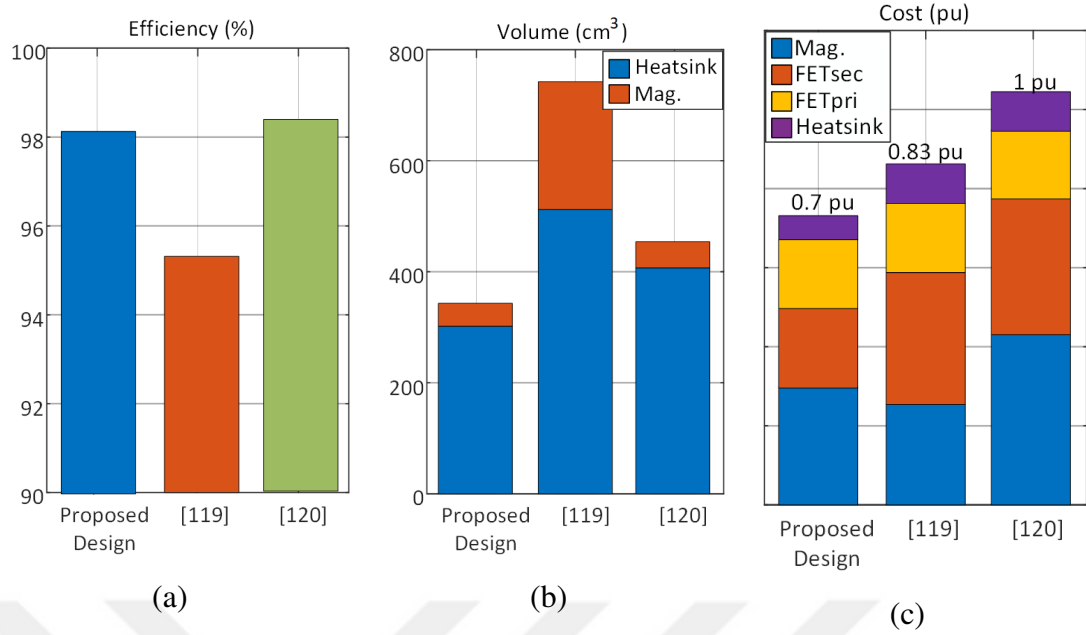


Figure 3.20 : Comparison of the proposed algorithm results with prototypes in [119] and [120] for a 3 kW design; (a) efficiency, (b) volume, and (c) cost.

To demonstrate the benefits of the suggested method, the recommended optimization technique has been implemented using volume-weighted design parameters for two benchmark designs presented in [119] and [120] with 3 kW output power and an output voltage range of 48-54V. Efficiency, boxed volume, and cost are compared for each design, and the results are displayed in Figure 3.20 as stacked bar charts. As the algorithm aims to minimize design volume, it achieves a slightly lower full load efficiency (98.1%) compared to [120] (98.4%), but significantly higher than [119] (95.3%).

Additionally, it suggests a much smaller boxed volume of 343 cm³, in contrast to the volumes of [119] and [120], which are 740 cm³ and 452 cm³, respectively.

The improvement in volume can also be observed separately for each design in the figure, specifically in terms of the magnetic components' volume, which comprises the total volume of the high-frequency transformer and the resonant inductor. Moreover, the proposed algorithm yields highly satisfactory results in terms of cost. The design in [120] is the most expensive due to the utilization of 16 switches in the secondary part and four magnetic components, including two parallel transformers and two series inductors. As a result, the cost of [120] serves as the base value for comparison and is considered as 1 pu. In contrast, [119] has the most cost-effective magnetic design, as the transformer's leakage inductance is used as a resonant inductor. However, due

to the use of 8 MOSFETs with very small R_{dson} values on the secondary side, the total cost of the [119] design rises dramatically to 0.83 pu. Finally, the proposed design's cost is determined to be 0.7 pu, making it the most affordable option with 2 parallel transformers, a resonant inductor, and 8 MOSFETs on the secondary side.

In this study, a systematic design approach has been presented for LLC converter that determines the resonance frequency, minimum and maximum switching frequency limits that satisfy the quality factors at corresponding voltage gains of a LEV battery charger by minimizing the system level fitness function for optimum design. During the design and optimization process L_r , L_m , and C_r values have been calculated for each resonance frequency. For each L-L-C combinations, a transformer and an inductor considering distributed core structure have been designed based on the developed analytical models for every possible core in the database. Among the designed candidates, the magnetic designs that minimize the local fitness function are recorded. For the operation conditions of each combination, the primary and secondary power device losses have been analytically calculated, and the required heat-sink volumes are estimated by using thermal models of bottom-side cooled devices cooled by mounting heat-sinks on a 4-layer 1.6 mm thick PCB.

After obtaining a design space, a system level fitness function including volume, efficiency and cost has been applied to find the optimal design. The proposed framework is essential to size the LLC converter optimally based on the given weights between cost, size and efficiency considering the load profile of a LEV battery charger. A 3700 W/48 V LLC prototype has been designed based on the outcome of the proposed framework, where minimum peak efficiency was constrained to 97%. The designed prototype achieves a peak efficiency of 98.2% at 60% load.

3.3 High Precision Model of LLC Resonant Converter Including Parasitic Elements

Various simulation programs are used for analyzing resonant converters. These analyses become more accurate and closer to experimental results when the parasitic values of the power components used in these programs are correctly estimated and included in the simulation model. For more realistic modeling, parameters like the parasitic capacitances of switching elements, transformer windings, leakage inductance, and the

parasitic capacitance of the resonance inductance should be considered. These parasitic capacitances arise from voltage differences between windings and the insulating material around the winding material. In LLC resonant converters, adjusting the parasitic values of transformers can help optimize the system's size. To reduce the circuit size, transformer windings are designed to provide leakage inductance equivalent to the desired resonance inductance, or they are wound to have a high leakage value to reduce the externally connected resonance inductance. Additionally, the required series or parallel capacitance for resonance can be integrated into the transformer by using high permeability magnetic materials and implementing winding methods on PCB. However, this PCB winding method is limited to several kilowatts and is not suitable for high-voltage applications [23].

Depending on the resonance topology used, unwanted parasitic capacitances can occur. For example, parasitic capacitance between primary windings of a transformer is beneficial for parallel resonance circuits, but detrimental for series resonance or LLC resonance circuits. These parasitic capacitors can cause unwanted high-frequency oscillations in the circuit, leading to high EMI [103]. This interference can cause errors in feedback readings and unintentional switching of power devices at improper times. The parasitic inductance of power switching elements significantly affects their characteristics by slowing down the switching speed and causing voltage overshoots and oscillations. In primary power stages with high-voltage switches, the created capacitance plays a role in determining parameters like magnetizing current and dead-time durations in resonant converters [104]. In soft-switching topologies, the main goal is to discharge these parasitic capacitances before the switches conduct, ensuring smooth switching. Therefore, knowing the parasitic capacitances of primary switches is crucial for circuit design. In LLC converters, primary power stage switches always engage in soft-switching. However, the secondary power stage operates intermittently under below-resonance conditions. In this case, circuit oscillations depend on the parasitic capacitances of secondary switches. Accurate modeling requires knowledge of these capacitances. In this study, a detailed modeling of an LLC Resonant Converter used in a 400-48 V, 3.7 kW battery charging system is conducted, including its parasitics, and then compared with experimental results.

3.3.1 Parasitics in Magnetic Elements

In the circuit shown in Figure 3.21, the transformers have three types of parasitic capacitances: primary side, secondary side, and between windings. These parasitic capacitances arise due to the voltage difference between the windings, which are surrounded by an insulating layer. The magnitudes of the primary leakage inductance (L_{l1}), secondary leakage inductance (L_{l2}), and magnetizing inductance are particularly significant as they directly affect the LLC resonant circuit's performance.

As illustrated in Figure 3.22, when the secondary side is left open-circuited and measurements are taken from the primary side at a low frequency using an RLC meter, the sum of the primary leakage inductance and magnetizing inductance (L_{ad}) is obtained. At this frequency, the impedance exhibited by the winding capacitances is very high, so in the equivalent circuit, these capacitances are considered as open circuits [103].

When the secondary side of the transformer is short-circuited as shown in Figure 3.23 and measurements are taken from the primary side, the sum of L_{l1} and the reflected value of L_{l2} on the primary side is obtained. This measurement is crucial for understanding the inductance dynamics of the transformer.

Using the equation that relates the winding and inductance ratios, the primary leakage inductance can be calculated as shown in equation (3.21). In this equation, L_{kd} represents the inductance value measured with an RLC meter when the secondary side is short-circuited. By leveraging these relationships and measurements, it's possible to accurately determine the primary leakage inductance, which is vital for the accurate modeling and efficient operation of LLC resonant converters.

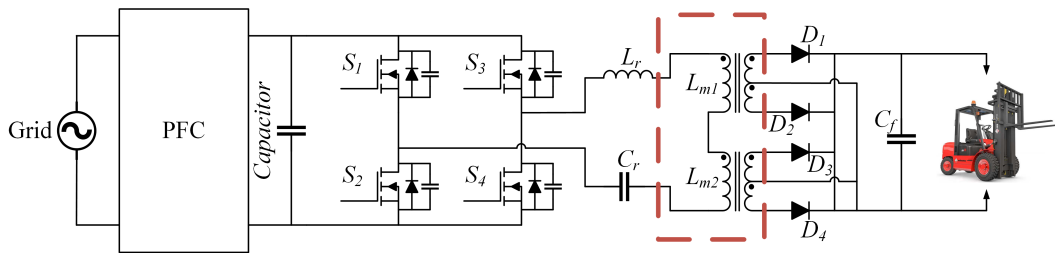


Figure 3.21 : Primary-side series secondary-side parallel connected LLC multi-core structure.

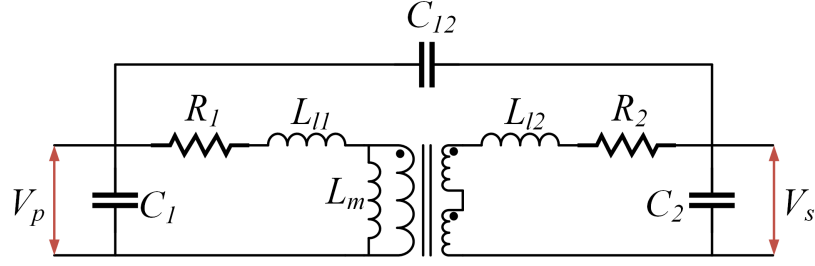


Figure 3.22 : Transformer equivalent circuit with secondary side open-circuit.

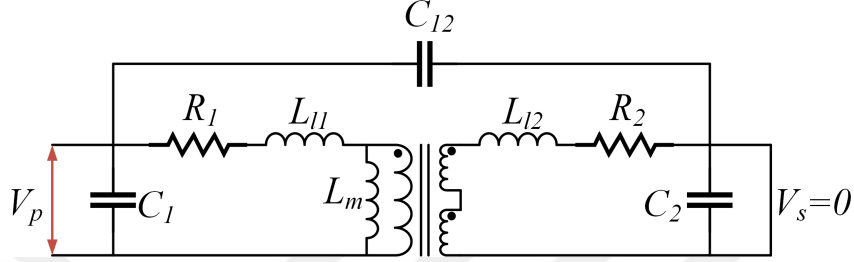


Figure 3.23 : Transformer equivalent circuit with secondary side short-circuit.

$$\begin{aligned} L_{l1} + L'_{l2} &= L_{kd} \\ L_{l1} + \frac{L_{l1}}{n^2} &= L_{kd} \end{aligned} \quad (3.21)$$

After calculating both the primary and secondary leakage inductance values, the magnetizing inductance can be determined. This is done by leaving the secondary side open-circuited and measuring the inductance from the primary side, which yields the inductance value (L_{ad}). From this L_{ad} value, the primary leakage inductance is subtracted to find the magnetizing inductance, as described in equation (3.22). This step is crucial for accurately characterizing the transformer's behavior in the LLC resonant converter and for ensuring optimal performance.

$$L_{l1} + L_m = L_{ad} \quad (3.22)$$

To measure the winding resistance values in transformers, a similar measurement procedure is followed. Firstly, with the secondary side open-circuited, the primary winding resistance is measured. Then, with the secondary side short-circuited, the combined resistance of the secondary winding's reflected value to the primary side and the primary winding resistance is obtained, as shown in Figure 3.23 and Figure 3.24.

To measure the parasitic capacitance (C_{12}) between primary and secondary windings, both sides of the transformer are short-circuited simultaneously, and the measurement

is repeated using an RLC meter. The measurement directly corresponds to the inter-winding parasitic capacitance.

To measure the primary side parasitic capacitance (C_1), the secondary side is short-circuited as illustrated in Figure 3.24. When the secondary side is reflected to the primary side, the equivalent circuit becomes as shown in Figure 3.25, and the measurement is taken from the primary side. During this measurement, the winding resistances can be neglected due to their minimal impact compared to the winding impedance at high frequencies.

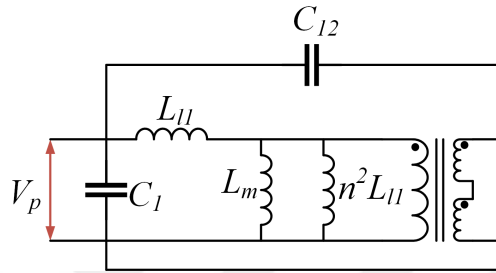


Figure 3.24 : Equivalent circuit used in the calculation of C_1 .

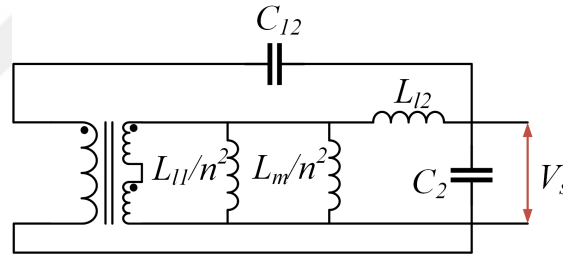


Figure 3.25 : Equivalent circuit used in the calculation of C_2 .

In the equivalent circuit shown in Figure 3.25, there are three parallel branches. Since a C_1 and C_{12} capacitances are in parallel, they are added together. The equivalent of the remaining inductances is calculated as shown in equation 3.23. In this equation, L_{C1} represents the equivalent inductance measured while targeting the C_1 capacitance. Utilizing equation (10), the total of C_1 and C_{12} capacitances can be calculated using the equivalent impedance and phase angle read from the RLC meter, along with previously measured leakage and magnetizing inductances. Finally, by subtracting the previously obtained C_{12} capacitance from this total, the C_1 capacitance, i.e., the parasitic capacitance between the primary windings, can be determined.

$$L_{C1} = L_{l1} + \frac{L_m \cdot n^2 L_{l2}}{L_m + n^2 L_{l2}} \quad (3.23)$$

A similar experimental method can be used to determine the secondary side parasitic capacitance C_2 as well. In this case, the primary side is short-circuited, and measurements are taken from the secondary side. According to the equivalent circuit shown in Figure 3.26, the capacitances C_2 and C_{12} are in parallel, and the equivalence of the remaining inductances can be calculated as shown in equation (3.24). This approach allows for an accurate assessment of the secondary side's parasitic capacitance, crucial for understanding and optimizing the performance of the transformer in an LLC resonant converter.

$$L_{C2} = L_{l1} + \frac{L_m L_{l1}}{(L_m + L_{l1})n^2} \quad (3.24)$$

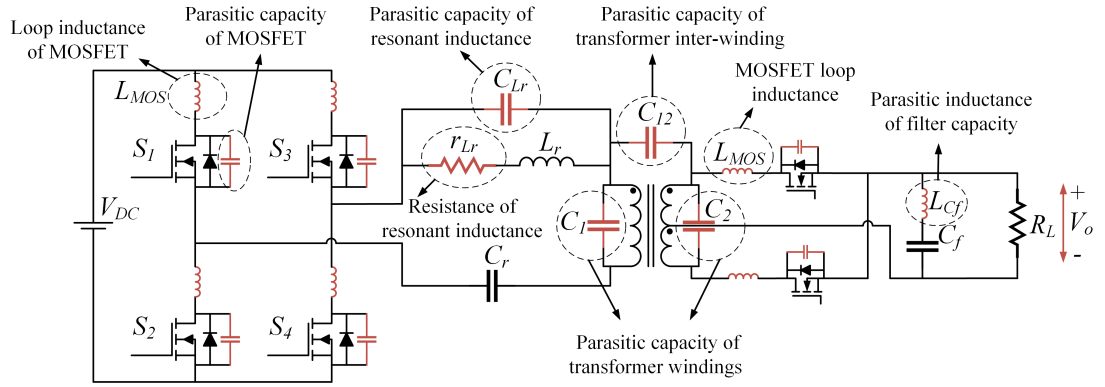


Figure 3.26 : LLC resonant converter with parasitic elements.

The inductance value of the resonance inductance in an LLC resonant converter can be directly measured at the LLC resonance frequency using an RLC Meter, while its capacitance value can be measured at a higher frequency. The parasitic values obtained from these measurements are listed in Table 3.6. The higher the frequency at which the internal winding parasitic capacitance measurements for the primary and secondary sides are performed, the less significant the effects of parallel L_{C1} and L_{C2} become.

However, the maximum frequency range of the used LCR meter is 8 MHz, and therefore, the winding parasitic capacitance measurements are conducted at this frequency. Since measurements at higher frequencies are not possible with this meter, there might be some error in determining the parasitic capacitance values. This limitation is important to consider in the analysis and optimization of the LLC resonant converter's performance.

Table 3.6 : Measured parasitic values of transformers.

Parameter	Value
L_m	17.5134 μH
L_{11}	0.5616 μH
L_{12}	0.0702 μH
$L_{r,cp}$	1.5 pF
$R_{P,DC}$	13 m Ω
$R_{S,DC}$	4 m Ω
C_1	35 pF
C_2	250 pF
C_{12}	31.5 pF

3.3.2 Semiconductor Power Switches Interferences

The output capacitance (C_{oss}) of semiconductors is a crucial parameter in many power converters. In hard-switching converters, the C_{oss} capacitance of a MOSFET gets charged when the MOSFET is off and discharges upon turning on.

In converters where Soft-Switching Gate Drive (SGA) conditions are met, the discharge of C_{oss} occurs during the off period of the MOSFET. This ideally results in the elimination of switching losses. In all power converters, the energy stored in the C_{oss} capacitance influences component selection, thermal design, and efficiency, making its accurate measurement critical [105].

C_{oss} values provided in data sheets are measured by applying a small AC disturbance signal to a steady-state DC source-drain voltage. Then, based on the MOSFET's voltage and C_{oss} value from the datasheet, the C_{oss} value in the circuit for the applied voltage is recalculated using an equation like (3.25).

$$Q_{oss} = \int_0^{V_{DS}} C_{oss}(V) dV \quad (3.25)$$

$$C_{oss}(V) = \frac{Q_{oss}}{\Delta V}$$

The circuit diagram of the simulation model, created for a detailed consideration of transient events and to achieve high accuracy between simulation and real-world data, is shown in Figure 3.27. In addition to the previously mentioned parasitic component values, the parasitic resistance of the resonance inductance was measured using an RLC meter, and its parasitic capacitance was found to be 1.5 pF. The effects of these parasitic elements on the circuit have been observed to be negligible.

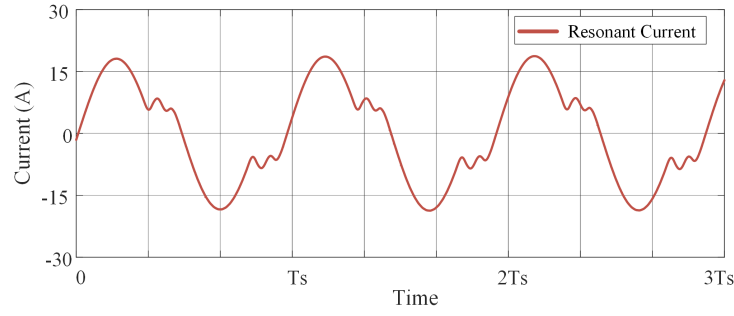


Figure 3.27 : Resonant current waveform obtained from the simulation model in below resonant operation.

In the primary side of the analyzed circuit, high voltage switches enter cut-off under current, leading to a higher di/dt value at cut-off compared to the conduction phase. This energy accumulates in the loop inductance, causing oscillations in the circuit. The loop inductance in the analyzed circuit is around 7 nH. Finally, the parasitic inductance of the output filter capacitor connected parallel to the load has been added to the simulation model. In the circuit, eight 4.7 $\mu\text{F}/100\text{ V}$ surface-mount film capacitors, housed in 6054 cases, are connected in parallel.

In the primary side of the circuit, high-voltage switches enter cut-off under load current, leading to a high di/dt value at the moment of cut-off. This high di/dt causes energy to accumulate in the loop inductance, which can lead to oscillations within the circuit. The parasitic capacitances of the inverter switches on the primary side are discharged before the switches are turned on, using soft-switching techniques. The smaller these parasitic capacitance values, the less current is needed for soft-switching, allowing the circuit to be switched at higher frequencies. The impact of these parasitic capacitances on the resonance current, particularly in the below-resonance operation region, is illustrated in Figure 3.27 using the simulation model. In comparison, the resonance current measured experimentally under the same operating conditions is shown in Figure 3.28. The effects of transformer primary side (C_1), secondary side (C_2), and inter-winding (C_{12}) parasitic capacitances on the transformer voltages are depicted in Figure 3.29. Finally, Figure 3.30 presents experimental measurement results showing the secondary side transformer voltage. These figures and experimental data highlight the importance of accurately accounting for parasitic elements in the design and simulation of high-frequency power converters, ensuring that their real-world performance aligns closely with theoretical predictions.

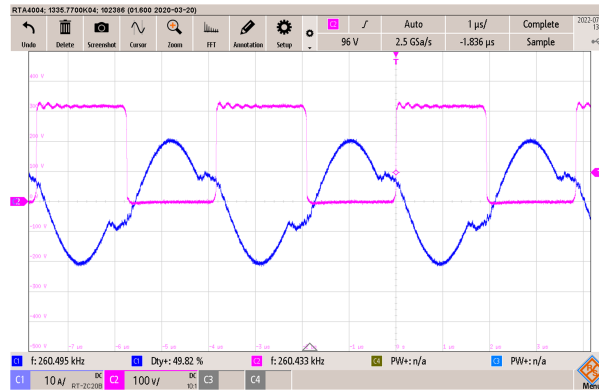


Figure 3.28 : Experimentally measured resonance current in below resonant operation.

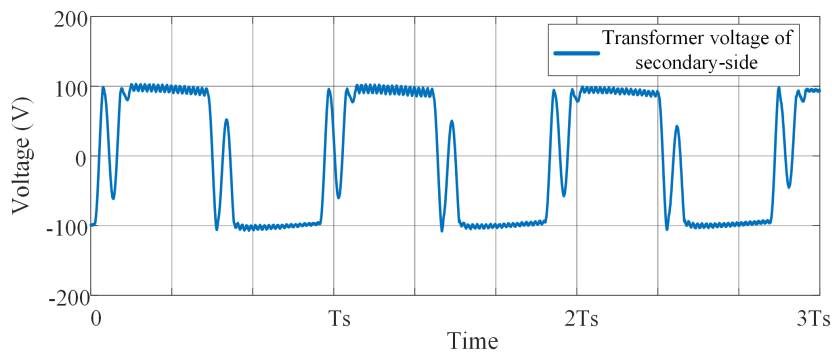


Figure 3.29 : Secondary side transformer voltage obtained from the simulation model.

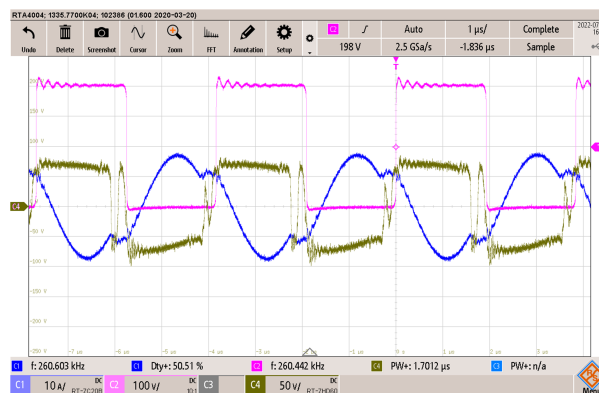


Figure 3.30 : Experimental measurement results; primary side switch voltage, resonance current, secondary side transformer voltage.

4. CLOSED LOOP CONTROL OF LLC RESONANT CONVERTER

Since the Li-ion battery technologies developed to drive electric motors in EVs have different operating modes such as CC and CV, it is necessary to design battery chargers that can adapt its control loops to the state of charge of the battery. Recently, several studies have been conducted on reliable and efficient battery chargers for EVs and their control. Among the power converter topologies, the LLC resonant converter is one of the widely preferred topologies for battery chargers due to its ZVS operation in a wide frequency and load range. This topology is proven to achieve both high-efficiency and high-power-density designs [8, 20, 24, 25]. LLC converters can obtain ZVS for zero to rated load with a low turn-off current for the primary switches, while achieving ZCS for the SRs. Soft turn-on and turn-off features allow minimizing the EMI [25]. However, as it is controlled with a variable switching frequency, control of LLC resonant converter remains as a difficult task, particularly for varying loads such as batteries in EVs [29].

Since a regulated output is desired in both PWM converters and resonant converters, a feedback loop is included in the control system to stabilize the system and regulate the output voltage or current. In general, small signal equivalent circuit models are used to design a robust control system for all converter structures. There are many approaches using small signal equivalent circuit models for PWM converters such as voltage mode control [81, 82], current mode control including inductor current sideband [83–86], V^2 control including both inductor current sideband and capacitor voltage sideband [87–89]. Since some state variables do not have dc components and contain switching frequency harmonics, the state-space averaging method is not preferred for resonant converters. Owing to the oscillations that occur in resonance situations, the switching frequency interacts with the resonance frequency. This is explained with the beat frequency dynamics, where a pair of double-poles located at the beat frequency determines the high frequency response [90]. Therefore, using an extended describing function method that considers the states of the resonance

frequency relative to the switching frequency provides the most successful approach for a robust LLC controller design [91].

CC and CV charging is a commonly used charging strategy for battery charging applications. The charging characteristic of a single Li-ion battery cell is shown in Figure 4.1. As shown in Figure 4.1, there are several key points in the charging process, namely, Region 1 to Region 7. Region 1 and Region 7 correspond to the beginning and end of the charging process, respectively. At the Region 4, the battery voltage is equal to the nominal voltage of the battery pack. The Region 6 marks the transition from CC to CV charging mode [9].

As seen from Figure 4.1, the dynamic response of LLC converter in battery charger changes with respect to these operation points. This requires a closed control system design that takes into account the battery current and battery voltage separately in the battery charging application.

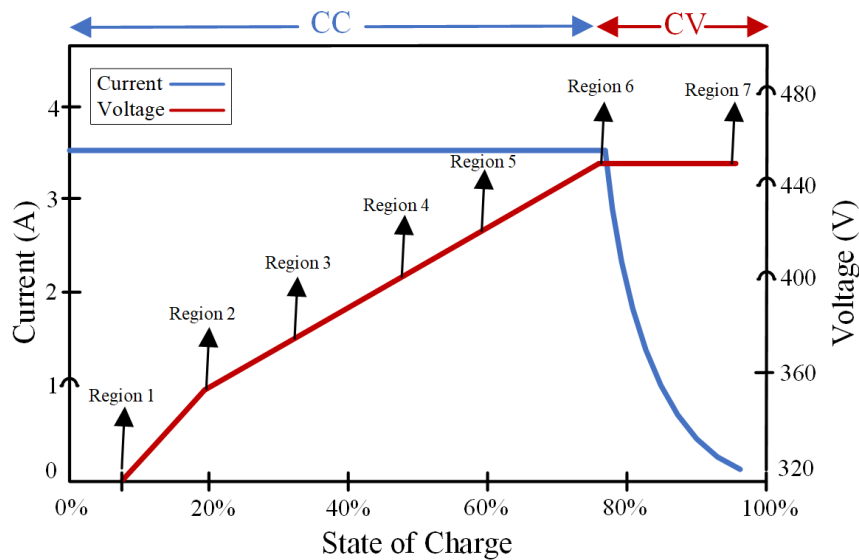


Figure 4.1 : Typical charging characteristics of a Li-ion battery cell.

This study proposes a closed-loop control for an LLC resonant converter, shown in Figure 4.2, designed for battery charging applications, in which the output current and output voltage are both considered. Extended describing function method considering the states of the resonance elements relative to the switching frequency for equivalent circuit approximation has been proposed for modeling LLC converter. Thus, a simple third order equivalent circuit model of the LLC resonant converter was obtained and the beat frequency dynamics and small signal behavior of the converter were very well

predicted with this circuit model. A proportional-integral control has been applied to the each transfer function.

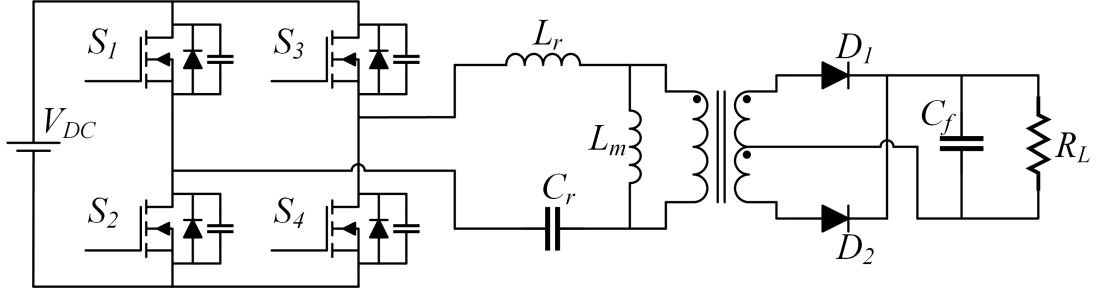


Figure 4.2 : Fundamental topology of LLC resonant converter.

4.1 Extended Describing Function Modelling of LLC Converter

The dynamic response of an LLC converter varies depending on its operating points. From the perspective of resonance, the behavior of the resonant tank differs when comparing scenarios where the switching frequency is greater than or equal to the resonant frequency with those where f_s is less than f_0 . In the case of f_s being greater than or equal to f_0 , only the resonant inductance resonates with the resonant capacitor, and the magnetization inductance is limited by the output voltage. This distinct resonant behavior necessitates the development of separate small-signal models for each case. As depicted in Figure 4.3, when f_s is greater than or equal to f_0 , the tank current lacks DC components but contains strong harmonics at the switching frequency. As a result, the concept of averaging, typically used in analyzing such systems, breaks down. This breakdown indicates a need for different analytical approaches or models to accurately understand and predict the behavior of the LLC converter under these specific conditions [121].

Up to this point, the most successful equivalent circuit model for LLC converters is the one proposed by E. Yang in [122], based on the extended describing function concept. During the development of this model, the fundamental approximation is used, meaning that only the fundamental components of the resonant tank variables are considered. The describing function method is applied separately to the inverter, rectifier, and resonant tank. For the inverter section, as depicted in Figure 4.4(a) [121], different approximation approaches are used for the input and output. The input of the inverter utilizes an average approximation, whereas the output employs a fundamental

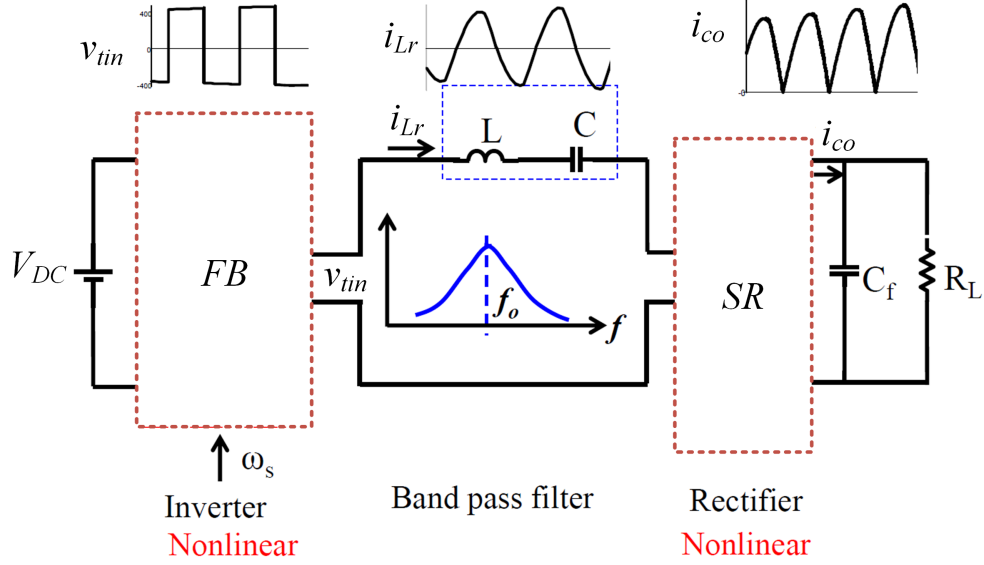


Figure 4.3 : Diagram of LLC converter with waveforms of major signals when $f_s \leq f_o$.

approximation. The small-signal model, shown in Figure 4.4(b), illustrates these approximations. On the input side, the model is represented as a controlled current source because the average value of the input current is correlated with the resonant current. On the output side, the model takes the form of a controlled voltage source, as the magnitude of the fundamental component v_{tin1} is related to the input voltage. This model effectively captures the essential dynamics of the LLC converter at different operating points, providing a more accurate and comprehensive understanding of its behavior. The expressions are shown in equations (4.1) and (4.2).

$$\langle \hat{i}_{in} \rangle = \frac{2}{\pi} \text{Re}(\hat{\mathbf{i}}_{Lr}) \quad (4.1)$$

$$\hat{\mathbf{v}}_{tin,1} = \frac{4}{\pi} \hat{v}_{DC} \quad (4.2)$$

For the rectifier in the LLC converter, the modeling approach is akin to that used for the inverter. As shown in Figure 4.5(a) [121], the waveforms indicate that the rectifier's input adopts a fundamental approximation, while the output uses an average approximation. The small-signal model for the rectifier, depicted in Figure 4.5(b), reflects these approximations in its structure. On the input side of the rectifier, the model is conceptualized as a controlled voltage source. This is based on the relationship between the magnitude of the fundamental component, $(v_{tout,1})$, and (v_o) .

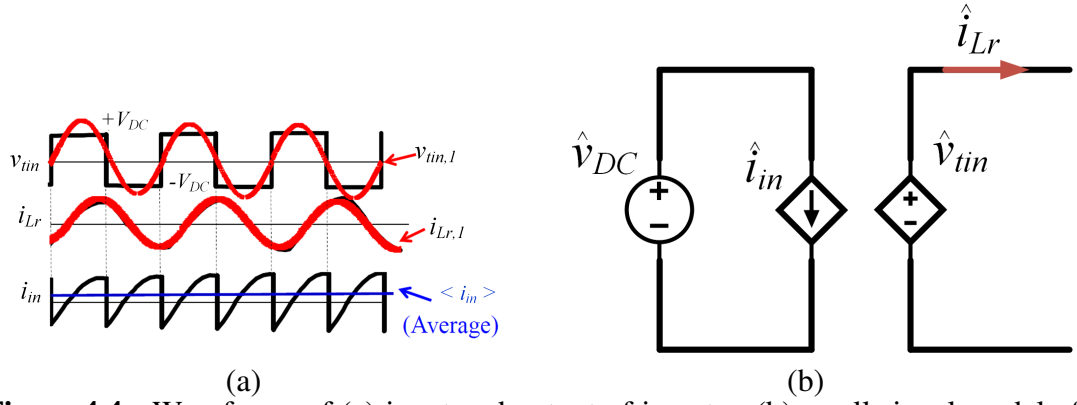


Figure 4.4 : Waveforms of (a) input and output of inverter, (b) small signal model of the inverter.

On the output side, the model is represented as a controlled current source. This is because the average value of the output current is related to the resonant current. This method of modeling allows for a more nuanced understanding of the rectifier's function within the LLC converter. By using different approximations for the input and output, the model can accurately reflect the distinct operational characteristics of each stage, providing valuable insights for the analysis and optimization of the converter's overall performance. The expressions are shown in equations (4.3) and (4.4) .

$$\hat{v}_{tout,1} = k_{vo} \hat{v}_o + k_{is} \text{Re}(\hat{\mathbf{i}}_{Lr,1}) + k_{ic} \text{Im}(\hat{\mathbf{i}}_{Lr,1}) \quad (4.3)$$

$$\langle |\hat{i}_{sec}| \rangle = k_s \text{Re}(\hat{\mathbf{i}}_{Lr,1}) + k_c \text{Im}(\hat{\mathbf{i}}_{Lr,1}) \quad (4.4)$$

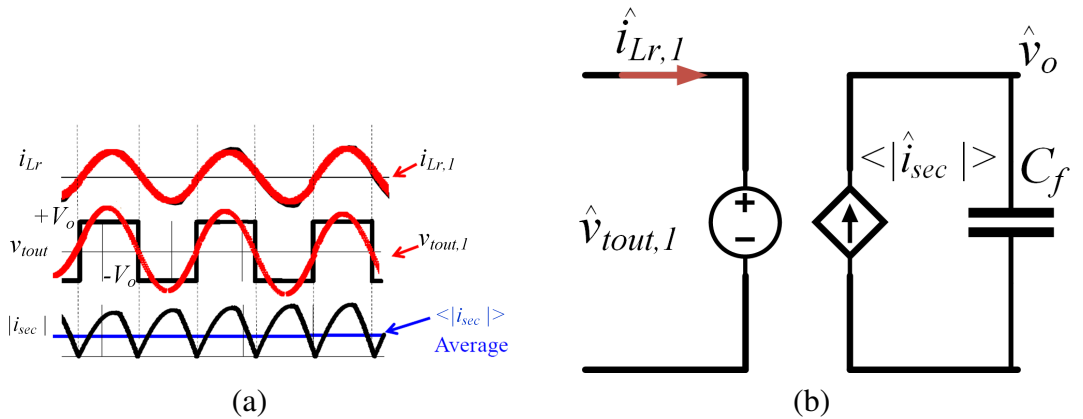


Figure 4.5 : Waveforms of (a) input and output of rectifier, (b) small signal model of the rectifier.

In the LLC converter model, the small-signal models for both the resonant inductor and the resonant capacitor, which are based on the describing function method, are

illustrated in Figures 4.6 and 4.7, respectively. These models are tailored to capture the effects of switching frequency on these components. For the resonant inductor model shown in Figure 4.6, there is a complex impedance that reflects the impact of the switching frequency. Additionally, this model includes a complex voltage source, which represents the perturbations caused by variations in the switching frequency. Similarly, in the resonant capacitor model presented in Figure 4.7, there is also a complex impedance due to the switching frequency effect. However, instead of a complex voltage source, this model features a complex current source. This current source accounts for the perturbations in the switching frequency specifically related to the resonant capacitor.

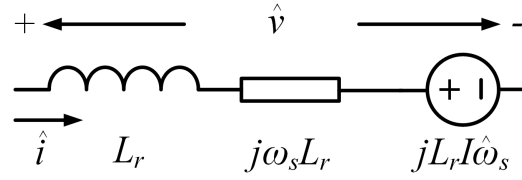


Figure 4.6 : Small signal model of the resonant inductor.

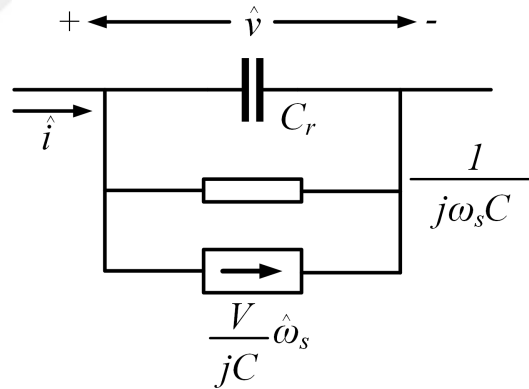


Figure 4.7 : Small signal model of the resonant capacitor.

4.1.1 Equivalent circuit model of LLC for $f_s \geq f_0$

In modeling the magnetizing inductor, a basic approximation similar to that used for the resonant inductor is employed. This involves considering only the primary component of the triangular switching frequency waveform while ignoring the harmonics. Using a method akin to that for the resonant inductor, a small-signal model of the magnetizing inductor is formulated, as depicted in Figure 4.8(a) [121]. When the modulation frequency (ω_m) is significantly lower than switching frequency (ω_s), the voltage

across the inductor L_m becomes negligible compared to the voltage across the complex impedance $j\omega_s L_m$. This leads to a more simplified small-signal model, illustrated in Figure 4.8(b). The model includes a complex impedance resulting from the switching frequency and a complex voltage source caused by the switching frequency's perturbation. It's important to note that the dynamic element L_m is absent in the small signal model. This is justified because, in this scenario, L_m is constrained by the output voltage, rendering its current no longer a state variable.

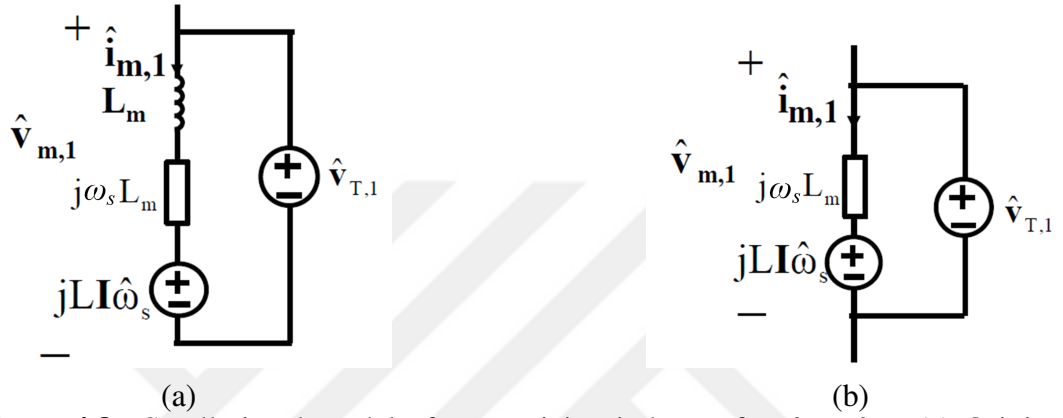


Figure 4.8 : Small signal model of magnetizing inductor for $f_s \geq f_0$. (a) Original small signal model, (b) simplified small signal model.

Integrating the small signal models of various components such as the inverter, resonant inductor, magnetizing inductor, capacitor, rectifier, and output filter, the small signal model of an LLC resonant converter operating at $f_s \geq f_0$ is developed, as illustrated in Figure 4.9. However, this equivalent circuit model is not suitable for simulation due to the complexity of the impedance and controlled sources.

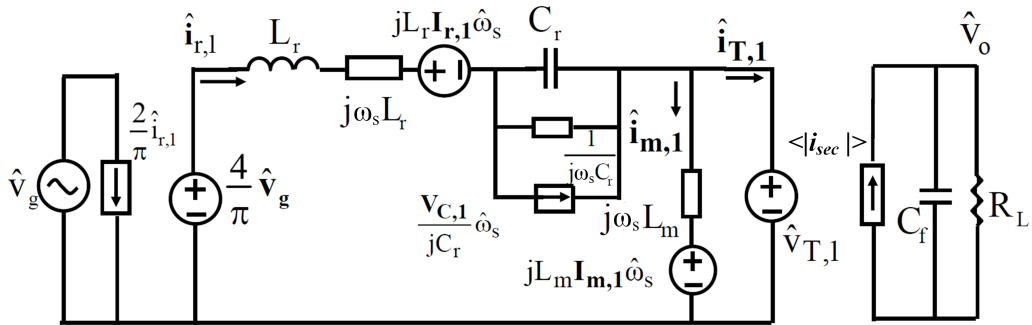


Figure 4.9 : Small signal model of LLC for $f_s \geq f_0$.

E. Yang's approach [121] to creating an equivalent circuit model suitable for simulation involves separating the sine and cosine components, which are conceptually orthogonal, as demonstrated in Figure 4.10 [121]. This method effectively eliminates the

complex terms, although it introduces some coupling terms. This approach is logical, considering that the essence of the complex term is to transform the sine part into the cosine part and vice versa. The expression of equivalent inductor L_e and impedance X_{eq} are shown in equations (4.5) and (4.6):

$$L_e = L_r + \frac{1}{C_r \omega_s^2} = L_r \left(1 + \frac{\omega_o^2}{\omega_s^2} \right) \quad (4.5)$$

$$X_{eq} = \omega_s L_r - \frac{1}{\omega_s C_r} \quad (4.6)$$

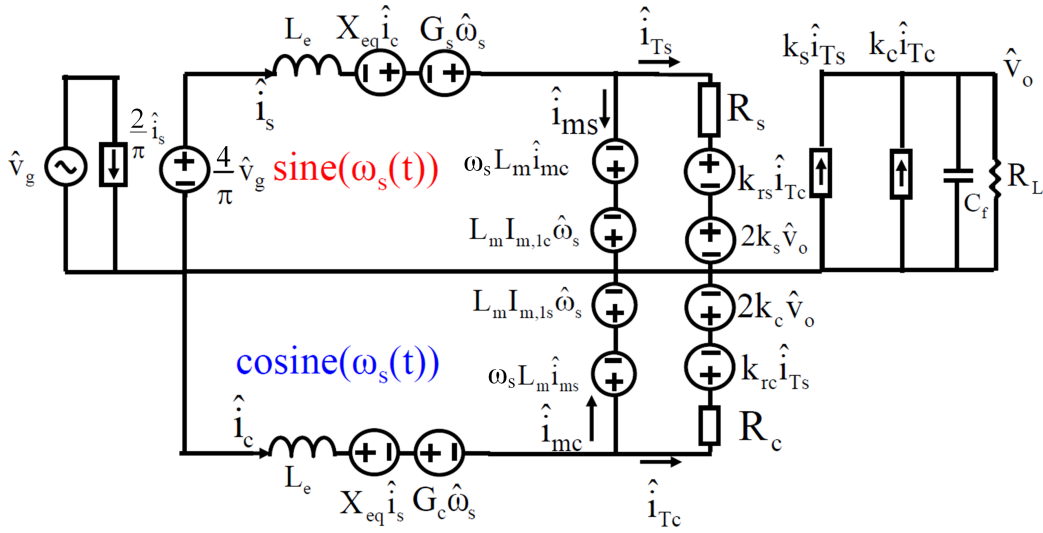


Figure 4.10 : Separated third-order equivalent circuit model of LLC for $f_s \geq f_0$.

The expressions of G_s , R_s , k_{rs} , k_s , G_c , k_{rc} and k_c are shown in equation (4.7):

$$\begin{aligned} G_s &= L_r I_{r,1c} + \frac{V_{C,1s}}{\omega_s}, G_c = L_r I_{r,1s} + \frac{V_{C,1c}}{\omega_s} \\ k_s &= \frac{4}{\pi} n \frac{I_{T,1s}}{\sqrt{I_{T,1s}^2 + I_{T,1c}^2}}, k_c = \frac{4}{\pi} n \frac{I_{T,1c}}{\sqrt{I_{T,1s}^2 + I_{T,1c}^2}} \\ R_s &= \frac{4n}{\pi} \frac{V_o}{\sqrt{I_{T,1s}^2 + I_{T,1c}^2}} \frac{I_{T,1c}^2}{I_{T,1s}^2 + I_{T,1c}^2}, \\ R_c &= \frac{4n}{\pi} \frac{V_o}{\sqrt{I_{T,1s}^2 + I_{T,1c}^2}} \frac{I_{T,1s}^2}{I_{T,1s}^2 + I_{T,1c}^2}, \\ k_{rs} &= k_{rc} = -\frac{4n}{\pi} \frac{V_o}{\sqrt{I_{T,1s}^2 + I_{T,1c}^2}} \frac{I_{T,1s} I_{T,1c}}{I_{T,1s}^2 + I_{T,1c}^2}, \end{aligned} \quad (4.7)$$

The expressions of $I_{r,1c}$, $I_{r,1s}$, $V_{C,1s}$, $V_{C,1c}$, $I_{T,1c}$, $I_{m,1s}$ and $I_{m,1c}$ are listed in equation (4.8):

$$\begin{aligned}
I_{T,1s} &= \frac{4 V_g}{\pi R_{eq}} \frac{F^2 L_n \left(L_n + 1 - \frac{1}{F^2} \right)}{F^2 \left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\frac{\pi^2}{8} Q (1 - F^2) L_n \right)^2}, \\
I_{T,1c} &= \frac{4 V_g}{\pi R_{eq}} \frac{F L_n^2 \frac{\pi^2}{8} Q (1 - F^2)}{F^2 \left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\frac{\pi^2}{8} Q (1 - F^2) L_n \right)^2}, \\
I_{m,1s} &= \frac{I_{T,1c} R_{eq}}{\omega_s L_m}, I_{m,1c} = -\frac{I_{T,1s} R_{eq}}{\omega_s L_m}, \\
I_{r,1s} &= I_{T,1s} + I_{m,1s}, I_{r,1c} = I_{T,1c} + I_{m,1c}, \\
V_{C,1s} &= \frac{I_{r,1c}}{\omega_s C_r}, V_{C,1c} = -\frac{I_{r,1s}}{\omega_s C_r}, R_{eq} = \frac{8}{\pi^2} n^2 R_L
\end{aligned} \tag{4.8}$$

The superposition theorem is utilized to develop a non-coupled equivalent circuit model, effectively eliminating the coupling effects observed in the previous approach. The resulting non-coupled equivalent circuit model is presented in Figure 4.11.

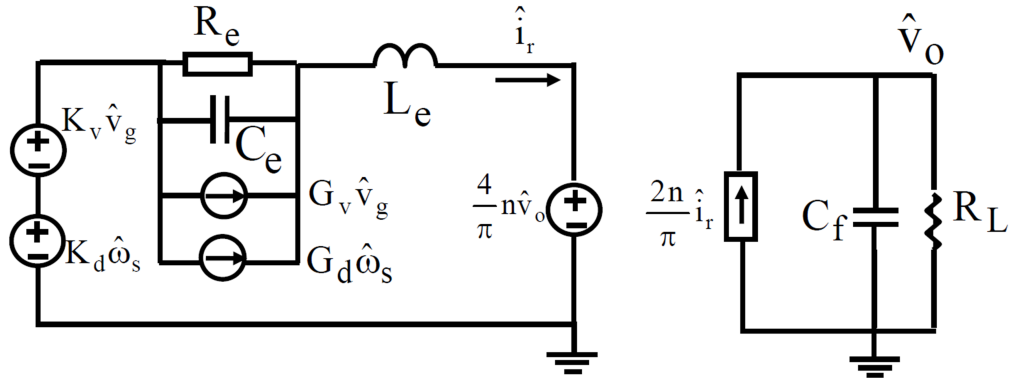


Figure 4.11 : Non-coupled third-order equivalent circuit model of LLC for $f_s \geq f_0$.

L_e , R_e , C_e are used to represent beat frequency dynamics. The expressions of K_v , G_v , K_d and G_d will be affected by the design of L_m . All the expressions are shown in equations (4.9) and (4.10).

$$\begin{aligned}
L_e &= L_r \left(1 + \frac{\omega_o^2}{\omega_s^2} \right), C_e = \frac{1}{L_e (\omega_s - \omega_o)^2}, R_e = \frac{L_e |X_{eq}| |\omega_s - \omega_o|}{R_{eq}} \\
G_d &= \frac{4V_g}{\pi} \frac{L_n}{\omega_o R_e} \left(\frac{2}{L_n^2} + \frac{\frac{1}{F} \left(\frac{1}{F^2} - F^2 \right) \left(\frac{\pi^2}{8} Q L_n \right)^2 - \left(L_n + 1 - \frac{1}{F^2} \right) \left(\frac{2}{F^3} \right)}{\left[\sqrt{\left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\left(\frac{1}{F} - F \right) \frac{\pi^2}{8} Q_r L_n \right)^2} \right]^3} \right),
\end{aligned} \tag{4.9}$$

$$\begin{aligned}
K_d &= -\frac{4V_g}{\pi} \frac{1}{\omega_0 L_n}, G_v = \frac{1}{\pi} \frac{X_{eq}}{\sqrt{X_{eq}^2 + R_{eq}^2}}, \\
K_v &= \frac{4}{\pi^2} \frac{V_g L_n F}{R_{eq}} \frac{L_n + 1 - \frac{1}{F^2}}{\sqrt{\left(L_n + 1 - \frac{1}{F^2}\right)^2 + \left(\left(\frac{1}{F} - F\right) \frac{\pi^2}{8} Q L_n\right)^2}}, \\
R_{eq} &= \frac{8}{\pi^2} n^2 R_L, X_{eq} = \omega_s L_r - \frac{1}{\omega_s C_r}
\end{aligned} \tag{4.10}$$

4.1.2 Equivalent circuit model of LLC for $f_s < f_0$

Figure 4.12 [121] illustrates the steady-state waveforms when the f_s is 0.8 times f_0 . In this scenario, the behavior of the resonant tank varies at different times. Specifically, during the intervals $[t_0, t_1]$ and $[t_2, t_3]$, the magnetizing inductor L_m is clamped by either V_o or $-V_o$, and does not engage in resonance. This situation mirrors the condition observed when f_s is greater than or equal to f_0 . Conversely, in the periods $[t_1, t_2]$ and $[t_3, t_4]$, the magnetizing inductor becomes part of the resonance, joining the series configuration of the resonant capacitor and resonant inductor. During these intervals, the output load is disconnected from the resonant tank. The alteration of the resonant tank within a single switching period signifies that the LLC resonant converter effectively adopts a multi-resonant structure when operating under the condition $f_s < f_0$ [121].

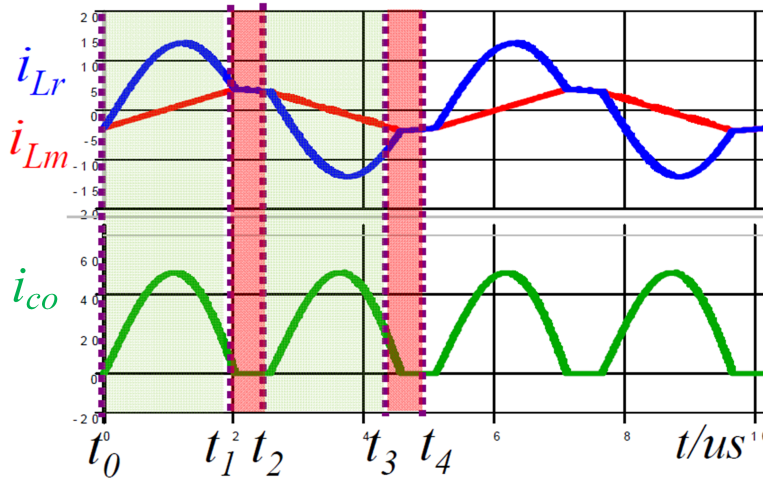


Figure 4.12 : Steady-state waveform for $f_s = 0.8 f_0$.

During the time intervals $[t_0, t_1]$ and $[t_2, t_3]$, the resonant tank and the load are coupled, and the relationship among the magnetizing inductor, the rectifier, and the output load is the same as in the scenario where the switching frequency is greater than or equal

to the resonant frequency. This configuration is depicted in Figure 4.13(a) [121]. Conversely, in the time periods $[t_1, t_2]$ and $[t_3, t_4]$, the tank and load are decoupled. In this situation, the resonant inductor is in series with the magnetizing inductor, and the combined inductance of these two inductors resonates with the resonant capacitor, as shown in Figure 4.13(b) [121].

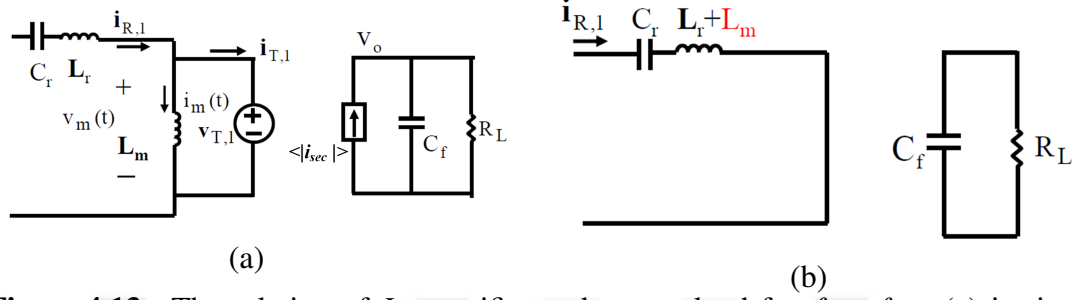


Figure 4.13 : The relation of L_m rectifier and output load for $f_s < f_0$. (a) in time periods $[t_0, t_1]$ and $[t_2, t_3]$, (b) In time periods $[t_1, t_2]$ and $[t_3, t_4]$.

The duration of the time intervals $[t_0, t_1]$ and $[t_2, t_3]$ corresponds to the resonant period T_0 , while the length of the intervals $[t_1, t_2]$ and $[t_3, t_4]$ is $T_s - T_0$, where T_s represents the total switching period. For modeling the entire switching cycle, the modulation model can be constructed by amalgamating the models depicted in Figure 4.13(a) and Figure 4.13(b). This combination is weighted by the respective time ratios of T_0/T_s and $(T_s - T_0)/T_s$. The resulting small signal modulation model, which incorporates these time-based proportions, is illustrated in Figure 4.14.

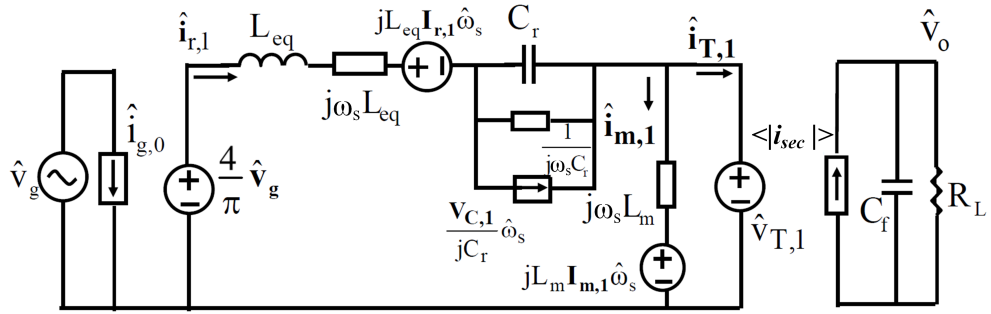


Figure 4.14 : Small signal model of LLC for $f_s < f_0$.

The equivalent resonant inductor, L_{eq} is related with resonant inductor and magnetizing inductor. The expression of L_{eq} is shown as equation (4.11):

$$L_{eq} = L_r + L_m \frac{\omega_o - \omega_s}{\omega_o} \quad (4.11)$$

Adopting a methodology akin to the previous case, the complex terms in the resonant tank model can be effectively eliminated by dividing it into sine and cosine components. This separation simplifies the analysis and is illustrated in Figure 4.15 [121].

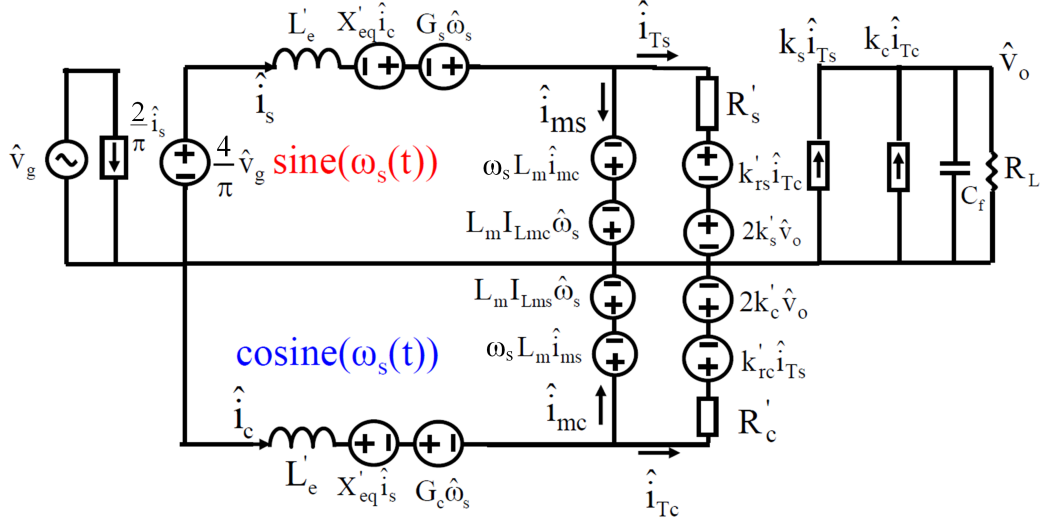


Figure 4.15 : Separated equivalent circuit model of LLC for $f_s < f_0$.

The expressions are shown in equation (4.12):

$$\begin{aligned} R'_s &= R_s \sin\left(\frac{\pi}{2}F\right), R'_c = R_c \sin\left(\frac{\pi}{2}F\right) \\ k_{rs}' &= k_{rc}' = k_{rs} \sin\left(\frac{\pi}{2}F\right) \\ k_s' &= k_s \sin\left(\frac{\pi}{2}F\right), k_c' = k_c \sin\left(\frac{\pi}{2}F\right) \end{aligned} \quad (4.12)$$

In a manner similar to the previous case, the superposition theorem is utilized to develop a non-coupled equivalent circuit model. This approach effectively disentangles the interactions within the circuit, leading to a more straightforward model. The resulting non-coupled equivalent circuit is presented in Figure 4.16.

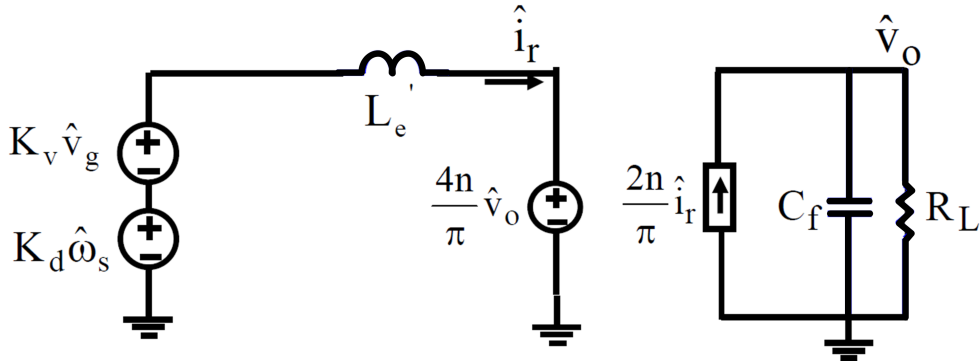


Figure 4.16 : Non-coupled equivalent circuit model of LLC for $f_s < f_0$.

In this case, there is no beat frequency dynamics and the circuit is especially second-order. The expressions of L_e' , K_v , and K_d are shown in equation (4.13) .

$$\begin{aligned}
L_e' &= L_r \left(1 + \frac{1}{F^2}\right) + L_m (1 - F), K_d = \frac{4V_g}{\pi} \frac{L_n}{\omega_o} (\alpha + \beta) \\
\alpha &= \frac{\left[\left(\frac{1}{F^2} - F^2 \right) \left(\frac{\pi^2}{8} Q_r L_n \right)^2 - \left(L_n + 1 - \frac{1}{F^2} \right) \left(\frac{2}{F^2} \right) \right] \frac{1}{F} \frac{1}{\sin\left(\frac{\pi}{2} F\right)}}{\left[\sqrt{\left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\left(\frac{1}{F} - F \right) \frac{\pi^2}{8} Q_r L_n \right)^2} \right]^3} \\
&\quad \left(-\frac{\pi}{2} \frac{\cos\left(\frac{\pi}{2} F\right)}{\sin^2\left(\frac{\pi}{2} F\right)} \right) \\
\beta &= \frac{\left(-\frac{\pi}{2} \frac{\cos\left(\frac{\pi}{2} F\right)}{\sin^2\left(\frac{\pi}{2} F\right)} \right)}{\sqrt{\left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\left(\frac{1}{F} - F \right) \frac{\pi^2}{8} Q_r L_n \right)^2}} \\
K_v &= \frac{2}{\pi} \frac{L_n}{\sin(F\pi/2)} \frac{1}{\sqrt{\left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\left(\frac{1}{F} - F \right) \frac{\pi^2}{8} Q_r L_n \frac{1}{\sin(F\pi/2)} \right)^2}}
\end{aligned} \tag{4.13}$$

4.1.3 Unified equivalent circuit model of LLC

The small-signal models shown in Figures 4.9 and 4.14 are combined to make a model that works for both when f_s is bigger than or equal to f_0 and when f_s is smaller than f_0 . This combined model is in Figure 4.17. In this model, L_e , R_e , and C_e are parts that show the beat frequency behavior. K_v , G_v , K_d , and G_d are special values that depend on the magnetizing inductance. The formulas for these parts are in equations (4.14), (4.15), (4.16) and (4.17).

$$G_d = \frac{4V_g}{\pi} \frac{L_n}{\omega_o} \frac{1}{R_e} U_d \tag{4.14}$$

$$U_d = \frac{\frac{1}{F} \left(\frac{1}{F^2} - F^2 \right) \left(\frac{\pi^2}{8} Q_r L_n \right)^2 - \left(L_n + 1 - \frac{1}{F^2} \right) \left(\frac{2}{F^2} \right)}{\left[\sqrt{\left(L_n + 1 - \frac{1}{F^2} \right)^2 + \left(\left(\frac{1}{F} - F \right) \frac{\pi^2}{8} Q_r L_n \right)^2} \right]^3} + \frac{2}{L_n^2} \tag{4.15}$$

$$K_v = \frac{8}{\pi^2} \frac{V_g L_n F}{R_{eq}} Y_V$$

$$Y_V = \frac{L_n + 1 - \frac{1}{F^2}}{\sqrt{\left(L_n + 1 - \frac{1}{F^2}\right)^2 + \left(\left(\frac{1}{F} - F\right) \frac{\pi^2}{8} Q L_n\right)^2}} \quad (4.16)$$

$$K_d = -\frac{8V_g}{\pi} \frac{1}{\omega_o L_n}, G_v = \frac{2}{\pi} \frac{X_{eq}}{\sqrt{X_{eq}^2 + R_{eq}^2}}$$

$$L_e = \begin{cases} \left(1 + \frac{1}{F^2}\right) L_r & \text{for } F \geq 1 \\ \left(1 + \frac{1}{F^2}\right) L_r + (1 - F) L_m & \text{for } F < 1 \end{cases}$$

$$R_e = \begin{cases} \frac{L_e |X_{eq}| |\omega_s - \omega_o|}{R_{eq}} & \text{for } F \geq 1 \\ 0 & \text{for } F \leq 1 \end{cases} \quad (4.17)$$

$$C_e = \frac{1}{L_e (\omega_s - \omega_o)^2}$$

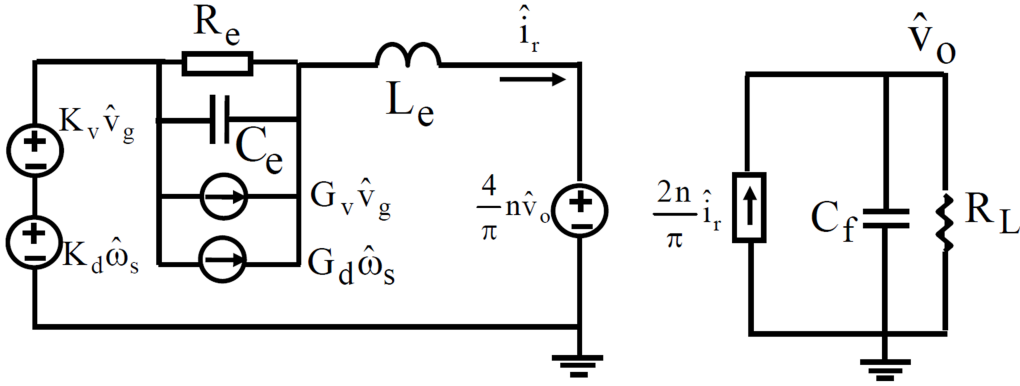


Figure 4.17 : Unified equivalent circuit model of LLC.

The equation of the unified equivalent circuit is expressed in terms of the perturbations in output voltage, input voltage, resonant current, output current and switching frequency [9, 121]. In order to find the expression for i_r , superposition is applied to the equivalent circuits as in equations (4.18), (4.19) and (4.20).

$$\begin{aligned} -K_d \hat{\omega}_s + I_1 \left(\frac{R_e}{sC_e R_e + 1} + sL_e \right) &= 0 \\ -K_v \hat{v}_g + I_2 \left(\frac{R_e}{sC_e R_e + 1} + sL_e \right) &= 0 \\ I_3 \left(\frac{R_e}{sC_e R_e + 1} + sL_e \right) + \frac{4n}{\pi} \hat{v}_o &= 0 \end{aligned} \quad (4.18)$$

$$\begin{aligned}
G_v \hat{v}_g \frac{\left(\frac{R_e}{sC_e R_e + 1} \right)}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} &= I_4 \\
G_d \hat{w}_s \frac{\left(\frac{R_e}{sC_e R_e + 1} \right)}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} &= I_5 \\
\hat{i}_r &= I_1 + I_2 + I_3 + I_4 + I_5
\end{aligned} \tag{4.19}$$

$$\begin{aligned}
\hat{i}_r &= \frac{(K_d \hat{w}_s + K_v \hat{v}_g)}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} - \frac{4}{\pi} \frac{n \hat{v}_o}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} \\
&+ \frac{\left(\frac{R_e}{sC_e R_e + 1} \right)}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} (G_v \hat{v}_g + G_d \hat{w}_s)
\end{aligned} \tag{4.20}$$

After finding the small signal expression of i_r , the relationships of output voltage and current can be found by

$$\begin{aligned}
\hat{v}_o &= \frac{2n}{\pi} \hat{i}_{Lr} \frac{R_L}{sC_f R_L + 1} \\
\hat{b}_r &= \hat{i}_o \left(\frac{1 + sC_f}{R_L} \frac{\pi}{2n} \right)
\end{aligned} \tag{4.21}$$

A simplified notation for the output current can be written as equation (4.22) .

$$\begin{aligned}
\hat{i}_o(s) &\approx G_{vg}(s) \hat{v}_g(s) + G_{vs}(s) \hat{w}_s(s) + G_{vo}(s) \hat{v}_o(s) \\
G_{vg} &= \left. \frac{\hat{i}_o(s)}{\hat{v}_g(s)} \right|_{\hat{v}_o=0, \hat{w}_s=0} \\
G_{vs} &= \left. \frac{\hat{i}_o(s)}{\hat{w}_s(s)} \right|_{\hat{v}_g=0, \hat{v}_o=0} \\
G_{vo} &= \left. \frac{\hat{i}_o(s)}{\hat{v}_o(s)} \right|_{\hat{v}_g=0, \hat{w}_s=0}
\end{aligned} \tag{4.22}$$

The control block diagram of the LLC resonant converter with a CC and CV regions are shown in Figure 4.18. The transfer function of $\frac{\hat{v}_o}{\hat{w}_s}$ for the CV region and the transfer functions of $\frac{\hat{i}_o}{\hat{w}_s}$ for the CC region are given in equation (4.23).

$$\begin{aligned}
\frac{\hat{v}_o}{\hat{w}_s} &= \frac{\left(K_d + \left(\frac{R_e}{sC_e R_e + 1} \right) \cdot G_d \right)}{\left(\frac{sC_f R_L + 1}{R_L} \cdot \frac{\pi}{2n} + \frac{4}{\pi} \cdot \frac{n}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} \right)} \\
\frac{\hat{i}_o}{\hat{w}_s} &= \frac{\left(K_d + \left(\frac{R_e}{sC_e R_e + 1} \right) \cdot G_d \right)}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right)} \\
&= \frac{\left(K_d + \left(\frac{R_e}{sC_e R_e + 1} \right) \cdot G_d \right)}{\left(1 + \frac{8R_L n^2}{\left(\frac{R_e}{sC_e R_e + 1} + sL_e \right) \cdot (sC_f R_L + 1) \cdot \pi^2} \right) \cdot \left(\frac{1+sC_f}{R_L} \cdot \frac{\pi}{2n} \right)}
\end{aligned} \tag{4.23}$$

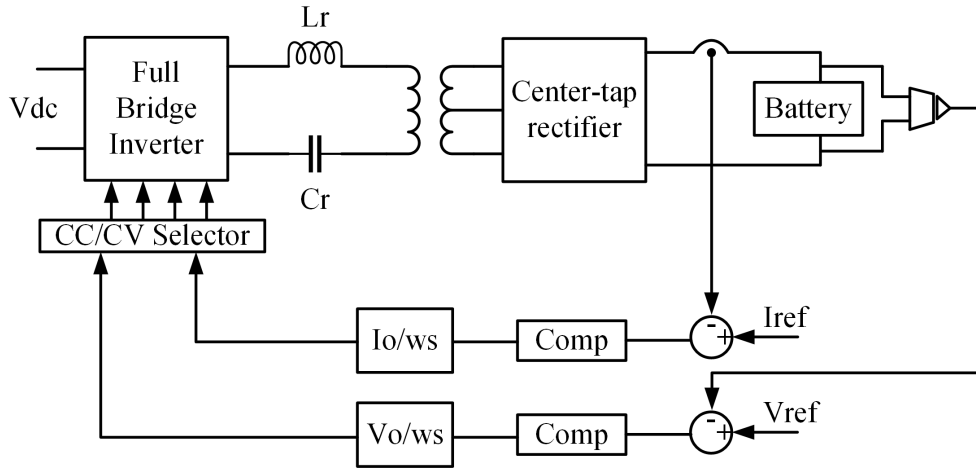


Figure 4.18 : Control block diagram of LLC for each region.

4.2 Compensator Design and Results

For a battery charging application with specifications of 370V-430V/36V-54V 3700W LLC converter with resonance frequency of 400 kHz, the magnetization inductance, resonant inductance, and resonant capacitor of 34.7 μH , 8.68 μH , and 18.25 nF, respectively, are considered in this study.

In order for the control loop system to operate at CC operating points effectively, the system is designed according to the Region 4 condition 400V/48V 3288W with 405kHz switching frequency. For the CV region, the control system is designed according to Region 6 400V/54V 3700W with 325kHz switching frequency.

The obtained transfer functions for the CC and CV regions are given in equation (4.24) for Region 4 and Region 6 respectively. Figure 4.19(a) and Figure 4.19(b) show the open loop Bode plots for every operating region.

$$\begin{aligned}\frac{\hat{v}_o(s)}{\widehat{\omega}_s(s)} &= \frac{-(9.893 \times 10^{11})}{(8.153)s^3 + (1.729 \times 10^7)s^2 + (1.072 \times 10^{11})s + (1.045 \times 10^{17})} \\ \frac{\hat{i}_o(s)}{\widehat{\omega}_s(s)} &= \frac{-(1.413 \times 10^{12})}{(8.153)s^3 + (1.729 \times 10^7)s^2 + (1.072 \times 10^{11})s + (1.045 \times 10^{17})}\end{aligned}\quad (4.24)$$

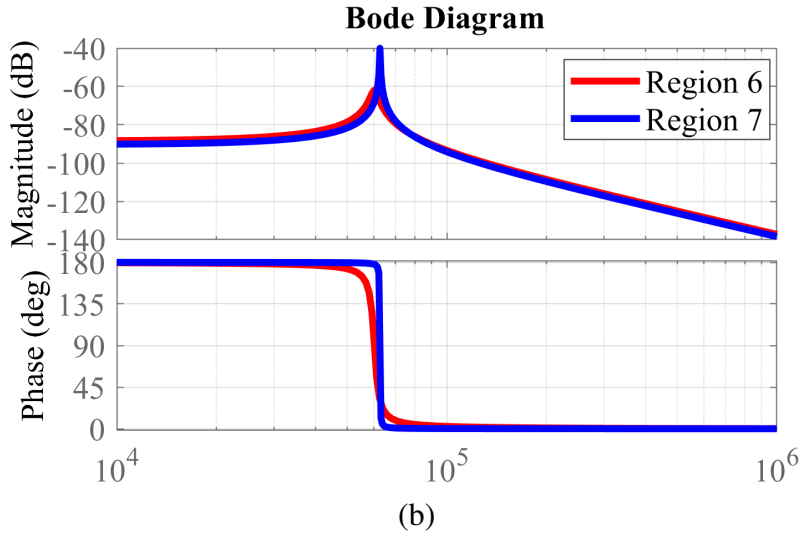
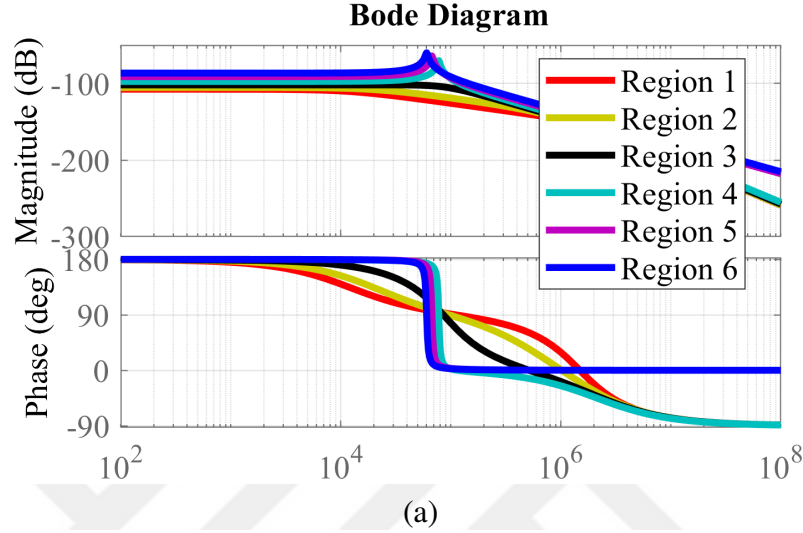


Figure 4.19 : Bode diagram of transfer functions (a) $\hat{i}_o/\widehat{\omega}_s$, (b) $\hat{v}_o/\widehat{\omega}_s$.

By using the obtained transfer functions, the closed-loop equivalent transfer function of the system is obtained for CC and CV region, and the transfer functions of the designed integral compensator are $I_{i_o} = -28190666$ and $I_{v_o} = -20750500$ (48ns) respectively. Several simulations are performed by setting references with different frequencies at the control input, measuring the system response and calculating its magnitude and phase by means of open PID Tuner in MATLAB environment. The control loop composed of only an integral (I) compensator. The reason why the integral compensator gain

is achieved so large is the amount of switching frequency. The bode diagrams of the closed loop transfer functions are shown in Figure 4.20 [123].

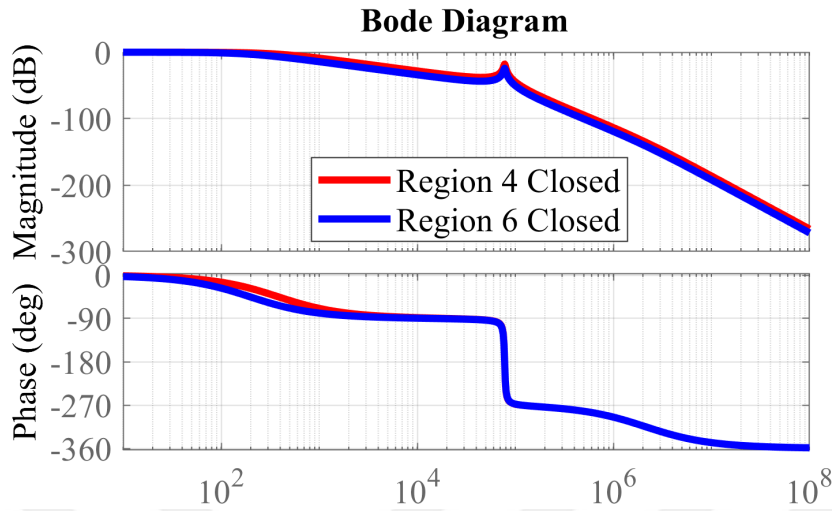
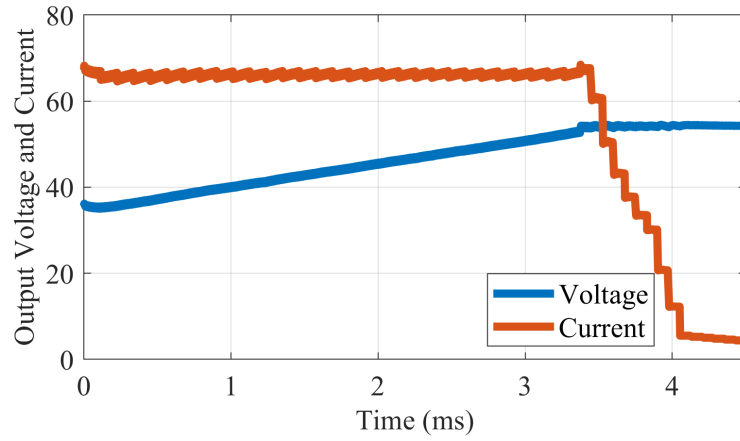


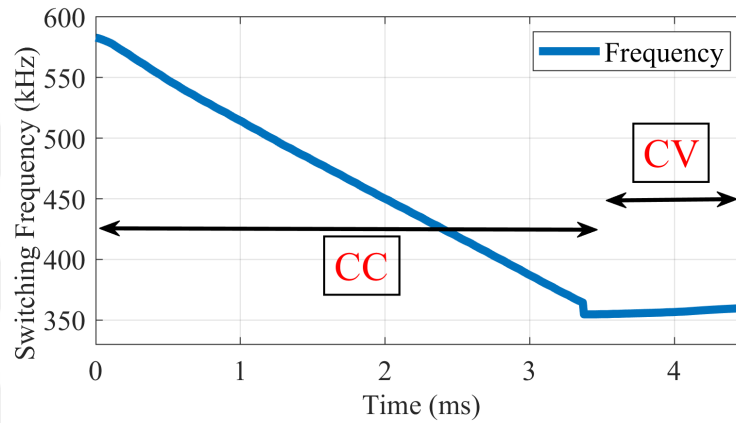
Figure 4.20 : Bode diagram of closed loop Region 4 and Region 6.

Figure 4.21(a) shows the response of the CC and CV control loops according to the given load profile. The whole charging process has been shortened to 5ms to save the simulation time. In the CC region, the closed loop current controller successfully regulates the current at 68A. In the CV region, the closed loop CV controller is activated and battery voltage is regulated at 54V, while the current slowly decreases as battery is being charged. The variation of the control variable switching frequency with respect to the varying load profile is shown in Figure 4.21(b). The system starts with high switching frequency and decreases as the battery is being charged [123].

The control structure designed according to certain operating points should be able to perform well and track the reference output voltages and currents under different operating points and instantaneous loading conditions. Figure 4.22(a) and 4.22(b) shows the response of the CC control loop for 6 different loading situations in favor of the output voltage and currents. These curves represent the worst loading conditions for each operating point where 68A reference is provided to the controller from a reference of 0 A. As it can be seen from Figure 4.22(c), the controller ensures a stable system by adjusting the switching frequency to below or above the resonance frequency in different load conditions [123]. Figure 4.23(a) shows that the designed control structure ensures regulated output voltage. Likewise, the control variable switching frequency variation under different load conditions in CV region is shown in Figure 4.23(b).



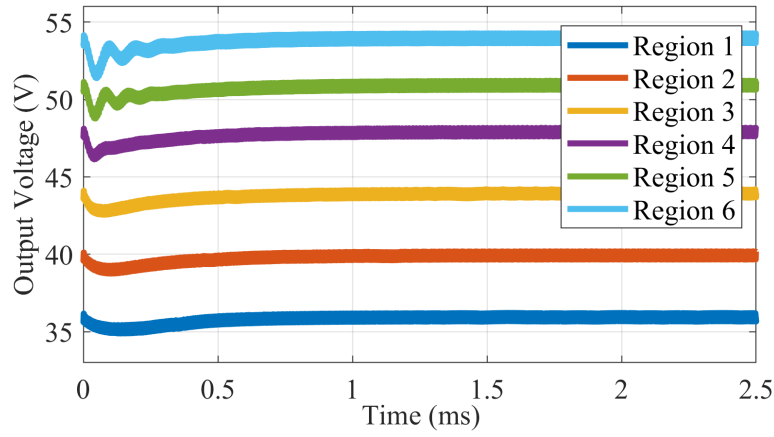
(a)



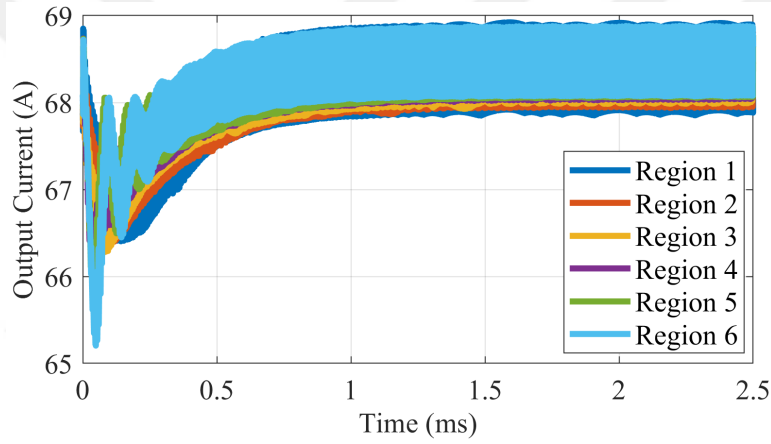
(b)

Figure 4.21 : Battery charger results of (a) output voltage and current (b) switching frequency variation.

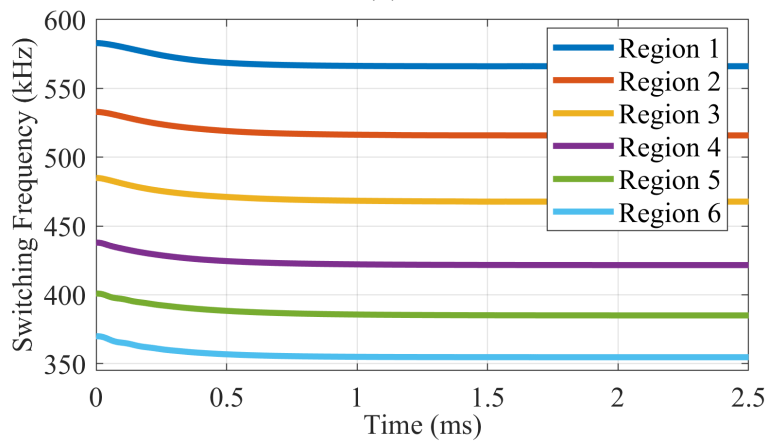
In this study, a CC and CV control scheme for a LLC resonant converter was presented in detail for battery charging applications. The small-signal model of the LLC converter is derived for both above and below resonance condition in order to obtain more accurate dynamic model. A simple third order equivalent circuit model of the LLC resonant converter was derived and the beat frequency dynamics and small signal behavior of the converter were very well predicted with this circuit model. The designed compensators for CC and CV operations at output voltage of 48V and 54V, respectively, have been simulated at different operating points. It has been shown that the output current and voltage references have been successfully tracked for whole load profiles.



(a)

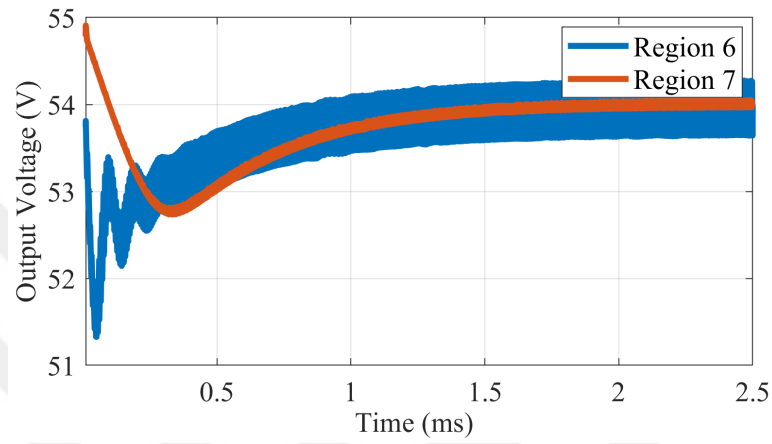


(b)

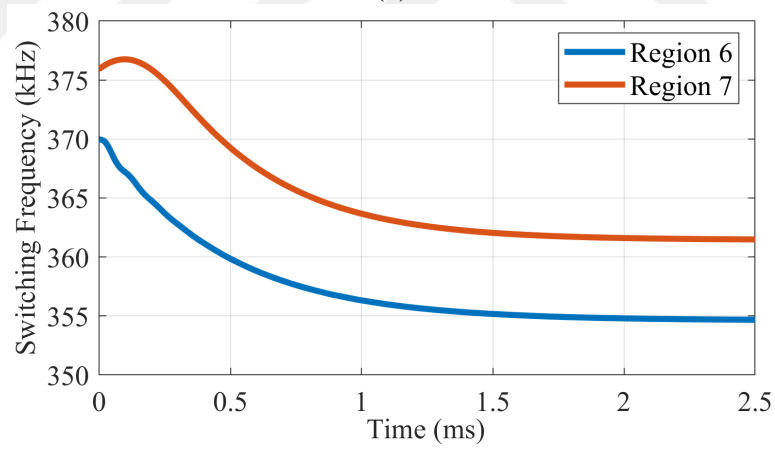


(c)

Figure 4.22 : Different load condition results in CC region (a) output voltage, (b) output current, (c) switching frequency.



(a)



(b)

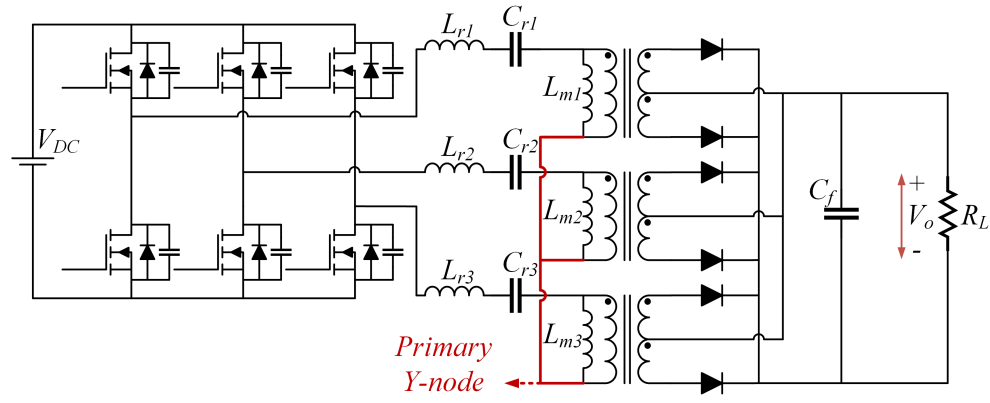
Figure 4.23 : Different load condition results in CV region (a) output voltage, (b) switching frequency variation.



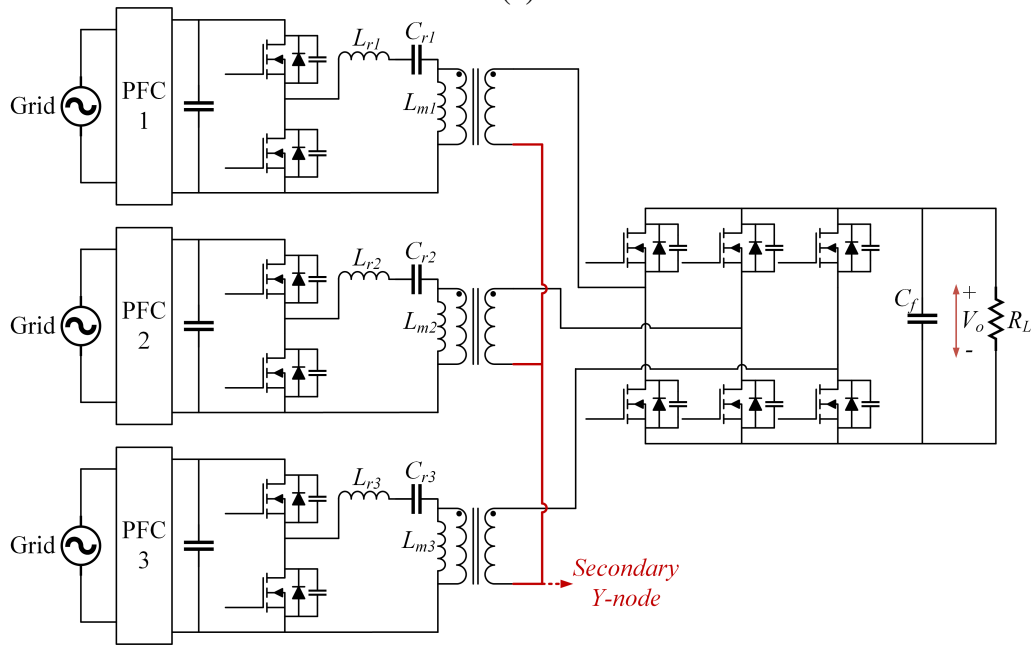
5. ANALYTICAL MODEL OF THREE-PHASE LLC RESONANT CONVERTER

The three-phase DC-DC resonant converter is beneficial for high power applications because it is easier on its components than a similar power single-phase converter [124]. There are several types of three-phase resonant converters. One type, the three-phase interleaved LLC converter, links the primary sides to a common Wye (Y)-node as shown in Figure 5.1(a) [44]. This setup allows primary switches, with a 120° phase difference, to share current evenly across phases, despite any variations in each phase's resonant tanks. Another type connects the secondary sides to a shared Y-node, as shown in Figure 5.1(b) [125]. This Y-node's average voltage is half the output voltage, similar to a voltage doubler's effect on the secondary side. Both layouts in Figure 5.1 allow the phases to interact via voltage changes at either the primary or secondary Y-nodes [125, 125]. This study explores and models combinations of primary Y, Delta (Δ), and Zigzag (Z) with secondary center-tapped (C), Y, Δ and Z configurations as illustrated in Figure 5.2.

Using three individual inductors and transformers in a three-phase LLC converter can significantly increase size and cost. Therefore, integrating magnetic components is crucial for the converter's application [126]. By using wide band-gap devices, including GaN and SiC, the switching frequency in power converters can reach up to several hundred kHz or MHz. This higher frequency reduces the voltage-second requirement across the transformer. With a consistent magnetic flux in the core, fewer turns are needed, allowing for the use of PCBs as transformer windings. The high switching frequency also means a smaller required inductor value, facilitating easier magnetic integration [124].



(a)



(b)

Figure 5.1 : Three-phase LLC converter topologies (a) Y-primary, (b) Y-secondary.

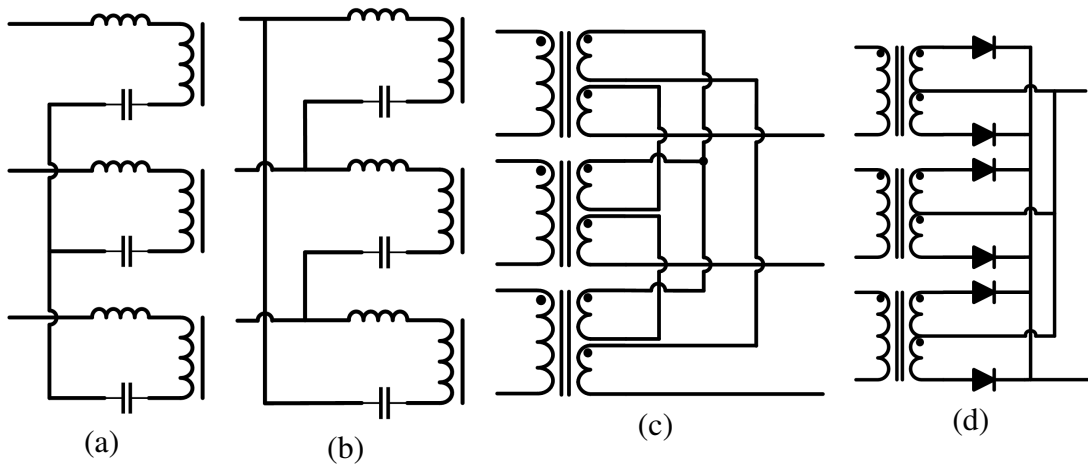


Figure 5.2 : Winding connections (a) Y-connection, (b) Δ -connection and (c) Zigzag-connection (d) secondary center-tapped.

5.1 First Harmonic Approximation of Different Topologies

The most common approach for designing and analyzing the resonance tank features of an LLC converter is the FHA. FHA assumes that the resonant current flows in a pure sinusoidal pattern. This method is used for all types of connection groups.

Step 1 involves determining the v_{tin} based on the V_{DC} . Figure 5.3 displays the voltage waveforms applied to the resonant tank's primary side under various primary side connections. To expand v_{tin} , the Fourier Series is used, and the outcomes of this expansion are presented in Table 5.1.

Step 2 involves calculating the v_{tout} in relation to the V_o . Since the tank only responds to the fundamental components of the applied waveforms, harmonics in v_{tout} can be ignored. Figure 5.4 shows the voltage waveforms applied to the secondary side of the resonant tank for various secondary side connections. v_{tout} is expanded for all secondary connection schemas and the results are summarized in Table 5.2.

Table 5.1 : The fundamental component of v_{tin} in terms of V_{DC} .

Topology	Expression
$\Delta - All$	$v_{tin} = \frac{2\sqrt{3}}{\pi} V_{DC}$
$Y - All$	$v_{tin} = \frac{2}{\pi} V_{DC}$
$Z - All$	$v_{tin} = \frac{2}{\pi} V_{DC}$

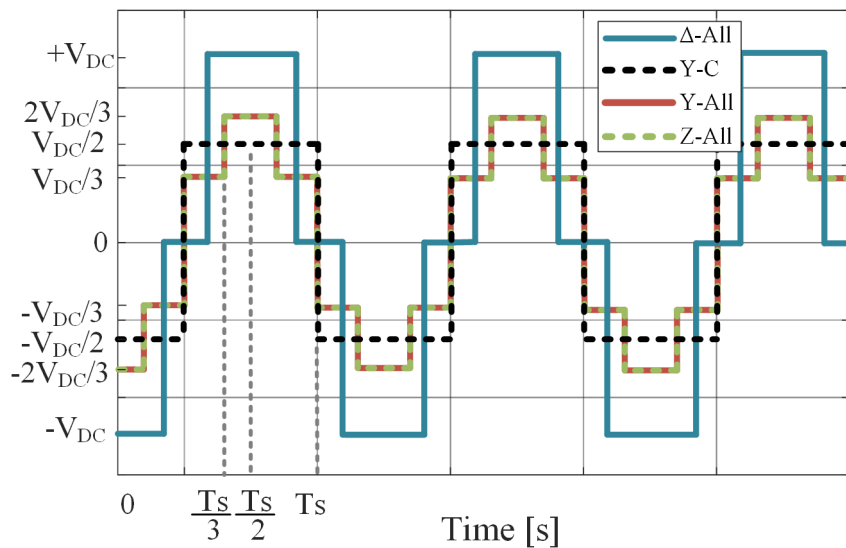


Figure 5.3 : The voltage waveforms applied to the resonant tank's primary side.

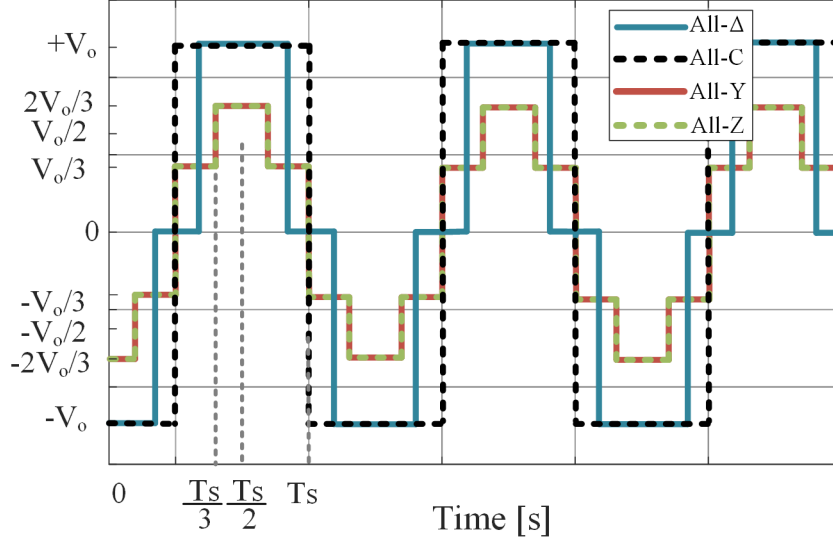


Figure 5.4 : The voltage waveforms applied to the resonant tank's secondary side.

Table 5.2 : The fundamental component of v_{tout} in terms of V_o .

Topology	Expression
<i>All</i> - Δ	$v_{tout} = \frac{2\sqrt{3}}{\pi} V_0$
<i>All</i> - Y	$v_{tout} = \frac{2}{\pi} V_0$
<i>All</i> - Z	$v_{tout} = \frac{2}{\pi} V_0$
<i>All</i> - C	$v_{tout} = \frac{4}{\pi} V_0$

Step 3 is about determining the I_{out} in terms of the i_{sec} . The average value of the rectified resonant tank output current, denoted as $\langle |i_{sec}| \rangle$, must match the steady-state load current I_{out} . The currents when secondary windings are connected in Δ and Y configurations are shown in Figure 5.5(a) and Figure 5.5(b), respectively. The results for all secondary connection types are provided Table 5.3.

Table 5.3 : The average value of the rectified resonant tank output current.

Topology	Expression
<i>All</i> - C	$I = \frac{3}{\pi} I_R$
<i>All</i> - D	$I = \frac{3\sqrt{3}}{\pi} I_R$
<i>All</i> - Y	$I = \frac{3}{\pi} I_R$
<i>All</i> - Z	$I = \frac{3}{\pi} I_R$

Step 4 is the calculation of the AC equivalent load resistance. The fundamental components of current and voltage are sinusoidal and in phase, indicating that the rectifier acts as a resistive load to the tank network. The AC equivalent load resistance,

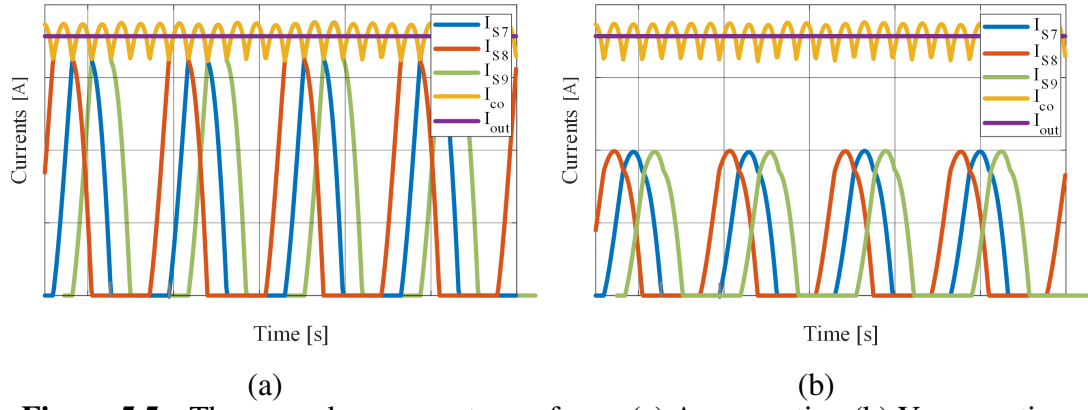


Figure 5.5 : The secondary current waveforms (a) Δ -connection (b) Y-connection.

R_e , can be calculated using equation (5.1). The results for all secondary connection types are again compiled in Table 5.4.

Table 5.4 : The AC equivalent load resistance expressions.

Topology	Expression
$Y - C$	$R_e = \frac{24}{\pi^2} R_L$
$Y - Y$	$R_e = \frac{6}{\pi^2} R_L$
$Y - Z$	$R_e = \frac{6}{\pi^2} R_L$
$Z - C$	$R_e = \frac{24}{\pi^2} R_L$
$Z - Y$	$R_e = \frac{6}{\pi^2} R_L$
$Z - Z$	$R_e = \frac{6}{\pi^2} R_L$

$$R_e = \frac{v_{\text{tout}}}{i_{\text{sec}}} \quad (5.1)$$

Step 5 involves deriving the H_s . $H(s)$, also known as the input-to-output voltage gain, is a mathematical relationship between input and output voltages. It incorporates several factors: L_m , L_r , C_r , M and switching frequency range. These factors are detailed in the equation presented in Chapter 2. Since the resonant tank type remains the same across all connection groups, the transfer function is identical to that of a single-phase system.

Step 6 is about deriving the M . The voltage conversion ratio of the resonant converter is defined as the magnitude of the product of the tank network transfer function as seen in equation (5.2). This ratio is influenced by a coefficient C , which varies depending on the types of connections on the primary and secondary sides. The C coefficients for various combinations of primary and secondary connections are listed Table 5.5.

$$M = \frac{V_o}{V_{DC}} = \frac{V_o}{I_{out}} \frac{I_{out}}{i_{sec}} \frac{i_{sec}}{v_{tout}} \frac{v_{tout}}{v_{tin}} \frac{v_{tin}}{V_{DC}} = C(H(s)) \quad (5.2)$$

Table 5.5 : The C coefficients for various combinations of primary and secondary connections.

Topology	C
$Y - C$	0.5
$Y - Y$	1
$Y - Z$	1
$Z - C$	0.5
$Z - Y$	1
$Z - Z$	1

5.2 Frequency Dependent Model of Three-Phase LLC Converter with Different Connection

For precise magnetic design, it's important to know the RMS and peak values of the resonance current, the peak value of the magnetization current, and the RMS value of the secondary-side currents. The operating currents of an LLC in a 48V LEV charger vary according to the state of the battery, which can be "below", "at", or "above" the resonance operating conditions. Thus, all current equations should be expressed in relation to the switching frequency. The methodology for a $\Delta - \Delta$ connection is outlined step by step.

Step 1 is determine the i_{Lr} in terms of the i_{sec} . To calculate the I_{Rm} in relation to the I_{out} , it's first necessary to derive mathematical expressions for the currents flowing through the MOSFETs Q_7 and Q_{11} of the same phase, as shown in equations (5.3) and (5.4). Figure 5.6 displays the current waveforms of the MOSFETs Q_7 and Q_{11} , i_{co} , and I_{out} .

$$i_{Q_7}(t) = \begin{cases} nI_{Rm} \sin(\omega_0 t - \varphi_{tout} + \frac{\pi}{6}) & 0 \leq t < \frac{T_0}{2} \\ 0 & \frac{T_0}{2} \leq t < T_s \end{cases} \quad (5.3)$$

$$i_{Q_{11}}(t) = \begin{cases} 0 & 0 \leq t < \frac{T_s}{2} \\ nI_{Rm} \sin(\omega_0 t - \varphi_{tout} + \frac{\pi}{6}) & \frac{T_s}{2} \leq t < \frac{T_s + T_0}{2} \\ 0 & \frac{T_s + T_0}{2} \leq t < T_s \end{cases} \quad (5.4)$$

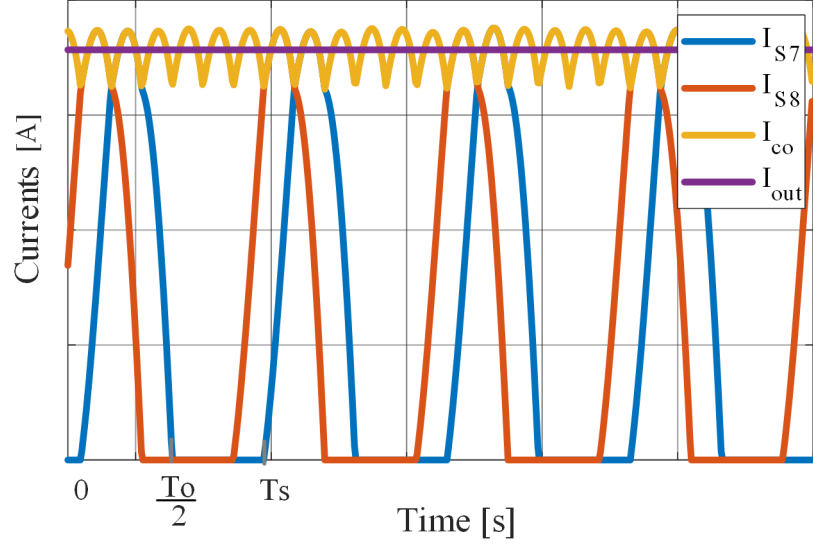


Figure 5.6 : The current waveforms of S_7 and S_7 switches, rectified output current and average output current.

L_m only contributes to the resonance when the i_{Lr} matches the magnetizing current. Commonly, the resonant frequency formed by the L_r , L_m and C_r is significantly lower than the frequency f_0 . Equation (5.5) describes both the i_{co} and I_{out} .

$$I_{out} = \langle i_{co} \rangle = \frac{2}{T_s/6} n I_{Rm} \int_0^{T_0/6} \sin\left(\omega_0 t + \frac{\pi}{6}\right) dt \quad (5.5)$$

Step 2 is specify the magnetizing current of the transformer. Figure 5.7 shows the magnetizing current and the transformer's secondary side voltage in a $\Delta - \Delta$ winding setup. The magnetizing current, based on the transformer's secondary voltage, is detailed in equation (5.6).

$$i_{Lm} = \left\{ \begin{array}{ll} \frac{nV_o}{L_m} t - \frac{nV_o}{L_m} \frac{T_s}{6} & 0 \leq t < \frac{T_s}{3} \\ \frac{nV_o}{L_m} \frac{T_s}{6} & \frac{T_s}{3} \leq t < \frac{T_s}{2} \\ -\frac{nV_o}{L_m} t + \frac{4nV_o}{L_m} \frac{T_s}{6} & \frac{T_s}{2} \leq t < \frac{5T_s}{6} \\ -\frac{nV_o}{L_m} \frac{T_s}{6} & \frac{5T_s}{6} \leq t < T_s \end{array} \right\} \quad (5.6)$$

Step 3 is define the resonant current. i_{Lr} is the sum of the referred secondary current of the transformer to the primary side, and the magnetizing current, which is expressed in equation (5.7) and (5.8).

$$i_{Lr} = i_{Lm} + \frac{1}{n} i_{sec} = i_{Lm} + \frac{1}{n} (i_{Q7} - i_{Q11}) \quad (5.7)$$

$$i_{Lr} = \begin{cases} \frac{nV_o}{L_m}t - \frac{nV_o}{L_m}\frac{T_s}{6} + I_{Rm} \sin\left(\omega_0 t + \frac{\pi}{6}\right) & 0 \leq t < \frac{T_s}{3} \\ \frac{nV_o}{L_m}\frac{T_s}{6} + I_{Rm} \sin\left(\omega_0 t + \frac{\pi}{6}\right) & \frac{T_s}{3} \leq t < \frac{T_o}{2} \\ \frac{nV_o}{L_m}\frac{T_s}{6} & \frac{T_o}{2} \leq t < \frac{T_s}{2} \\ -\frac{nV_o}{L_m}t + \frac{4nV_o}{L_m}\frac{T_s}{6} - I_{Rm} \sin\left(\omega_0 t + \frac{\pi}{6}\right) & \frac{T_s}{2} \leq t < \frac{5T_s}{6} \\ -\frac{nV_o}{L_m}\frac{T_s}{6} - I_{Rm} \sin\left(\omega_0 t + \frac{\pi}{6}\right) & \frac{5T_s}{6} \leq t < T_o \\ -\frac{nV_o}{L_m}\frac{T_s}{6} & T_o \leq t < T_s \end{cases} \quad (5.8)$$

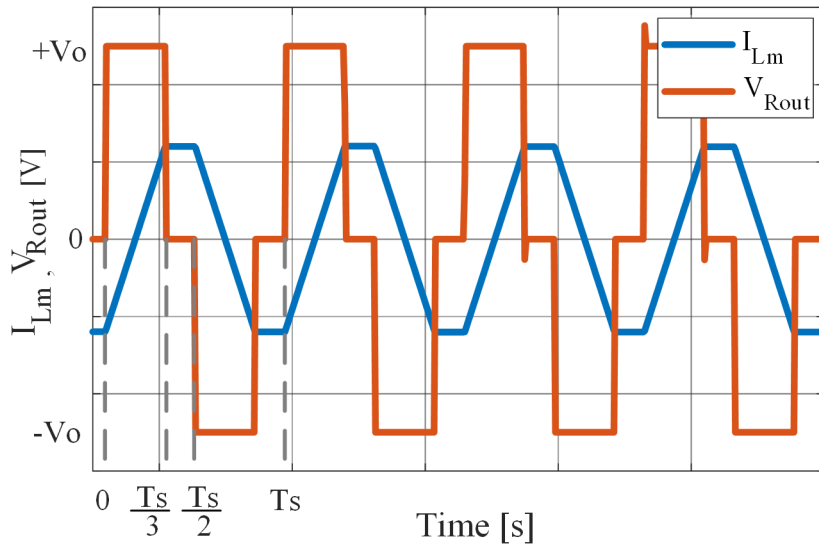


Figure 5.7 : The magnetization current and transformer secondary side voltage.

Step 4 is define RMS value of secondary current of transformer. The RMS value of the transformer secondary side current is derived as in equation (5.9).

$$i_{sec_{RMS}} = \sqrt{2}i_{Q7_{RMS}} \quad (5.9)$$

5.3 Optimization Based on the 3.7kW Three-Phase LLC Converter

A systematic design framework is provided for three-phase LLC converter with $\Delta - \Delta$ winding configuration that attempts to find the optimal resonance frequency and switching frequency range that satisfies the quality factors at corresponding voltage gains and minimizes a system level fitness function to obtain the optimum magnetic designs. Different than the previous studies, the algorithm provides newly derived current equations depending on the frequency and FHA circuit characteristics. For

each resonance frequency, satisfying L_r , L_m , and C_r values are calculated, and the resonant transformer and inductor are designed based on the analytical models for every feasible core in the core database. Finally, the performance of the optimized three-phase LLC converter with $\Delta - \Delta$ winding configuration, as shown in Figure 5.8, is compared with single-phase primary-series secondary-parallel connected 3-core configuration transformer to quantify the benefits obtained by using the proposed framework.

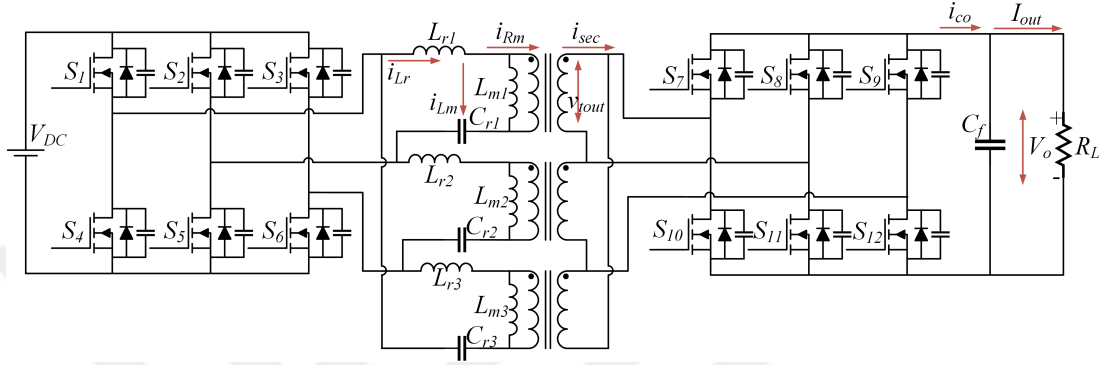


Figure 5.8 : Three phase LLC converter with $\Delta - \Delta$ winding configuration.

The proposed design algorithm is based on the principle of searching for the best transformer and inductor by minimizing the defined system level fitness function including loss, cost and volume in the given core database. The flowchart of the proposed framework is shown in Figure 5.9. More information on the air-gapped transformer design can be found in 2.

The transformer and resonant inductors of a 3700 W LLC with an input voltage of 370-430 V, and output voltage of 48-54 V is designed using the proposed algorithm as a case study. The derived current equations, that are expressed as a function of switching frequency, are compared with the simulation results in Table 5.6. The RMS resonant currents are calculated numerically as 9.797 A and 3.9279 A for single phase 3-core and three-phase with $\Delta - \Delta$ winding configuration, respectively. As seen from the Table 5.7, the derived equations and the simulation results are matching closely. The algorithm sweeps the resonance frequency between 300-500 kHz, while the minimum and maximum frequencies are internally swept through L_n and Q pairs. The weights of the fitness function are chosen as 60%, 20%, and 20% for volume, power loss, and cost, respectively. The design space for feasible $L_n - Q$ combinations at each resonance frequency are provided in Figure 5.10(a) and Figure ??(b) for single-phase

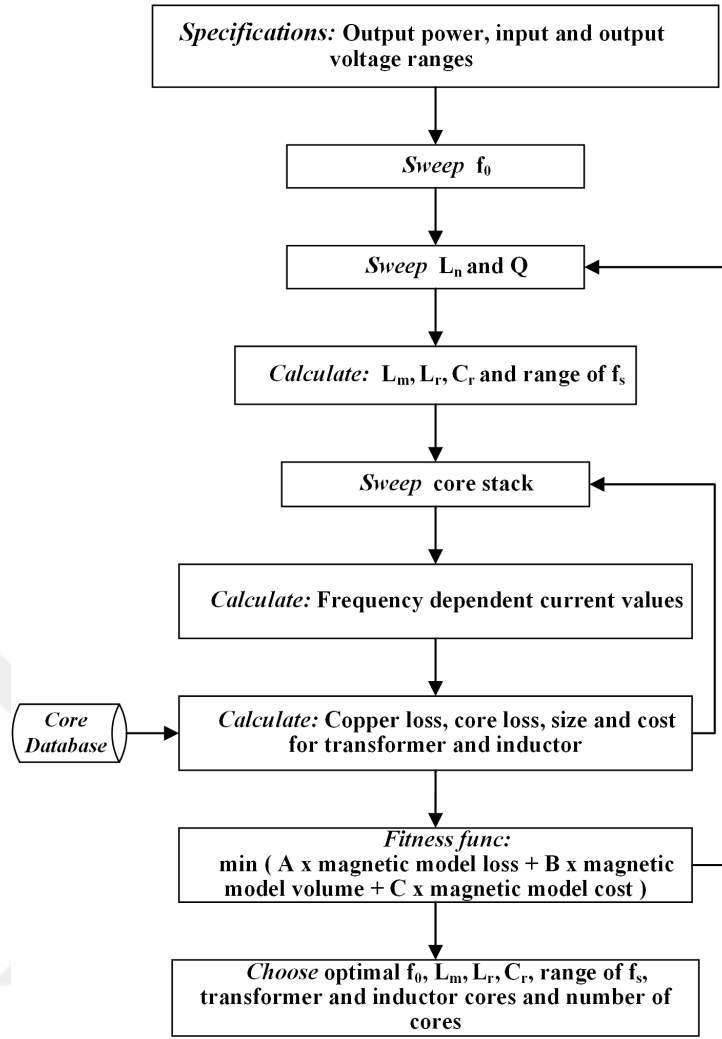
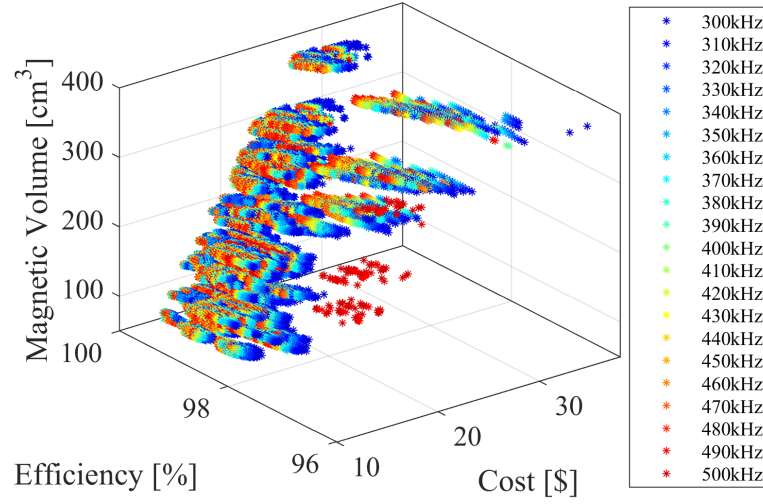


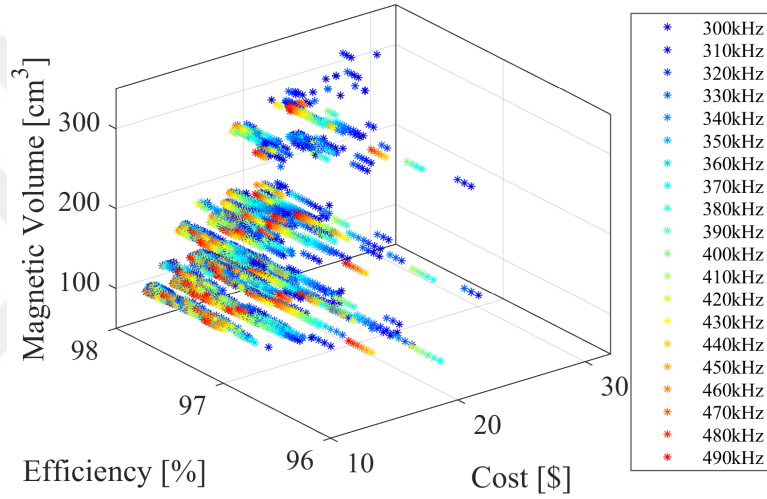
Figure 5.9 : Flowchart of the proposed design algorithm for three-phase LLC converter with $\Delta-\Delta$ winding configuration.

3-core and three-phase $\Delta-\Delta$ winding configurations, respectively. These design points show the total volume, cost and efficiency of each viable magnetic model for each L_n - Q combination at the corresponding resonance frequency. It must be noted that this design space also considers the multi-stack core proposed in [111]. As it can be seen from Figure 5.10, an increase in the resonance frequency decreases the total volume and cost of the magnetic components, however, increases the total power loss.

The optimum design parameters found by the proposed design algorithm are shown in Table 5.7 for single-phase 3-core and three-phase with $\Delta-\Delta$ winding configurations. A total of 29642 individual designs were acquired. Among this design space, the algorithm chooses the resonance frequency as 370 kHz according to the fitness function with assigned weights for both configurations. For detailed in-depth analysis, the volume, power loss, and cost of the transformer and the inductor are shown in the



(a)



(b)

Figure 5.10 : Magnetic design space; a) single-phase 3-core, b) three-phase with $\Delta - \Delta$ winding configuration.

radar chart given in Figure 5.11. As it can be seen from the results, the LLC converter with three-phase transformer with $\Delta - \Delta$ winding configuration reduces the magnetic power losses by 34.8% in comparison to the design with single-phase primary-series secondary-parallel 3-core transformer configuration at the expense of increased cost.

Three-phase LLC converter with $\Delta - \Delta$ winding configuration is a strong candidate for high power battery charging applications. Like in a single-phase LLC converter, the design starts off with the selection of the resonance frequency, which is usually determined based on the designer's experience, and a $L_n - Q$ pair that satisfies the required voltage conversion over a wide load range. Without fully modeling the magnetic components, it is not possible to optimize the LLC converter. Therefore, this paper proposed an optimized design framework to choose the resonance frequency

Table 5.6 : Derived resonant current equation values and simulation results.

	Single-phase with 3-core		Three-phase with $\Delta - \Delta$ winding	
Resonant	Numeric	Simulation	Numeric	Simulation
Current (A)	9.797	9.99	3.9279	4.01

Table 5.7 : Results of the design and optimization algorithm.

Symbol	Description	Values	
		Single-phase with 3-core	Three-phase $\Delta - \Delta$ winding
L_m	Magnetizing Ind.	$37.52\mu H$	$84.41\mu H$
L_r	Resonant Ind.	$9.38\mu H$	$21.1\mu H$
C_r	Resonant Cap.e	$19.73nF$	$8.77nF$
f_0	Resonant Freq.	$370kHz$	$370kHz$
Wire thickness	Primary	13AWG	17AWG

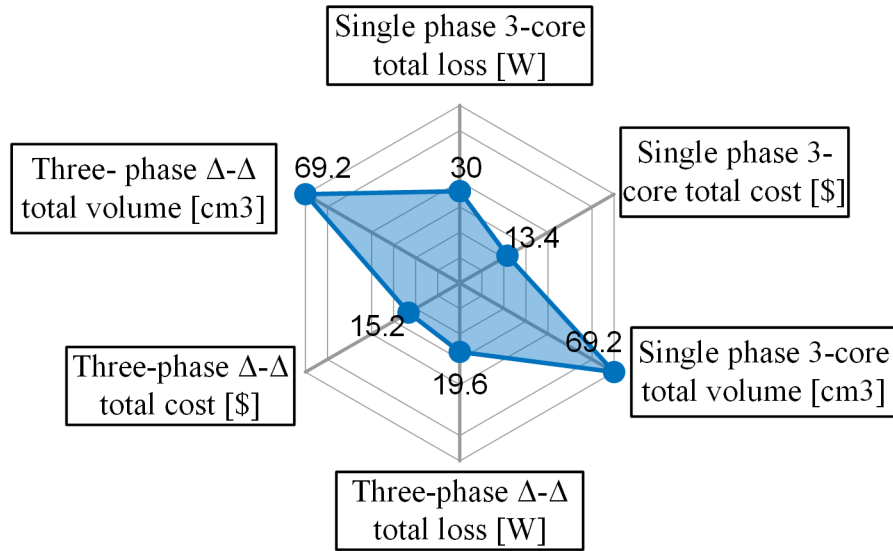


Figure 5.11 : Comparison of optimized LLC converters with single phase with 3-core and three-phase with $\Delta - \Delta$ winding configuration.

and $L_n - Q$ pair that minimizes the fitness function consisting of power loss, volume and cost of the three-phase LLC transformer and the resonant inductor. To do this, firstly the switching frequency dependent current quantities of three-phase LLC converter with $\Delta - \Delta$ winding configuration are derived for the first time in literature. The design aspects such as power losses, thermal limits, saturation flux density, etc. are calculated using these derived mathematical equations.

The optimization algorithm has been run to design a 3700W 48V LLC converter for both three-phase LLC converter with $\Delta - \Delta$ winding configuration and single-phase primary-series secondary-parallel connected 3-core LLC converter. The results suggest that the volume of these two different LLC topologies are very similar, while the power losses of three-phase LLC converter with $\Delta - \Delta$ winding configuration comes out 34.8% less at the expense of 13.88% increased cost.



6. CONCLUSION

This thesis consolidates multiple research efforts into a coherent narrative on the design optimization of LLC converters, with a focus on applications in LEV chargers. The thesis is structured to sequentially build upon each study's findings, culminating in a comprehensive approach to magnetic component design, system optimization, and the inclusion of parasitic elements in high-fidelity simulation models.

The significant findings can be summarized as follows:

1. **Optimized Magnetic Design:** The research developed a comprehensive magnetic design algorithm, optimizing the balance between copper and core losses. Key highlights include:
 - Unlike classical methods where a pre-determined B_{max} is selected, the operating flux density is swept up to 65% of the B_{sat} .
 - Distributed core structures and various winding types were considered, resulting in a design that reduces losses and volume.
 - A prototype confirmed the algorithm's effectiveness, showing considerable improvements in performance over conventional single-core designs.
2. **Systematic Design Framework:** A systematic approach was presented to determine the optimal resonance frequency and switching frequency limits for LLC converters, which emphasized:
 - The importance of considering the load profile of a LEV battery charger when sizing the LLC converter.
 - A design that achieved a peak efficiency of 98.2% at 60% load, surpassing the minimum targeted efficiency of 97%.
3. **Control Scheme Accuracy:** The thesis detailed a Constant Current (CC) and Constant Voltage (CV) control scheme for LLC converters, showcasing:
 - A derived third-order equivalent circuit model that accurately predicts the converter's dynamics.

- Successful tracking of output current and voltage references across various load profiles.
4. Three-phase LLC Converter Design: The research extended to three-phase LLC converters, finding that:
- A novel mathematical derivation was introduced for current quantities dependent on switching frequency.
 - Comparatively, the three-phase design with a $\Delta - \Delta$ winding configuration showed a significant reduction in power losses at a slightly increased cost.

All in all, this study contributes to a holistic design methodology that integrates magnetics design optimization, systematic framework development for converter sizing, accuracy in control schemes, and the advancement of three-phase converter designs. The thesis validates the hypothesis that a systematic design framework can enhance the performance and efficiency of LLC resonant converters for LEV charging, resulting in designs that are not only energy-efficient but also optimized for volume, cost, and thermal performance.

The impact of the design and simulation methods proposed in this thesis on practical applications has significant implications for the advancement and implementation of LEV charging systems. These contributions are of value to both industry and academia and have the potential to guide future design strategies.

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A HOLISTIC DESIGN OPTIMIZATION METHOD FOR LLC
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