Alternative Structures for Integrated Electromagnetic Passives

by

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Abstract

The demand for high power density keeps driving the development of electromagnetic integration technologies in the field of power electronics. Based on planar homogeneous integrated structures, the mechanism of the electromagnetic integration of passives has been investigated with distributed-parameter models. High order modeling of integrated passives has been developed to investigate the electromagnetic performance. The design algorithm combining electromagnetic design and loss models has been developed to optimize and evaluate the spiral winding structure. High power density of 480 W/in$^3$ has been obtained on the prototype.

Due to the structural limitation, the currently applied planar spiral winding structure does not sufficiently utilize the space, and the structure is mechanically vulnerable. The improvement on structures is necessary for further application of integrated passives. The goal of this research is to investigate and evaluate alternative structures for high-power-density integrated passives. The research covers electromagnetic modeling, constructional study, design algorithm, loss modeling, thermal management and implementation technology.

The symmetric single layer structure and the stacked structure are proposed to overcome the disadvantages of the currently applied planar spiral winding structure. Because of the potential of high power density and low power loss, the stacked structure is selected for further research. The structural characteristics and the processing technologies are addressed.

By taking an integrated LLCT module as the study case, the general design algorithm is developed to find out a set of feasible designs. The obtained design maps are used to evaluate the constraints from spatial, materials and processing technologies for the stacked structure.
Based on the assumption of one-dimensional magnetic filed on the cross-section and linear current distribution along the longitudinal direction of the stacked structure, the electromagnetic field distribution is analyzed and the loss modeling is made. The experimental method is proposed to measure the loss and to verify the calculation.

The power loss in the module leads to thermal issues, which limit the processed power of power electronics modules and thus limit the power density. To further improve the power handling ability of the module, the thermal management is made based on loss estimation. The heat extraction technology is developed to improve the heat removal ability and further improve the power density of integrated passives.

The experimental results verify the power density improvement from the proposed stacked structure and the applied heat extraction technology. The power density of 1147 W/in³ (70 W/cm³) is achieved in the implemented LLCT module with the efficiency of 97.8% at output power of 1008W.
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Chapter 1: Introduction

1.1. Current trends in power electronics

In the past two decades, with the increasing amount of electric and electronic machines and devices, the field of power electronics has become increasingly important. High performance power supplies are required in all variety of consumer electronics from personal computers, shavers to tanks and space shuttles. Demands for higher efficiencies, lower costs and ever-decreasing volume and profile have driven the development of power electronics converters and components. Efforts on increasing operation frequency [1], improving topologies [2], planarization [3] and modularization [4] have been made to satisfy these demands. However, the reduced space, and increased frequency and power density have in turn lead to the concerns of parasitic parameters [5], EMI noise [6] and thermal issues [7]. To solve the induced problems, the future of the power electronics will require more revolutionary ideas. The electromagnetic integration of the power electronics components has been proposed as the solution.

In conventional power electronics, the converter is constructed from discrete components and packages that are interconnected though printed circuit boards and/or wires. Each component or package is created to be a mechanically robust and thermally reliable block with sole electric function, such as an inductor, transformer or switch, and so on. Actually, no real component has single electromagnetic function as the ideal component described in circuit diagram. Every real component has inherent parasitic capacitance and inductance inside the package. Parasitic parameters also exist outside the component package due to the interconnections and the layouts. When the operation frequencies increase, the circuit operation become more and more sensitive to the parasitic effects. The parasitic inductance and capacitance induces voltage and current stresses on active components and leads to additional losses. In high frequency applications, such as, EMI filter and LLC resonant converter, the induced parasitic parameters will change the overall values of passive parameters and impact the circuit performance.

Integrated power electronics modules are proposed to replace the discrete conventional components to solve the problems induced by parasitic parameters and to improve power
density. By functionally integrating discrete power electronics components into modules, the equivalent series inductance (ESL) is reduced for shortened interconnections, while the equivalent parallel capacitance (EPC) is controllable with the specific dimensions and properties of applied materials. The application of integrated power electronics modules has become a more and more successful solution to achieve good electromagnetic performance, high power density and low cost of power converter.

1.2. Integrated power electronics modules

According to different electrical functions and processing technologies, the integrated power electronics modules can be chiefly classified into two categories, active integrated power electronics modules (IPEMs) and passive IPEMs. Active power electronics modules have been successfully developed as standardized commercial products [8].

In recent years, electromagnetic integration of passive components is becoming more important, because these energy storage components contribute to most of the total volume. By integrating the electromagnetic functions of inductor, capacitor and transformer into a functional module, the number of the passive components reduces and the overall passives volume decreases. Different from the ideal models of passive components in circuit theory, integrated passives are modeled and designed with the electromagnetic energy propagation theory, which has been proven to be more applicable and accurate in high frequency range. The design of integrated passives concerns how the stored energy influences the electromagnetic performance by studying the electromagnetic field distribution and energy propagation. By treating the “parasitic parameter” as part of the design variables, integrated passives have good electromagnetic performance, and utilize the space efficiently. Integrated power passive modules have been studied for several years. Prototypes have been developed and adopted in various power electronics applications [9].

The integrated passives have been applied in several areas in power electronics, such as the LC resonator [10], integrated igniter [11], the EMI filter with high frequency attenuation [12], the integrated passive module for ZVS asymmetric half bridge converter [13], the LCT application [14-16] and the LLCT application [17] in resonant converters. With the spiral winding structure, a developed LLCT module [17] reaches the power
density of 480W/in$^3$. With the developing passive integration technologies, integrated passives get into more and more areas in power electronics.

Compared to discrete passive components, passives integration aims to achieve the following characteristics:

- Functional integration for better performance.
- Reduction in volume, profile and mass for high power density.
- Modularization for mass manufacturing and low cost.
- Improved manufacturability.
- Improvement in mechanical reliability.

Incorporating with the active integration technology, a fully integrated power converter could be constructed to consist of only a number of integrated active stages and integrated passive stages, as shown in Figure 1-1. Such integration of reactive components in resonant and soft-switching power electronic converters has been reported in the literature [10-18].

1.3. Integration of power passives

The electromagnetic integration of passives takes on many different forms. The most common form of the utilization of “extra energy” is in the case of transformer leakage inductance. Since the value of coupling coefficient between transformer windings can never reach one, the leakage inductance cannot be completely removed. By applying the magnetic theory and circuit model, the leakage inductance can be represented by an inductor in series with the transformer. Many different methods have been proposed to
utilize the leakage inductance as a circuit element. The leakage inductance is used as a series resonant inductor in resonant circuit [19].

Another method of electromagnetic integration is the use of the integrated magnetics. This approach is based on the use of core designs with multiple winding areas and core sections, which provide common flux paths for inductor and transformer operation. By arranging the core structure, the magnetic interactions and core volume can be minimized [20, 21]. Due to the benefit of reduction on volume and loss, this approach has received much attention in recent years. Integrated magnetics, such as integrated coupled-inductors and current doubler have been successfully applied.

The application of transmission line structures in power electromagnetic has also been investigated. By applying good electromagnetic coupling between conductors of transmission lines, the transmission line transformer can work at both low power levels [22, 23] and high power levels [24]. By applying the impedance characteristic of transmission lines, two cascaded transmission lines functioned as a resonator in [25].

Another category of the electromagnetic integration is the integration of resonant structures. The typical structure of this category is the integrated LC series/parallel resonator. By exploiting dielectric material within winding conductors, the electrical and magnetic energy can be stored in one module, which carries both the inductive and capacitive functions. The integration of the parallel intra-winding capacitance to facilitate parallel resonance has been suggested [26]. The series LC resonance has been investigated with barrel type windings of bifilar foil conductors [27]. By combining the integrated LC resonance structure and the leakage inductance of a transformer, the integrated LCT and LLCT structures have been developed and applied in power converters [14-17].

This study focuses on the integrated resonant structures incorporating the capacitive energy storage within inductor and transformer structures. With the evaluation of existing integration structures, alternative structures are expected to improve power density of the integrated power passives. The previous work on integrated resonant structure is discussed in the next section.
1.4. Previous work on integrated passives

In the previous work, integrated passives have been theoretically and experimentally investigated on aspects of electromagnetic modeling, structure, processing technologies, loss modeling, and design algorithm based on the planar spiral winding structure.

In the past 30 years, the electromagnetic modeling of integrated passives has been developed with both lumped-parameter models and distributed-parameter models. Extended from the classical transmission line theory, the two lately proposed models [28, 29] represent generalized applications of the integrated passives, illustrate the physical nature of the performance, and accurately predict the electromagnetic behaviors at high frequencies. Though the developed theoretical models can predict the electromagnetic behaviors of integrated passives with known dimensions and materials, the parameter extraction from the real samples is still an obstacle for obtaining precise prediction. In design stage, simplified models need to be developed for specific integrated structures to optimize designs.

Several structures have been proposed to implement passive integration. In the early age, the wound structure was applied [30, 31]. To be wound, the dielectric needs to be flexible materials, such as polyester film. Due to the limited material properties, the wound structure cannot obtain large capacitance. Then, the planar structure was proposed associated with the application of high permittivity materials. The spiral winding structure has been successfully applied in power electronics field. To increase the design flexibility and mechanical reliability, the multi-cell structure was developed. However, this structure comes with large power loss and does not sufficiently utilize the space. It is clear that structural improvement is needed for further improvement on power density. Associated with the structures and materials applied, special interconnection and processing technologies also need to be developed.

Losses exist in every part of an integrated passive module: the magnetic core, the winding conductor, the dielectric material, and the leakage layer material, if applied. The losses in the core, the leakage material and the dielectric material are normally calculated with empirical equations, respectively, while the conduction loss is dependent on the structural factors. For planar structure, one-dimensional electromagnetic field is
practically assumed for simplification. Based on the orthogonality theory of skin and proximity effects, the conduction loss model was proposed. The conduction loss model for planar integrate modules concerns the non-uniform current distribution along the longitudinal direction. However, the estimation of losses has not been verified by either simulations or experimental results. To obtain more accurate loss estimation, the loss in interconnections, which was ignored in previous works, is needed to be identified.

The design algorithm for an LLCT module with the planar spiral winding structure has been developed by J.T. Strydom [17]. The design algorithm is also valid for the design optimization and the evaluation of this structure. The design algorithm is based on the assumption that all the required material properties are available. A practical design algorithm with practically available materials and processing technologies is needed to evaluate the new integrated structure proposed in this research work.

The thermal constraint is a major limitation for high power density. By applying thermal management to improve the thermal performance of integrated passive, the power handling ability of the power passives is improved and thus the power density increases without violating the thermal constraint.

1.5. **Aim of this study**

From the previous sections, it can be seen that the power density of integrated passives can be improved by applying integration technologies, while the previous work can be ameliorated from several aspects: structure, electromagnetic modeling, design algorithm, loss modeling, and thermal management. In this research, the LLCT integrated module is taken as the study case to investigate and evaluate the alternative integrated structures for high power density purpose.

The following research will be included in this dissertation.

Chapter 2 illustrates the principle of electromagnetic integration. The performance of integrated passives is related to the structure, including the detailed layout and terminal conditions. The behavior of integrated passives refers to the electromagnetic energy storage in the desired frequency range. The simplified models with lumped parameters are introduced and used to study the low frequency performance of integrated passives. As an important observation of the integrated passive behaviors, the asymmetric behavior
is reported and analyzed to present comprehensive knowledge of electromagnetic integration.

Chapter 3 illustrates the development of structures previously applied in integrated passives. The planar structure is suitable for power electronics applications and the planar spiral winding structure has been applied in high power density applications. However, the disadvantage of the spiral-winding structure is significant. Two alternative structures are proposed for further investigation. The interconnection structure and processing technologies are investigated.

Chapter 4 focuses on the electromagnetic design of the stacked structure proposed in Chapter 3. The comprehensive design algorithm covers the selection of design variables, analysis of the electromagnetic field distribution, calculation of the inductance and capacitance, design criteria and design maps for optimization and evaluation. The developed design maps are used to investigate the influencing factors and design limitation.

Chapter 5 applies loss models to estimate the losses in an LLCT module, respectively. Core loss model is based on the improved empirical equation, while the conduction loss model is based on the approximation of electromagnetic field distribution inside the module. A simple and accurate loss measurement is proposed to verify the loss models and to identify the efficiency of the components.

Chapter 6 illustrates heat extraction technology and its application on integrated power passives to improve the overall power density. With the calculated loss distribution, the internal temperature distribution of the module is investigated with simulation results. By embedding the high thermal conductivity electromagnetic inactive material as the heat extractors into the magnetic core, the module increases the heat removal ability and thus supports higher loss density under the same internal temperature drop. With slightly sacrificing the efficiency, the overall power density is improved.

In Chapter 7, two prototypes with the stacked structure are built based on the design approach proposed in chapter 4. The test results are compared with the calculation to verify the design and loss models. To verify the heat extraction technology, one of the prototypes is built with the embedded heat extractors and tested under the similar
operational conditions with the one without heat extractors. The evaluation of the experimental results of the stacked structure and the heat extraction technology is presented.

Chapter 8 presents the conclusions and the suggested future research work.
Chapter 2: Electromagnetic Integration of Passives

2.1. Principle of electromagnetic integration

In this dissertation, the electromagnetic integration of power passives refers to the combination of inductive and capacitive functions on one power electronics unit. The behavior of the integrated unit, which is described by frequency response curve, is the same as that of the combination of several discrete inductors and capacitors within certain frequency range. The simplest integrated electromagnetic passive is an integrated series or parallel resonator.

In power electronics field, the phenomena of combining capacitive and inductive functions on one component have been observed for a long time on two typical aspects: parasitic parameters and transmission line.

The parasitic effects exist in all power electronics components as the equivalent parallel capacitance (EPC) of the inductors and the equivalent series inductance (ESL) of the capacitors. In normal applications, the parasitic parameters are not expected and their existence leads to undesired impact to the circuit. The parasitic parameters associated with the active components induce additional current and voltage stress and increase operation loss. The equivalent parallel capacitance of a switch causes current stress by discharging the capacitor when the switch is turned on; while the equivalent series inductance leads to high voltage stress because high voltage is created across the ESL when the switch is turned off. The parasitic parameters in passive components limit the applicable frequency range of the components. In some applications, such as EMI filter, the parasitic parameters would be a serious problem influencing the performance [32, 33]. The passive electromagnetic components can be classified into three main types: resistors, capacitors and inductors (and transformers). Every real passive component is expected to have only one of the three characteristics as it is represented by the ideal model in the circuit diagram. However, every practical component exhibits these three characteristics to greater or lesser extent. Figure 2-1 shows a commercial capacitor and its electric performance represented by the impedance vs. frequency curves. The small signal test result shows that the capacitor has capacitive function at frequencies lower
than 1 MHz. When the frequency goes higher, the effect of the equivalent series inductance becomes stronger and starts to influence the electric behavior. The ESL and the capacitor form a series resonator, which has the resonant frequency of 13.39 MHz. At the resonant point, the magnitude of impedance is about 84 mΩ and the phase is 0°. At higher operation frequencies, the ESL dominates the electric performance and the component functions as an inductor.

(a) commercial capacitor 333

(b) Equivalent mode

(c) Impedance vs. frequency curves

Figure 2-1 Impedance vs. frequency curves and equivalent mode of a real capacitor

The observation of the frequency response curve can be summarized as following. A real component excited by alternating current always stores both electric energy and magnetic energy. It is because when electromagnetic wave propagates, both electric field and magnetic field exist. The reason that some components show inductive behavior and some show capacitive behavior is the components stores different amount of electric energy and magnetic energy due to different structures and materials. The ratio of the two types of energy stored is also related to frequency. For a specific component, such as a commercial capacitor stated above, at low frequencies, the stored electric energy dominates, thus the component demonstrates capacitive functions. At high frequencies, the stored magnetic energy dominates and the component demonstrates inductive functions.

It has been clearly recognized that any passive component with specific structure and material is possible to present capacitive function or inductive function or a combination
of both at certain frequency range. However, this phenomenon cannot be directly applied in power electronics applications. In power electronics circuit, the operation frequency is normally limited within a small range. The passive component needs to be designed with appropriate structure and materials to provide the desired function in the desired frequency range. Since the conventional capacitor and inductor are specially designed to store only electric energy and magnetic energy, respectively, the parasitic inductance of the capacitor or the parasitic capacitance of the inductor normally has very small value. Thus, in most cases, the parasitic parameters of the conventional passive components cannot be directly applied in the circuit. Only in some special cases, where the required passive parameter has the value close to the parasitic parameters, can the parasitic parameters be applied as part of the circuit parameters.

In power electronics applications, the integration of large capacitance and large inductance in one component is expected to reduce the total volume of passive components. To store equivalent amount of electric energy and magnetic energy in one component, the structures for conventional discrete component are no longer suitable. The transmission line has been studied as an alternative structure with good electric coupling and magnetic coupling between the two conductors of a conventional transmission line. The transmission line may be defined as a device for transmitting or guiding electromagnetic energy from one point to another [36]. The most common forms of transmission line are coaxial, two-wire and infinite-plane transmission lines.

Figure 2-2 shows a planar transmission line excited with alternating current and the corresponding electric and magnetic field lines. As we can see, both the electric field and the magnetic field concentrate within the space between the two conductors. The dimensions and materials in this space impact the stored electric and magnetic energy. Thus, from the electromagnetic integration viewpoint, transmission line combines distributed inductance and capacitance. Practically, the integrated inductance and capacitance is expected to be designable for different requirements, in which it normally requires relatively larger values than the distributed inductance and capacitance in the conventional transmission line structure. If a material having high permeability and high permittivity is inserted in the space between the two conductors, the stored magnetic and electric energy will be increased and adjusted by means of controlling the permittivity
Chapter 2: Electromagnetic Integration of Passives

and permeability of the material. However, though such a high permeability and high permittivity material is theoretically achievable according to the present material technology, there is no applicable commercial product so far. Moreover, the adjusting of values of permeability and permittivity would lead to difficulty in processing procedure. The conventional transmission line structure has been applied in power electronics field as transmission line transformers [22, 23], but it can hardly satisfy the requirement of most power electronics applications.

The inductance and capacitance to be achieved in one module is not necessarily correlated. Values of inductance and capacitance vary within a wide range. To practically implement integration, the magnetic energy and electric energy are separately stored in different space in an integrated module. To do so, the conventional transmission line structure is excited at two terminals on different conductors at the two ends, as terminal A and D, shown in Figure 2-3. With this structure, the magnetic energy is stored in the space around the two conductors and can be adjusted easily by controlling the surrounding magnetic cores, while the electric energy is stored in the space between the two conductors and controlled in the inserted dielectric material.

An additional benefit for the structure shown in Figure 2-3 is that it is easy to obtain multi-turn windings, which is important for obtaining large inductance. Because the magnetic energy is store outside of the two conductors, the shape and dimensions of the magnetic material is not necessarily limited by the conductors. The conductors can be wound around the magnetic material to form more than one turn.

Due to the benefits listed above, the structure in Figure 2-3 is taken as the basic structure of electromagnetic integration to implement integrated LC resonator. More complicated designs are based on the study of this basic integrated electromagnetic structure.
Chapter 2: Electromagnetic Integration of Passives

Figure 2-2 Electric and magnetic field distribution in a transmission line structure excited in the conventional way

Figure 2-3 Electric and magnetic field distribution in a transmission line structure excited at terminals at the two ends

2.2. Basic structure of integrated electromagnetic passives

The basic structure for integrated electromagnetic passives is theoretically demonstrated by the generalized two-conductor transmission structure model [28]. As shown in Figure 2-4, the transmission structure is a symmetric structure consisting of two conductors insulated with dielectric material. It can be represented as a two-port network and used to build the simplest integrated passive module, an LC series or parallel resonator.

The overall and distributed inductance and capacitance of a specific transmission structure are determined by the detailed dimensions and materials. However, the electromagnetic performance, which is represented by the frequency response curve of the structure, is determined by the terminal conditions. The terminal conditions include the selection of excitation terminals, the interconnections between terminals and the load connected to the terminals.
One two-conductor transmission structure has four terminals, A, B, C and D. Due to the symmetry, the structure has two typical forms of terminal combination. The excitation terminals can be terminals A and B or terminals A and D. If excited at terminals A and B, the transmission structure functions as a conventional transmission line. Its first LC series resonance occurs when the structure length meets the quarter wave length, while its high frequency resonant frequencies are multiples of the first resonant frequency. If excited at terminals A and D, the transmission structure functions as an LC series resonator at the first resonant frequencies, which is much lower than the frequency of the quarter wave length, but it has the same high frequency resonance as conventional transmission lines.

The interconnections between the redundant terminals also influence the performance. If the structure is excited at terminal A and B, and terminal C and D are shorted, the structure functions as an LC parallel resonator at the first resonant frequency, which is corresponding to the quarter wave length. If the structure is excited at terminal A and D, and terminal B and C are shorted, the structure functions as an LC parallel resonator at the first resonant frequency, which is much lower than the frequency of the quarter wave length. The classification of typical integrated resonators is shown in Figure 2-5. The corresponding frequency response curves shown aside are obtained from the small signal measurement results of a real integrated passive module.

The load connected to the terminals also takes part in the energy propagation and influences the terminal performance. For example, the structures shown in Figure 2-5 (b) is regarded to have load of infinite resistance connected to terminals C and D; while the structure in Figure 2-5 (d) has a load of zero resistance. L. Zhao presented the calculation results for this structure with different loads, as shown in Figure 2-6[28]. As it indicates, the load impedance, to some extent, determines the function of the transmission structure. With different resistive load values, the terminal function can vary between series resonator and parallel resonator. In the boundary situation, the transmission structure functions as a resistor. If the load is capacitive, the structure functions as a series resonator at the first resonance. If the load is inductive, the structure functions as a parallel resonator at the first resonance. If the load consists of more than one passive component, such as a combination of a capacitor and an inductor, the situation would be more complicated.
Though the load across terminals is a crucial factor determining performance of the transmission structure, it is not practical to design integrated passives by adjusting the load types and values. Generally, only the structures shown in Figure 2-5 are considered in the design of integrated power electronics passives. Due to the separation of the storage space of magnetic and electric energy, the structures shown in Figure 2-5 (b) and (d) are widely applied in power electronics component designs.

Figure 2-5 Classification of typical integrated resonators and corresponding frequency response curves
Chapter 2: Electromagnetic Integration of Passives

2.3. *Modeling of the transmission structure*

2.3.1. **Lumped parameter models**

The modeling of conventional transmission line structure, in which the two excitation terminals are at the same end, has been well developed [36]. To analyze the performance of the other structures shown in Figure 2-5, several lumped models have been proposed.

In 1975, Reeves first made the electromagnetic modeling of integrated passives with lumped models [37]. Later, P.N. Murgatroyed [38], Stielau [39] and M. Ehsani [40] also studied with lumped models. However, these lumped models cannot describe the high frequency behavior. Kemp’s quasi-distributed model [41] and J. Smit’s distributed model may predict the higher order resonance, but they have the difficulty to determine the
parameters and cannot show a deep insight into the physical meanings. Extended from the classical transmission line theory, L. Zhao proposed the generalized transmission structure theory [28].

2.3.2. Generalized transmission structure model

The transmission structure has infinite resonances at high frequencies. That means the full spectrum performance of the transmission structure cannot be completely represented by models consisting of limited number of lumped parameters. As the extension of the conventional transmission line model, the generalized transmission structure model has been proposed by L. Zhao [28].

For a structure as in Figure 2-4, a generalized equivalent circuit for an infinitesimal length $\Delta x$ is proposed by L. Zhao as shown in Figure 2-7. This model takes the magnetic and electric coupling and the losses of the structure into account. Different from the models for conventional transmission line, this model assumes that the current in the two conductors are not necessarily balanced. They can have any magnitude, flowing in either direction.

![Figure 2-7 Equivalent circuit of an infinitesimal length $\Delta x$ of a transmission structure with arbitrary terminal conditions](image)

Applying Kirchhoff’s loop and node equations for the circuit yields the differential equations (2-1) with the definitions in equations (2-2) (2-3) and (2-4).

\[
\begin{align*}
-\frac{d}{dx}V(x) &= zI_{1-2}(x) \\
-\frac{d}{dx}I_{1-2}(x) &= yV(x) \\
I_{1-2}(x) &= \frac{1}{2}(I_1(x) - I_2(x))
\end{align*}
\]  

(2-1)
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\[ z = 2[R + j\omega(L - M)] \quad (2-3) \]

\[ y = G + j\omega C \quad (2-4) \]

The propagation coefficient is

\[ \gamma = \sqrt{zy} = \alpha + j\beta \quad (2-5) \]

The characteristic impedance is

\[ Z_0 = \sqrt{\frac{z}{y}} = \sqrt{\frac{2[R + j\omega(L - M)]}{G + j\omega C}} \quad (2-6) \]

The differential equations has the general solutions of (2-7)

\[
\begin{align*}
V(x) &= V_a e^{\alpha x} + V_b e^{-\beta x} \\
I_{1-2}(x) &= I_a e^{\alpha x} + I_b e^{-\beta x} = -\frac{V_a}{Z_0} e^{\alpha x} + \frac{V_b}{Z_0} e^{-\beta x}
\end{align*}
\]

(2-7)

By applying different terminal conditions, the special solutions can be obtained and the terminal impedance and current and voltage distribution can be calculated.

Zhao’s model can accurately estimate the performance of integrated passives with simple structure. The model has been extended to analyze the cascaded structure and multi-conductors successfully.

However, there exist some shortcomings. The model requires tedious equations derivation for every change of boundary conditions in a simple structure. For complicated multi-conductor structures, the numerical solution becomes difficult, and the convergence problem is the key point to get the result.

2.3.3. Multi-conductor lossy transmission line model

To solve these problems, R. Chen proposed the coupled lossy transmission line theory, in which the common ground surrounding the transmission structure is added into to analysis. A two-conductor structure is regarded as two transmission lines (with respect to ground) with magnetic and electric coupling between them. The cross-sectional view is shown in Figure 2-8. The equivalent circuit of an infinitesimal section is modeled in Figure 2-9. To solve the coupled differential equations, the excitation wave is separated into several modes, which indicate that a wave can travel independently on a system of
ininitely long, uniformly coupled transmission line. A system of two coupled lines consisting of three conductors (including the ground) has two modes, even mode and odd mode. In general, a transmission line system consisting of \( n+1 \) conductors has \( n \) modes. For each mode, there is an equivalent transmission line corresponding to the excitation wave. From the solution obtained from classical transmission line theory can be combined through superposition law to obtain the special solution [29]. The multi-conductor lossy transmission-line model greatly simplifies the calculation and obtains the same estimation accuracy as the generalized transmission structure model. By extending this model, the performance of integrated passives can be explained as the performance of the combination of a discrete resonator determining the fundamental resonant frequency and a conventional transmission line determining high frequency resonances. The theory has been successfully applied to develop the EMI filter in DC/DC converters.

![Figure 2-8 A two-conductor transmission structure with ground](image)

![Figure 2-9 Equivalent circuit of an infinitesimal section](image)

### 2.3.4. Current and voltage distribution

The current and voltage distribution in the transmission structure is important for conduction loss modeling, which will be detailed in Chapter 5. The study of current and voltage distribution along the propagation direction is based on sinusoidal waveforms. The magnitude and phase of the current and voltage can be calculated with the models stated above. The current and voltage distribution is related to the ratio of the wave length and the structure dimension along the propagation direction. In other words, for a practical transmission structure with specific length, the current and voltage distribution is related to frequency.
For the structure in Figure 2-5 (b), the normalized current and voltage distributions at various frequencies are shown in Figure 2-10 [28]. In general power electronics applications (not including EMI filters), only the behavior around the first resonance is concerned. Compared to the frequencies of the high order resonance, the operation frequency of power electronics circuit is so low that the ratio of the wave length to the structure dimension is a small value. Thus, the linear current and voltage (magnitude) distribution shown in Figure 2-10 (a) best fits the real situation in integrated power passives.

(a) $\lambda/L = 0.016$

(b) $\lambda/L = 0.158$
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(c) $\lambda/L = 0.791$

(d) $\lambda/L = 1$

(e) $\lambda/L = 1.582$
2.3.5. Simplified models for design

Though the models discussed in section 2.3.3 can help to illustrate detailed behaviors of the integrated structures, they cannot be directly used to design integrated passive modules. In most power electronics designs, only the behavior around the first two resonances is of concern. Thus, a simplified model would be sufficiently accurate for design. As the transmission structure in Figure 2-11 (a), the typical frequency response curve in Figure 2-11(b) indicates that the structure functions as a series resonator at the first resonant frequency, which is much smaller than the frequency of the second series resonance. From design point of view, we can use the LC series resonator shown in Figure 2-11(c) to represent the transmission structure. The inductance in the model is the overall inductance generated by the winding; the capacitance is the overall capacitance generated by the two conductors and approximated as the parallel plate capacitance. In some cases, the lossless symmetric model in Figure 2-11(d) is used to intuitively represent the integrated structure with the emphasis on the symmetry feature of the structure. In some cases, the second resonant frequency is also concerned. The higher the second resonant frequency the better the high frequency performance of the structure is. The improved model in Figure 2-11(e) considers the parasitic parallel capacitance and is sufficient to predict the behaviors around frequencies up to the second resonance.

Figure 2-10 Normalized current and voltage (magnitude) distribution of the top and bottom conductors at various ratio of the wave length to the structure dimension (length) [28]
Chapter 2: Electromagnetic Integration of Passives

(a) Transmission structure applied for design of integrated power passives

(b) Frequency responds curve

(c) 1st order model for fundamental design

(d) Lossless symmetric model

(e) Improved model for second resonance estimation

Figure 2-11 Simplified models of integrated passives
2.4. Asymmetric structure of integrated passives

With the developing application of integrated power passives, the asymmetric performance has been observed in multi-layer-winding structure. Further, it has been found that almost all practical integrated passive modules, to some extent, are asymmetric. For the four-terminal integrated structure, if it is symmetric, the terminal pair of A and D is interchangeable with the terminal pair of B and C. The performance measured on the terminal pair of A and D should be the same with that measured on the terminal pair B and C. If the performance measured on different terminal pairs is different, that means the two terminal pairs are not exchangeable and the structure is asymmetric.

In a symmetric transmission structure, the distribution of the inductance and capacitance along the wave propagation direction is symmetric to a certain central line. However, due to the manufacturing tolerance induced in the processing procedure, the parameter distribution in practical modules is not uniform. In some structures, such as planar spiral winding structure, the parameters distribution is guaranteed by the processing technologies, and the asymmetry is not significant. However, in some structures, such as magnet wire winding structure with the hand-made windings, the asymmetric phenomenon is obvious. In complicated structures, for example multi-layer interleaved winding structure, the asymmetry is determined by the structure and unavoidable. The details of modeling and analysis of asymmetric integrated structure are presented in Appendix A.

The asymmetric phenomena in current integrated passive modules are not significant and the discrepancy between the unbalanced parameter values is small compared to the design value, because the numbers of turns and layers are not large and the discrepancy on winding position is small. However, in some applications, where the parasitic parameters, such as inherent leakage inductance in the integrated LLCT module, are applied as the part of the design parameters, the asymmetry of the structure would lead to practical issues in the real components. The details about such an asymmetric example, an integrated LLCT module with multi-turn center-tapped windings are addressed in Chapter 7.
2.5. Summary

Due to the characteristics of electromagnetic wave propagation, every physical component carrying alternating waves will transfer and store both magnetic and electric energy at the same time. Every physical component has infinite resonances in the full spectrum. The principle of electromagnetic integration for power electronics components is to design the specific structure, and obtain the required performance and resonance parameters within the desired frequency range. The basic structure for electromagnetic integration consists of two conductors. Associated with the selection of excitation terminals, the two-conductor structure can separate the spaces for magnetic energy storage and electric energy storage. More importantly, the two galvanically isolated conductors can achieve series resonance at the fundamental resonant frequency by getting the conduction current and displacement current in series. By shorting of the two conductors at the terminals, the conduction current and displacement current are parallel and the parallel resonance is obtained at the fundamental resonant frequency. The models for full spectrum behaviors have been applied to study the high frequency resonances and analyze the detailed current and voltage distribution in the integrated structure. For the design purpose, the simplified models have been proposed to calculate the dimensions for specific structures and estimate the behaviors around the first two resonances. The asymmetric behavior of practical integrated passive modules is observed and investigated.
Chapter 3: Structures for Power Passive Integration

3.1. Previous integrated electromagnetic components

In 1929, the first recorded integrated LC resonator, known as a “Wickelkondensator”, was made [42]. In 1949, the first patent for such a component was granted to Stickely in British [43]. This integrated LC resonator consists of a multi-turn inductor with barrel type windings of bifilar foil conductors with flexible dielectric material inserted in between, as shown in Figure 3-1. The dielectric material is used for insulation and to create capacitance. There is no magnetic material applied. Parallel or series resonance can be obtained by either connecting or insulating the two bifilar windings.

In 1975 Kemp reported the research about self-resonance in foil inductors [41]. He developed parallel resonant foil inductors by making coils from a single strip of aluminium foil and a slightly wider strip of plastics foil. The two foils are stacked and wound together on a core of plastic foil, using a capacitor-winding machine. The method of construction and the principal dimensions are shown in Figure 3-2. Though there is only one conductor in the module, the multi-turn structure leads to different electrical potential in different windings and thus stores electric energy, while the conduction current flowing in the multi-turn conductor generates magnetic energy. The single-conductor structure is the simplest form to obtain an integrated parallel resonator.
In 1975, Reeves reported the research work on an inductor-capacitor hybrid structure [37]. The four-terminal inductor-capacitor hybrid structure shown in Figure 3-3(a) is similar to the classical barrel integrated LC structure. A conventional type of capacitor is wound around a core of magnetic material. The metal plates are connected as in Figure 3-3 (b). It is just the same structure as shown in Figure 2-6 (a). Since the excitation terminals are at different ends, the conduction current flows in the same direction in each conductor, while the displacement current exists in the capacitive materials. The combination of conduction current in the two conductors generates magnetic flux in the core and stores magnetic energy in the air gap. The other two terminals are used as the output terminals and connected to a load resistor R. The unit is characterized by the capacitance C between the plates and by its “single-plate inductance” L which can be measured by using either one of the two windings to represent the combination of the two windings.
In 1990, Stielau proposed two alternative structures to build series and parallel resonant elements, as shown in Figure 3-4 [44]. They are actually based on the same concept as the classical barrel type LC. The two plates of a capacitor were rolled up as windings. The coil can have an air core or a core made of a magnetic material.

In 1992, M. C. Smit indicated that the physical limitation of the barrel type integrated structure is the requirement of flexible dielectric materials, which are normally of low permittivity. To obtain large capacitance, the ferro-electric ceramic (dielectric constant possibly larger than 1000) is preferred. Since these ceramics are brittle and rigid, it is most suited to planar structures. Though quarter wavelength transmission line has been widely applied as the solution for microwave resonators, the operation frequency is too high for power transmission. By cascading two different sections of transmission lines,
Smit proposed the transmission line type of integrated power passives and reduced the resultant resonance frequency to a value below 1 MHz [45].

The research started from the quarter wavelength transmission line, as shown in Figure 3-5. When the load is short-circuited, the input impedance of the structure is infinite, thus resulting in a parallel tuned circuit. If the quarter wavelength transmission line has an open-circuited load, the input impedance appears to be zero, and results in a series tuned circuit.

If the same quarter wavelength transmission line is used, but the transmission line medium is changed at some discrete point along the length of the line, it will now show that the resonance can occur at a much lower frequency than that of the initial quarter wavelength transmission line. As illustrated in Figure 3-6, the structure has the initial length L as the quarter wavelength, but actually separated into two sections. Section 1 consists of a short circuited transmission line in ferrite medium ($\mu_r >> 1$ and $\varepsilon_r = 1$) resulting in a large characteristic impedance, while section 2 consists of an open circuit transmission line in a dielectric medium ($\mu_r = 1$ and $\varepsilon_r >> 1$) resulting in a low characteristic impedance. The structure can also be regarded as a short-circuited high characteristic impedance transmission line (section 1) connected to a low characteristic impedance transmission line (section 2) as the load. Apparently, the high characteristic impedance part stores magnetic energy and the low characteristic impedance part stored electric energy. Since the two conductors in the structure are connected at the terminals, the conduction current in the two conductors is in series, while in parallel with the displacement current, the structure functions as a parallel resonator. If the two conductors are galvanically isolated, the conduction current in the two conductors are in series with the displacement current, the structure would function as a series resonator. The
structures and the corresponding performance are shown in Figure 3-7 and Figure 3-8. By changing the properties of the mediums, the resonant frequency will shift.
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Since the space between the two plates of section 2 is magnetic field free, a compact structure was proposed by folding section 2 into section 1, as shown in Figure 3-9.

![Figure 3-9 Folded integrated resonator structure](image)

In 1990, L-C-T integration was introduced by Ferreira and Smit. The technology was successfully demonstrated in a high frequency series resonant converter [46]. Two configurations were proposed and illustrated in Figure 3-10 and 3-11. The primary winding is made of bifilar foils, separated by a suitable dielectric, wound directly on the core. The single-turn secondary forms a shell which contains the core and the primary. The space between the primary winding and the internal walls determines the leakage or series inductance.

![Figure 3-10 Toroidal L-C-T transformer structure](image)
In 2001, the wound structure was developed by Crebier to obtain an integrated L-C-T structure, based on the flexible conduction and dielectric materials, such as copper foils and Kapton tapes. Figure 3-12 shows an LCT module with wound structure [31]. With the wound structure, the integrated modules with simple functions, such as, LC resonator and LCT module, can be fabricated in a compact volume. However, if complicated functions are required, this structure would be difficult to assemble due to the interconnection problem. If many turns/layers are applied in this structure, the diameter of the outer layer will be much larger than that of the inner layers. In addition, the wound structure requires flexible dielectric materials, which are normally of low permittivity. Thus, a large copper area is needed to create a small value of capacitance. All of these limit the application of the wound structure.
In 2005, the magnet wire winding structure was proposed by Wenduo for the applications that require large inductance and small capacitance [47]. Figure 3-13 shows an integrated igniter (LC series resonator) with the magnet wire winding structure. The structure can be regarded as an extension of an inductor. The commercial magnet wire and core are assembled as an inductor. By introducing an additional winding and applying insulation tape of high breakdown voltage, the inter-winding capacitance is created between the two windings and utilized as the integrated resonant capacitance. For the applications that large inductance (turns number) and small integrated capacitance is required, the inter-winding capacitance is large sufficient to meet the design requirement. Since the value of the inter-winding capacitance is greatly dependent on the winding layout, dimensions and processing procedure, the obtained capacitance varies within a certain range. Thus, this structure is not suitable for cases where accurate parameter values are required.

3.2. Planar structure for integrated passives

Along with the proposal of integrated structures listed above, the planar structure has become more and more concerned due to the advantages of low profile, achievable large capacitance, precise parameters guaranteed with the processing technologies, and mass production ability.

3.2.1. Planar spiral winding structure

The planar spiral winding structure has been widely applied to fabricate integrate passive modules in all kind of power electronics circuit. Figure 3-14 shows the exploded view of an integrated LLCT (inductor-inductor-capacitor-transformer) module with planar spiral winding structure. Due to the planar feature, the rigid ceramic material can
be applied to create large capacitance. Associated with the application of ceramic, the metallization technology is applied to deposit copper on the ceramic substrate. Other solid state technologies, such as chemical etching and laser cutting are also needed to fabricate the planar windings. Though the processing technology is relatively complicated, the spiral winding structure is widely used due to the merits of large capacitance, low profile, large heat dissipation area and the ability to obtain high power density. An integrate LLCT module with the spiral winding structure reaches the power density of 480 W/in³ [17].

However, the spiral winding structure has several disadvantages. With the metallization technologies, sputtering and electroplating, large area metallization is difficult to obtain and the metalized structure is mechanically unreliable. The overhang of the spiral winding functioning as interconnections occupies large space and materials, which do not contribute to the energy storage and transformation. The inefficiency utilization of material and space limits the power density and leads to unnecessary conduction loss. The interconnection between different spiral winding layers is a hurdle for fabrication.

![Image](image1.png)

Figure 3-14 The exploded view of an integrated LLCT module with planar spiral winding structure

### 3.2.2. Multi-cell structure

As an improved planar structure, the multi-cell structure was proposed to provide the ability of flexible design and mass production. The multi-cell structure applies many LC cells as the basic electromagnetic element. The so-called LC cells are high permittivity
ceramic strips covered with copper layers on both sides. Surrounded by the magnetic core, the internal LC cells can be connected by outside PCB through the interconnections on the terminals, as shown in Figure 3-15 [48]. The many terminals of the internal LC cells provide great interconnection possibility and design flexibility. With different interconnections, a module can have different functions. The identified construction of internal cells is suitable for mass production. The small metallization area reduces the mechanical vulnerability.

The disadvantage of this structure is that the outside PCB interconnection reduces the power density. To pursue high power density, two alternative structures are proposed in this dissertation and will be discussed in the following two sections, respectively.

3.3. **Symmetric single layer structure**

The symmetric single layer structure comes from the idea of stacking two multi-layer structures and fabricating interconnections on the vertical surface [49]. Figure 3-16 shows the symmetric single layer structure.
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The schematic diagram has been proposed to represent the symmetric single layer structure in the analysis. The schematic diagram abstracts the dimension characteristic of the 3D module and represents the structure feather and the interconnections in a 2D form, as shown in Figure 3-17.

The symmetric single layer structure has more design flexibility than the multi-layer structure for consisting of more internal LC cells. Complicated Multi-functional modules can be obtained with the symmetric single layer structure.

Figure 3-18 illustrates the process to derive a complicated integrated structure. It needs to be noted that an example of a multi-functional module. All parameters in the equivalent lumped parameter model are corresponding to the materials in the integrated module, and the parameter values can be controlled by changing the dimensions and material properties.
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(a) Two windings magnetically coupled

(b) add series capacitor and leakage inductance

(c) add parallel capacitor

(d) add inter-winding capacitor

Figure 3-18 Derivation of a complicated integrated structure

By exploiting the vertical surface to fabricate interconnections, the symmetric single layer has more compact size. The implementation of interconnections requires special interconnection structure, material, and processing technologies. One of the implementation approaches is illustrated in Figure 3-19.

(a) LC cell and lead

(b) layout

37
The symmetric single layer structure has been applied to build integrated LLCT module. Figure 3-20 shows the structure of an integrated LLCT module with a two-turn winding in the primary and a two-turn winding in the secondary.

Since there is only one winding layer, the internal cells are placed parallel inside the magnetic core. This leads to some disadvantages. When this structure is applied for an integrated LLCT module, the inherent leakage inductance would be too larger due to the small winding height. As shown in Figure 3-20 (a), since the height of the LC cells is small, the winding window height is small and close to the air gap in the core. Thus, the inherent leakage inductance is close to the magnetizing inductance. In cases that the inherent leakage inductance is larger than the required leakage inductance, this structure is not valid.

The parallel layout of the cells leads to high conduction loss due to proximity effect. For an LLCT module, the primary winding and secondary winding are in the same winding window and carrying current in opposite directions. The magnetic field inside the winding window cannot be regarded as a one-dimensional magnetic field. The magnetic field in the vertical direction penetrates the conductor and induces eddy current at the edge. Figure 3-21 shows the simulation results of the magnetic field distribution in the winding window and the current distribution in the cells of one layer. Moreover, the insulation space between the adjacent cells results in a wide winding window, which leads to large core volume and thus core loss. To overcome these shortcomings, the stacked structure is proposed in the next section.
Chapter 3: Structures for Power Passive Integration

(a) Structure of the integrated LLCT module

(b) the real sample

(c) the schematic diagram

(d) the equivalent circuit of lumped parameter model

Figure 3-20 The integrated LLCT module with the symmetric single layer structure

(a) Magnetic field distribution in winding window

(b) Current distribution in cell conductors of one layer

Figure 3-21 Significant proximity effect in the symmetric single layer structure
3.4. **Stacked structure**

Instead of placing internal cells horizontally, the stacked structure has the cells stacked vertically, as shown in Figure 3-22 [50]. By doing so, the disadvantages of the symmetric single layer structure are avoided. The inherent leakage inductance can be eliminated to a small value, since the leakage flux between the two windings is small due to the compact structure; the proximity effect loss on the horizontal direction can be suppressed by the approximate one-dimensional field, while the proximity effect loss on the vertical direction is not significant due to the small thickness of conductors; from the constructional viewpoint, the stacked structure has the most compact size among the structures discussed above.

![Stacked structure images](image)

Figure 3-22  An integrated LC series resonator with the stacked structure

The stacked structure has compact size and small vertical surface area. Thus, the implementation of interconnections becomes the key issue of this structure. The development of the interconnection structure is illustrated in the following.

In the 1st interconnection approach shown in Figure 3-23, leads are made on the terminals of LC cells, and are extended out of the core to penetrate the vias on Pyralux. The leads are soldered with the pad on the Pyralux circuit. The disadvantage appears in the fabrication. It is difficult to make via on Pyralux and get the cells penetrate the vias. In addition, the interconnection structure is not mechanically robust.
In the 2nd approach shown in Figure 3-24, the leads are made on the vertical side of the cells. The interconnections are made on PCB or Pyralux circuit, on which the pads are soldered with the leads. With Pyralux or PCB, complicated interconnection traces can be fabricated, but the area on the vertical side of the cells is too small. The small area ceramic is not able to mechanically support the Pyralux or PCB circuit.
In the 3\textsuperscript{rd} approach shown in Figure 3-25, the core and LC cells are encapsulated as a sealed module. The side surface is metalized after polishing. By etching off the undesired copper, the interconnection traces remains on the surface. The parameter of the encapsulated module, such as the leakage inductance and magnetizing inductance is not adjustable. The encapsulated module has a CTE mismatch problem.

![Figure 3-25 3\textsuperscript{rd} approach of the interconnection structure](image)

In the 4\textsuperscript{th} approach shown in Figure 3-26, each lead has a small width, and the positions of corresponding leads are in the same vertical line. Only the central core and the cells are encapsulated. The vertical interconnections are fabricated on the side surface of the encapsulated module.

The four approaches are suitable for different applications. The fourth approach is currently preferred. In the interconnection structures discussed above, the lead occupies a small space. The small conduction area leads to current concentration, which is associated with the increase of loss. If the number of cells is too large, the space occupied by each lead would be too small to fabricate. For this case, multi-layer interconnection structure has to be applied.
Chapter 3: Structures for Power Passive Integration

3.5. Comparison of integrated structures

The spiral winding structure, the symmetric single layer structure and the staked structure have been proposed in the above. A question comes that which structure uses space most efficiently. With the same core cross-sectional area and magnetic flux density, to provide the same electromagnetic performance (inductance), which structure occupies the smallest volume? For simplification, we ignore the insulation and dimensions tolerance and assume the three integrated structure has the same number of turns, $n$, core length, $l$, and core thickness, $T$. We also assume the integrated windings in the three structures have the same width, $w$ and thickness, $h$. The spiral winding has $k$ turns per layer and totally $m$ layers. The product of $k$ and $m$ is $n$. The winding layout and dimensions of the three structures are shown in Figure 3-27.

The volume of the three structures can be expressed as the following.

Volume of the spiral winding structure:

$$V_{spiral} = 2(l + 2kw)(nwh + 2mhT + 2kwT + 4T^2)$$  \hspace{1cm} (3-1)

Volume of the symmetric single layer structure:

$$V_{SSL} = 2l(nwh + 2hT + 2nwT + 4T^2)$$  \hspace{1cm} (3-2)

Volume of the stacked structure:

$$V_{Stack} = 2l(nwh + 2nhT + 2wT + 4T^2)$$  \hspace{1cm} (3-3)
Besides the winding layout parameters, \( k, m, \) and \( n \), the four parameters, \( w, h, l \) and \( T \), represent the dimensions of the integrated structure. Practically, the length \( l \) is much larger than the winding width \( w \), and the winding width is much larger than winding thickness \( h \) \((l \gg w \gg h)\). The winding width could be larger or smaller than the core thickness. Set the core thickness as a constant. Define the ratio of the winding width to the core thickness as

\[
k_{wT} = \frac{w}{T}
\]  

(3-4)

Define the core dimensional ratio of the length to the core thickness as

\[
k_{LT} = \frac{l}{T}
\]  

(3-5)

Define the winding dimensional ratio of the winding height to the winding width as

\[
k_{hw} = \frac{h}{w}
\]  

(3-6)

The three ratios are set as the three variables to determine the overall volume. Practically set the values of the three variables in the ranges of \((0.1 < k_{wT} < 10)\) and \((5 < k_{LT} < 25)\), and \((10 < k_{hw} < 50)\).

The volume comparison can be represented by the value of the volume ratios. Firstly, we consider the volume ratio of the symmetric single layer structure to the stacked structure, \( R_{ss,st} \). Comparing equations (3-2) and (3-3), we can see the difference between the two equations is related to \( k_{wT} \) and \( k_{hw} \). Set the turn number 10 and take the two variables to do calculation. The volume ratio \( R_{ss,st} \) is shown in Figure 3-28. It shows that the ratio \( R_{ss,st} \) is always larger than one. That means the volume of the symmetric single layer structure is always larger than that of the stacked structure, when providing the same electromagnetic performance. The reason is because the winding width is larger than the winding thickness. Since the core surrounds the winding, the corresponding core dimension in the width direction is larger than that in the height direction. In the symmetric single layer structure, the cells are horizontally placed in parallel, while the cells are vertically stacked in the stacked structure. The difference of winding layout enlarges the difference of the horizontal dimension and the vertical dimension. As shown
in Figure 3-28, with the reduction of the value of the winding dimensional ratio, the $R_{ss_st}$ value increases.

Figure 3-27  winding layout and dimensions of the three integrated structures
The spiral winding structure has two extreme winding layouts as shown in Figure 3-27 (d) and (e). When there is only one column of windings, the spiral winding structure is similar with the stacked structure, but its overall length is larger than that of the stacked structure due to the overhang. When there is only one winding layer in the spiral winding structure, it is similar with the symmetric single layer structure, but its overall length is larger than that of the symmetric single layer structure. Again it is due to the overhang.

Set three layout conditions for the spiral winding structure as: (1) $k = 1$, $m = 10$, (2) $k = 2$, $m = 5$, (3) $k = 10$, $m = 1$. Take $k_{wT}$ and $k_{hw}$ as two variables and set the constant length of 50. The calculated ratios are shown in Figure 3-29. Take $k_{wT}$ and $k_{LT}$ as two variables and set the constant winding dimensional ratio of 0.1. The calculated ratios are shown in Figure 3-30. The calculation results show that the volume ratio of the spiral winding structure to the stacked structure increases, when the ratio of winding width to core thickness increases, the winding dimensional ratio and the core dimensional ratio decrease.
Chapter 3: Structures for Power Passive Integration

(a) $m = 10$, $k = 1$

(b) $m = 5$, $k = 2$
Figure 3-29 Volume comparison between the spiral winding structure and the stacked structure with the constant length L = 10
Figure 3.30  Volume comparison between the spiral winding structure and the stacked structure with the constant winding dimensional ratio, $k_{hw} = 0.1$
3.6. Summary

In the past several decades, the integrated structures developed with various forms, but the basic concept of electromagnetic integration is unchanged. To integrate inductive and capacitive functions in one unit, the structure of the unit should be able to store the required amount of magnetic energy and electric energy. Basically, the integrated structure consists of two conductors. The electric energy storage occurs in the dielectric material between the two conductors. The magnetic energy is normally stored in the magnetic field with a magnetic core or air core.

To form the capacitor, the barrel type structure requires flexible dielectric materials, which are normally of low values of permittivity. This limits the achievable capacitance. Another problem of the barrel type structure is that the terminals are covered by the windings and it is difficult to make interconnections. These prevent the barrel type structure from complicated integration structures consisting of multi-functions. The barrel type structure is also an asymmetric structure that the inner diameter is much smaller than the outer diameter when the number of turns is large. To contribute the same inductance, the outer windings have significantly larger conduction loss than the inner windings. It is the geometric limitation preventing the barrel type structure from large size. Similar disadvantages occur on the magnet wire winding structure. Furthermore, restricted processing technologies are required to get accurate control on the parameters obtained in the magnet wire winding structure, due to the uncertainty induced in the processing procedure.

The planar structure has been developed widely due to the geometric characteristics. The planar windings make it possible to apply high permittivity dielectric materials to obtain large capacitance normally required in power electronics field. The thin windings help to reduce the profile of the whole module and provide large cooling area. The regular shapes provide good packaging factors and improve power density. Since the terminals are bare on the same surface with the windings, it is easy to make interconnections on a planar structure. The proposal of LC cells induces great design flexibility into the integrated structure. The large amount of terminals creates numerous possibilities to make interconnections, while one interconnection structure corresponds to a specific combination of electromagnetic functions.
Every integration structure is suited for certain applications. Among the four existing planar integration structures, the spiral winding structure, the multi-cell structure, the symmetric single layer structure and the stacked structure, the stacked structure shows the best potential to minimize loss and maximize power density. With existing material technology and processing technologies, the stacked structure is suitable to the power range from several hundred watts to several kilo-watts. The detailed investigation will be presented in the next several chapters.
Chapter 4: Electromagnetic Design

4.1. Introduction

In the previous chapter, alternative structures to implement integrated passive modules have been presented. Due to the capability to achieve high power density, the stacked structure is selected as the alternative structure to implement integrated power passive module. To further investigate the characteristics and limitation of the stacked structure, the structure is applied in an integrated LLCT module in a DC/DC converter, as shown in Figure 4-1. Though the models of basic integration structure have been introduced in Chapter 2, the integrated LLCT module is with more complicated structure and needs specific model for design. The electromagnetic design approach for the integrated LLCT module with spiral winding structure has been developed by J.T. Strydom [51].

Loosely based on the previous design work, the design approach for the integrated LLCT with the stacked structure is proposed in this dissertation. The constructional details about integrating the resonant capacitance $C_r$ and inductance $L_r$, the transformer $T$ and the parallel inductance $L_m$ with the stacked structure are illustrated. The electromagnetic field distribution inside the integrated LLCT structure is investigated to calculate the electromagnetic energy and estimate the losses. Several design variables are set and vary in certain ranges to develop the design maps showing the relationship between interested characteristics and the design variables. Based on the valid designs in the design maps, the optimal design can be selected according to various criteria.

![Figure 4-1 The circuit and diagram of the integrated LLCT module and the DC/DC converter](image-url)
4.2. **Construction of the integrated LLCT module**

The integrated LLCT module provides the combined functions as a transformer with the primary winding in series with a series resonator and in parallel with an inductor. The equivalent lumped parameter model is shown in Figure 4-1.

The integrated structure is developed based on a transformer having copper foil windings and implemented with the stacked structure. By using LC cells to replace the copper foil in the primary side, the resonant capacitance $C_r$ is induced. The resonant inductance $L_r$ can be generated by using the leakage inductance between the primary and secondary windings. Since the leakage inductance between transformer windings is not avoidable, normally it is applied as the resonant inductance in the LLC resonant converter. The inherent leakage inductance is due to the magnetic energy stored in the winding space. It is fixed by the detailed winding structure and dimensions and normally of a small value, smaller than the required resonant inductance. Thus, the additional leakage inductance is needed to adjust the overall leakage inductance to the required value. The low permeability material, such as ferrite polymer composite (FPC), is inserted into the two transformer windings as the leakage layer to provide additional path for magnetic flux. By adjusting the dimensions (mainly the thickness) of the leakage layer, the additional leakage inductance is tunable. According to the transformer modeling, a transformer with small magnetizing inductance can be regarded as an ideal transformer with the primary winding connected in parallel with an inductor. By increasing the air gap and lowering the magnetizing inductance of the transformer, the parallel inductor $L_m$ can be integrated into the module. The structure of the integrated LLCT module and the interconnections are shown in the schematic diagram in Figure 4-2. For easy description, a rectangular plane coordinate system is associated with the structure. The axes $x$, $y$, and $z$, represent the dimensions along the length, width and height directions, respectively.

The number of the turns is an important design factor, which is normally determined at the beginning of the design. The large number of windings turns leads to much conduction loss, large volume, large inherent leakage inductance and difficulty to implement interconnections. Considering efficiency and power density, the number of turns in the planar integration structures is normally set one or two times of the minimum
winding number. For the design requirement shown in Figure 4-1, the turns ratio is 4:1:1, thus, the primary winding has four turns and the secondary has one turn in each of the center-tapped windings.

4.3. Design assumptions and parameters

The electromagnetic performance is determined by many factors. To simplify the design procedure and pay attention on the dominant factors representing the structure features, several assumptions are made.

- All material properties are independent of temperature. The properties of the materials are selected for a temperature that is representative of those within the
prototype under operating conditions. Since the Curie temperature of the Ferro-electric ceramic and ferrite are the main temperature constraints, the design temperature for the material should below 100°C. Since the electrical parameters of the prototype are verified with small signal test, which is made at room temperature, all material properties used in the design are based on the values at room temperature.

- The materials are considered to have electrical characteristics that are independent of electromagnetic field and frequency. The permittivity and loss of the dielectric material is regarded constant during the operation. The background knowledge about the characterization of the dielectric is given in [52, 53].

- The space occupied by insulation material and the manufacture tolerance is estimated with fixed practical values according to the application and the processing technologies. It is taken into account in the design calculation. The induced parasitic capacitance by insulation materials is not considered in the design.

- Perfect processing technologies are assumed, so that the LC cells, interconnections, cores and the whole module can be built exactly as they are designed.

- The influence of the skin and proximity effect on the values of electrical parameters is ignored in the electromagnetic design.

- The influence of the leads and interconnections is ignored. The space occupied by leads and interconnections is not taken into account. The current concentration and conduction loss in leads and interconnections are not considered in design approach.

- One-dimensional electromagnetic field distribution is assumed in the cross section of the module. The distribution of conduction current is assumed uniform along the winding width direction, y axis.

- Thermal issues are not considered in the electromagnetic design, because the temperature estimation requires complicated 3D models.
Chapter 4: Electromagnetic Design

With the simplifications to the structure model, it is possible to design an LLCT structure to a given set of equivalent electrical circuit requirements as listed in Table 4-1. A list of all the structural parameters is given in Table 4-2 and the material parameters are listed in Table 4-3.

The three target quantities, resonant inductance, \( L_r \), resonant capacitance, \( C_r \), and magnetizing inductance \( L_M \), are determined by not only the constructional parameters, but also the electromagnetic field distribution. As the vital background of the quantity calculations, the electromagnetic field distribution inside the LLCT module will be discussed in the next section.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant inductance [H]</td>
<td>( L_r )</td>
</tr>
<tr>
<td>Resonant capacitance [F]</td>
<td>( C_r )</td>
</tr>
<tr>
<td>Magnetizing inductance [H]</td>
<td>( L_M )</td>
</tr>
<tr>
<td>Transformer voltage [V(_{\text{rms}})]</td>
<td>( V_T )</td>
</tr>
<tr>
<td>Voltage waveform parameter for core</td>
<td>( k_t )</td>
</tr>
<tr>
<td>Resonant capacitor voltage, rms [V]</td>
<td>( V_{Cr} )</td>
</tr>
<tr>
<td>Operating frequency rang [Hz]</td>
<td>( f_i - f_h )</td>
</tr>
<tr>
<td>Primary side transformer current, rms [A]</td>
<td>( I_{pri} )</td>
</tr>
<tr>
<td>Secondary side transformer current, rms [A]</td>
<td>( I_{sec} )</td>
</tr>
<tr>
<td>Peak inductor current [A]</td>
<td>( I_L )</td>
</tr>
<tr>
<td>Turn’s ratio of the transformer</td>
<td>( n )</td>
</tr>
</tbody>
</table>
Table 4-2 LLCT constructional parameters

<table>
<thead>
<tr>
<th>Constructional Parameters</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>Turn’s number of the primary winding</td>
<td>$n_p$</td>
</tr>
<tr>
<td>Turn’s number of the secondary winding</td>
<td>$n_s$</td>
</tr>
<tr>
<td>Window width [m]</td>
<td>$w_{win}$</td>
</tr>
<tr>
<td>Window height [m]</td>
<td>$h_{win}$</td>
</tr>
<tr>
<td>Module width [m]</td>
<td>$w_m$</td>
</tr>
<tr>
<td>Module height [m]</td>
<td>$h_m$</td>
</tr>
<tr>
<td>Module length [m]</td>
<td>$l$</td>
</tr>
<tr>
<td>Core thickness [m]</td>
<td>$T_{Core}$</td>
</tr>
<tr>
<td>Leakage layer thickness [m]</td>
<td>$T_L$</td>
</tr>
<tr>
<td>Cross-sectional area of the leakage layer [m$^2$]</td>
<td>$A_L$</td>
</tr>
<tr>
<td>Insulation layer thickness, vertical [m]</td>
<td>$T_{in}$</td>
</tr>
<tr>
<td>Insulation layer thickness, horizontal [m]</td>
<td>$T_{inh}$</td>
</tr>
<tr>
<td>Copper thickness of the primary winding [m]</td>
<td>$T_{cp}$</td>
</tr>
<tr>
<td>Copper thickness of the secondary winding [m]</td>
<td>$T_{cs}$</td>
</tr>
<tr>
<td>Winding width [m]</td>
<td>$w_{cu}$</td>
</tr>
<tr>
<td>Aspect ratio of the cell</td>
<td>$k_{cell}$</td>
</tr>
<tr>
<td>Dielectric material thickness [m]</td>
<td>$T_{di}$</td>
</tr>
<tr>
<td>Air gap length [m]</td>
<td>$T_{gap}$</td>
</tr>
<tr>
<td>Height of the primary winding [m]</td>
<td>$h_{pri}$</td>
</tr>
<tr>
<td>Mean magnetic path length [m]</td>
<td>$l_{mmL}$</td>
</tr>
<tr>
<td>Core cross-sectional area [m$^2$]</td>
<td>$A_{core}$</td>
</tr>
<tr>
<td>Core volume [m$^3$]</td>
<td>$V_{core}$</td>
</tr>
<tr>
<td>Module volume [m$^3$]</td>
<td>$V_m$</td>
</tr>
</tbody>
</table>
### Table 4-3 LLCT material parameters

<table>
<thead>
<tr>
<th>Material Parameters</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>Relative permittivity of dielectric material</td>
<td>$\varepsilon_r$</td>
</tr>
<tr>
<td>Relative permeability of magnetic core</td>
<td>$\mu_r$</td>
</tr>
<tr>
<td>Permittivity of free space [F.m$^{-1}$]</td>
<td>$\varepsilon_0$</td>
</tr>
<tr>
<td>Permeability of free space [H.m$^{-1}$]</td>
<td>$\mu_0$</td>
</tr>
<tr>
<td>Design flux density of transformer [T]</td>
<td>$B_T$</td>
</tr>
<tr>
<td>Flux density due to leakage inductance [T]</td>
<td>$B_{\text{leakage}}$</td>
</tr>
<tr>
<td>Flux density in leakage layer [T]</td>
<td>$B_l$</td>
</tr>
<tr>
<td>Average conduction current density at primary [A/m$^2$]</td>
<td>$J_{cu}$</td>
</tr>
<tr>
<td>Resistivity of conductor material [Ω.m]</td>
<td>$\rho_{cu}$</td>
</tr>
</tbody>
</table>

#### 4.4. Electromagnetic field distribution

The required parameters in the equivalent circuit are corresponding to the electromagnetic energy stored in the appropriate space inside the integrated module. The analysis of the electromagnetic field distribution provides the necessary relationship between the stored electromagnetic energy and the module dimensions and material properties.

For a module with the length much larger than the width, the fringing effect at the terminals is ignored and the magnetic flux is considered parallel to the cross section. Since the width of LC cells is much larger than the thickness, the electromagnetic field on the cross-sectional plane is assumed a one-dimensional field. In other words, the magnetic flux inside the winding window is regarded parallel to the $y$ axis.

Thus, on any specific cross section inside the module, the magnetic field is a one-variable function with respect to the height and represented as $H(z)$. The magnetic field distribution on the cross section in half of the integrated structure is shown in Figure 4-3. Here we assume that alternating current is input into the primary winding and output at the secondary windings and all primary windings carry the same amount of current, so do
the secondary windings. There is no phase difference between the current in primary side and that in secondary side. The magnetic field can be estimated with the equation of ampere’s law. The red line in Figure 4-3 shows the simulation results of the magnetic field from a 2D model. Approximately, it can be simplified as a piecewise linear function, shown as the black dash line.

In power electronics applications, the module length is much shorter than the length of the excitation waves. It is reasonable to assume that the conduction current in the primary winding distributes linearly along the propagation direction, $x$ axis. So does the electric potential distribution on the conductor, while the voltage between the two conductors is constant along $x$ axis. Thus, the displacement current is constant at any longitudinal position.

In an LC series resonator, the conduction current is in series with the displacement current. The overall equivalent current in the LC cell windings is the combination of the conduction current and the displacement current, and has the same magnitude with the input current. In the calculation of magnetizing inductance, the induced magnetizing inductance concerns only the number of the current loops, while the detailed current distribution inside the cells does not matter. Thus, in this calculation, the LC cell winding can be replaced with a conductor winding carrying current with the same magnitude.
However, in the calculations of the conduction loss, the variance of magnetic field along the propagation direction should be taken into account. In the primary winding, the conduction current distribution and the magnetomotive force distribution along the $x$ axis is shown in Figure 4-4 and 4-5.

The electromagnetic field distribution discussed above is used in the electromagnetic design. The electromagnetic field distribution in the secondary winding will be discussed for the calculation of conduction losses in next chapter.

Figure 4-4 Conduction current distribution in the primary winding
4.5. Variables relationship for design

The cross-sectional winding layout and the relevant dimensional parameters are shown in Figure 4-6. The core length $l_{core}$, is equal to the cell length $l_{cell}$, and the module length $l_m$ and denoted as $l$.

The dimensional parameters have the relationship as following.

Effective core area:

$$A_{core} = 2 \times T_{core} \times l$$  \hspace{1cm} (4-1)

Mean magnetic flux length:

$$l_{mmL} = (2 \times T_{core} + w_{win} + h_{win}) \times 2$$  \hspace{1cm} (4-2)
Chapter 4: Electromagnetic Design

Primary winding height:

\[ h_{pri} = 4 \times (T_{dl} + 2 \times T_{cp} + 2 \times T_{in}) \]  \hspace{1cm} (4-3)

Secondary winding height:

\[ h_{sec} = 2 \times (T_{cu} + 2 \times T_{in}) \]  \hspace{1cm} (4-4)

Winding window height:

\[ h_{win} = h_{pri} + h_{sec} + T_{k} \]  \hspace{1cm} (4-5)

Module height:

\[ h_{m} = 2 \times (h_{win} + 2 \times T_{core}) \]  \hspace{1cm} (4-6)

Copper area per cell:

\[ A_{cu} = w_{cu} \times l \]  \hspace{1cm} (4-7)

Cell width:

\[ w_{cell} = w_{cu} + 2 \times T_{inh} \]  \hspace{1cm} (4-8)

Winding window width:

\[ w_{win} = w_{cell} + 2 \times T_{inh} \]  \hspace{1cm} (4-9)

Module width:

\[ w_{m} = w_{win} + 2 \times T_{core} \]  \hspace{1cm} (4-10)

Module volume:

\[ V_{m} = w_{m} \times h_{m} \times l \]  \hspace{1cm} (4-11)

Based on the selected material properties, the design is to find out appropriate dimensions to obtain the three required quantities, the magnetizing inductance \( L_M \), the leakage inductance \( L_r \) and the resonant capacitance \( C_r \).

Taking the reluctance in the air gap and the magnetic core into account, the magnetizing inductance is calculated as follows:

\[ L_M = 2 \cdot \frac{\mu h_{p}^2}{2l_{gap} + \frac{l_{mm}}{A_{core} \cdot \mu_r A_{core}}} \]  \hspace{1cm} (4-12)
With the assumption of uniform voltage distribution, the capacitance of each cell is calculated as

$$C_{cell} = \frac{\varepsilon_0 \varepsilon_r A_{cell}}{T_{di}}$$

(4-13)

The resonant capacitance is the combination of all the cell capacitance according to the interconnections. With different interconnections, the overall equivalent capacitance is different [50]. For the interconnection structure shown in Figure 4-2 (b), the resonant capacitance is the product of the cell number and the capacitance of a single cell.

The inherent leakage inductance can be calculated based on the electromagnetic field distribution. To simplify the calculation, the primary winding is considered to carry uniform current along the propagation direction. The assumption of the one-dimensional magnetic field distribution illustrated in Figure 4-3 is applied. When the secondary winding is short-circuited, the electromagnetic energy stored in the space inside the winding window is calculated as

$$E_w = 2 \cdot w_{win} l_{core} \left[ \int_0^{h_{pwin}} \frac{1}{2} \mu_0 \left( \frac{n_p^2 I_p^2 z}{w_{win} h_{pwin}} \right)^2 dz + \int_0^{h_{swin}} \frac{1}{2} \mu_0 \left( \frac{n_p^2 I_p^2 z}{w_{win} h_{swin}} \right)^2 dz \right]$$

(4-14)

$$E_w = \frac{1}{3} \frac{\mu_0 n_p^2 l_p^2 (h_{pwin} + h_{swin})}{w_{win}} l_{core} \approx \frac{1}{3} \frac{\mu_0 n_p^2 l_p^2 h_{win} l_{core}}{w_{win}}$$

(4-15)
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The stored electromagnetic energy is equal to the magnetic energy stored by the inherent leakage inductance, thus the inherent leakage inductance is calculated as

\[
L_{\text{int}} = \frac{2E_w}{I_p^2}
\]  

(4-16)

and simplified as

\[
L_{\text{int}} = \frac{2}{3} \frac{\mu_0 n_p^2 h_{\text{win}} l_{\text{core}}}{w_{\text{win}}} \approx \frac{2}{3} \frac{\mu_0 n_p^2 h_{\text{win}}}{k_{\text{cell}}}
\]  

(4-17)

Normally the inherent leakage inductance is smaller than the required resonant inductance, thus the additional leakage inductance is needed and calculated as

\[
L_{\text{add}} = 2 \cdot \frac{\mu_0 n_p^2}{2T_{\text{gap}}} + \frac{1}{A_{\text{core}}} + \frac{w_{\text{win}}}{\mu_r A_{\text{core}}} + \frac{w_{\text{win}}}{\mu_l A_L}
\]  

(4-18)

The inherent leakage inductance is also the design criterion used to filter those invalid designs having inherent leakage inductance larger than the required resonant inductance.

To calculate the geometry parameters, the equations represent the relationship between the geometry parameters and the circuit design parameters are needed.

The relationship between the flux density and the core cross-sectional area:

\[
V_T = A_{\text{core}} B_T n_p k_{Tf}
\]  

(4-19)

The flux density in the leakage layer:

\[
B_{\text{leakage}} = \frac{1}{2} \frac{L_r I_L}{n_p A_{\text{core}}}
\]  

(4-20)

The definition of the skin depth:

\[
\delta_{\text{cu}} = \sqrt{\frac{\rho_{\text{cu}}}{\pi f_h \mu_0}}
\]  

(4-21)

The definition of the cell aspect ratio:

\[
k_{\text{cell}} = \frac{w_{\text{cell}}}{l}
\]  

(4-22)

The definition of average current density:


4.6. **Design approach and design map**

Different from the design approach proposed by Strydom [51] and Hofsajer [54], the design in this research work is with more practical considerations. Instead of assuming that material characteristics can be realized to any required value, the material properties used in the design calculation are based on the practical values obtained from manufacture datasheets.

The proper materials and interconnection structure, as well as the number of turns, are selected before the calculation of dimensions begins. Three parameters are set as the independent design variables and have their values vary within practical ranges, respectively. The magnetic flux density $B_T$ represents the stored magnetic energy; the average current density $J$ is concerned with the conduction loss; the aspect ratio of the cell, $k_{cell}$, is a geometry factor representing the cell shape. For a practical design the copper thickness is optimized for minimum conduction loss based on the assumption of one-dimensional electromagnetic field distribution as discussed in Chapter 5. Thus, the copper thickness is fixed with respect to the frequency. According to the equations (4-19), (4-23), (4-13) and (4-22), there are only two independent variables, either $B_T$ and $J$, or $B_T$ and $k_{cell}$.

To satisfy the assumption of one-dimensional field distribution, the cell aspect ratio is expected to be less than a certain value, for example 0.5. Compared to the average current density, the cell aspect ratio has more practical meaning in the design. Thus, parameters $B_T$ and $k_{cell}$ are selected as the design variables.

The design steps are as follows:

1. Select the structure to implement the module and determine the interconnection structure.
2. Select the magnetic and dielectric materials to obtain property values and minimum dimensions for processing.

\[
J_{cu} = \frac{I_{ps}}{W_{cell} T_{cu}} \quad (4-23)
\]
3. Determine the appropriate processing technologies, and estimate mechanical tolerance and insulation space.

4. Determine the number of turns. Experience shows that the lowest and second lowest possible number of turns often yields the smallest structures.

5. Determine the copper thickness based on the optimal value obtained from loss calculation, illustrated in next chapter. To minimize the conduction loss, normally the copper thickness is set within the range from half to two times of the skin depth.

6. Calculate the geometry parameters and losses with the values of the two design variables.

7. Generate the design maps with the calculation results to show the relationship between the interested characteristics and the design variables.

8. Apply design constraints to filter undesired or invalid designs and provide the design maps within valid design range.

9. Select the optimal design with various tradeoffs.

In step 8, the leakage inductance is applied as one of the design constraints to filter invalid designs due to the geometric limitation. After step 9, the thermal simulation is needed to estimate the thermal conditions. The flow chart of the design steps is shown in Figure 4-7.

![Figure 4-7 The flow chart of the electromagnetic design](image-url)
Since each of the two design variables has a set of values, the design results are two-dimensional arrays showing how the design variables influence the interested parameters. The design maps can be presented as 3D graphs. The calculation is implemented by programs coded in Matlab.

For a design example with the specifications listed in Table 4-4, the design maps without any constraints are calculated and shown with respect to four interested characteristics, power density, total loss, inherent leakage inductance and core thickness, in Figure 4-8. The loss models included in the design program will be detailed in Chapter 5.

The design maps can also be shown in form of contour. Figure 4-9 shows the design maps with respect to power density and total loss in form of contour.

By applying design constraints, the invalid designs can be filtered for the design maps. By changing values of the specific design constraints, the influence of the constraints to the valid design range can be demonstrated in design maps.

Table 4-4 Specifications of an example design

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant inductance $L_r$</td>
<td>$0.5\mu H$</td>
</tr>
<tr>
<td>Resonant capacitance $C_r$</td>
<td>$50 , nF$</td>
</tr>
<tr>
<td>Magnetizing inductance $L_M$</td>
<td>$30 , \mu H$</td>
</tr>
<tr>
<td>Transformer voltage $V_T$</td>
<td>$200 , V$</td>
</tr>
<tr>
<td>Primary winding current, rms $I_{pri}$</td>
<td>$3 , A$</td>
</tr>
<tr>
<td>Secondary winding current, rms $I_{sec}$</td>
<td>$12 , A$</td>
</tr>
<tr>
<td>Operating frequency rang $f_r$</td>
<td>$1 , MHz$</td>
</tr>
<tr>
<td>Turns ratio of the transformer $n_p: n_s$</td>
<td>$4 : 1:1$</td>
</tr>
</tbody>
</table>
(a) Power density

(b) Total loss
Figure 4-8 Design maps with respect to the interested characteristics – example, no constraint applied

(c) Inherent leakage inductance

(d) Core thickness
Figure 4-9 Design maps shown in form of contour – example, no constraint applied
(a) constraint: air gap length should be larger than 5μm

(b) constraint: air gap length should be larger than 15μm

(c) constraint: inherent leakage inductance should be smaller than the required resonant inductance
Figure 4-10 shows the design map with respect to power density under different constraints.

In the integrated LLCT module, the magnetizing inductance is used as one of the resonant inductance and expected to have a small value. So, normally the air gap of the core is used to reduce the magnetizing inductance. However, in the stacked structure, there are only four turns of primary winding and the core cross-sectional area is set small to minimize core volume. It results in a small magnetizing inductance. In some designs, the maximum achievable magnetizing inductance is smaller than the required value and that leads to invalid design. Magnetic cores used in the stacked structure are planar C core and I core. When the C core and I core are assembled to form a closed loop for magnetic flux, air gap exists between the two cores. The minimum value of the air gap is
determined by the manufacturing tolerance and the assemble technique. For example, in certain conditions, very small manufacturing tolerance exists, and the minimum air gap length is 5μm. The consideration of air gap length can be applied as the design constraint to filter those designs in which the calculated air gap is less than 5μm. The resultant design map with respect to the power density is shown in Figure 4-10 (a). It shows that when the magnetic flux density and the cell aspect ratio both reach large values, the design becomes invalid. It is because the larger magnetic flux density corresponds to the smaller core cross-sectional area, while the larger cell aspect ratio corresponds to the larger mean magnetic flux length. Decreasing core cross-sectional area and increasing mean magnetic flux length leads to the same result -- lowering maximum achievable magnetizing inductance. More generally, the minimum air gap length should be larger than 15μm. The corresponding design map is shown in Figure 4-10 (b). The valid design range is smaller than that in Figure 4-10 (a).

To implement the resonant inductance, the inherent leakage inductance should be smaller than the required resonant inductance, so that the additional leakage layer can be used to adjust the overall leakage inductance to the required value. By applying this as the design constraint, the design map is shown in Figure 4-10 (c). It shows the inherent leakage inductance is influenced little by the magnetic flux density (core cross-sectional area), but related to the cell aspect ratio. It is because the inherent leakage inductance corresponds to the magnetic energy stored in the winding window, whose dimensions are influenced by the cell aspect ratio.

Considering the limitation from the processing technology, the minimum acceptable core thickness is 1.5 mm. The design map with this constraint is shown in Figure 4-10 (d). The boundary between the valid designs and the invalid designs represents those designs having exactly core thickness of 1.5 mm.

To obtain valid designs, all the above constraints need to be satisfied. The design map with all the three constraints is shown in Figure 4-10 (e). For various design specifications and various limitations from materials and processing technologies, similar constraints can be set to obtain the valid designs.

The presented design approach provides a tool to evaluate designs with respect to the interested characteristics. The design maps show the relationship between the interested
characteristics and the design variables intuitively based on numerical calculation. It can be easily extended to the design of other integrated structures.

4.7. Summary

The implementation of the integrated LLCT module is based on a planar transformer. By applying the leakage inductance and magnetizing inductance as the two resonant inductances, the integrated LLCT module can be implemented with the stacked structure using LC cells to form the primary winding. With the analysis results of electromagnetic field distribution, the electromagnetic parameters values can be calculated. Based on the assumptions of materials properties and processing technologies, the calculation of the electromagnetic parameter values is simplified. With considerations of practical issues, the design approach is proposed. Two independent design variables are set to providing the design maps showing the relationship between the interested characteristics and the design variables. With the design constraints, the valid design range shrinks. The resultant design maps show the influence of the constraints to the valid design range. In specific applications, it can be used to evaluate the limitation of a specific structure under certain conditions.
Chapter 5: Electromagnetic Loss Modeling for the Stacked Structure

5.1. Introduction

In this chapter, the losses in different materials of the integrated LLCT module are estimated, respectively. The losses in the integrated module can be categorized into three parts: dielectric, magnetic and conductive.

The dielectric loss generally refers to the loss in the resonant capacitance. So far, there is no accurate loss model and valid data for the dielectric materials. Normally the loss in dielectric materials is estimated by the imperial equations with the estimated value of loss factor. It hardly provides accurate estimation. High permittivity materials show nonlinear performance under various electric field, frequency and temperature. Both the permittivity and the loss factors have significant change with the changing electric field, frequency and temperature [55, 56]. The loss factor value obtained from the manufacturer’s datasheet has a fixed value, but will be applied for all the various conditions. The waveform applied to the resonant capacitor generally consists of significant harmonic components; nevertheless, the empirical equation is obtained from experiments made by sinusoidal waveforms. These are reasons leading to the inaccurate estimation of dielectric loss.

The magnetic losses include the losses in the magnetic core and in leakage material. Since the difficulty of obtaining necessary parameter values of the leakage material, the loss in the leakage material is not estimated. Due to the leakage flux, the flux density varies in different parts of the core. Thus the core loss density varies in the different position of the core. The variance is not taken into account and then will increase the error in core loss estimation. Generally, the voltage exerted on the transformer is not sinusoidal; the loss analysis needs to accommodate an arbitrary voltage excitation.

The conduction losses can be separated into two parts – skin effect and proximity effect losses. The conduction current distribution is a function of geometry. So is the induced magnetic field. Since the in-circuit current is not sinusoidal, Fourier analysis is applied
and the harmonics are used to calculate the conduction loss. The conduction loss in interconnections is also calculated and added into the estimated total loss.

Since different materials in the integrated module are electromagnetically coupled and thermally coupled, it is difficult to measure the losses separately. To verify the accuracy of the loss models, we compare the estimated total loss with the measured in-circuit loss under various operational conditions. If the estimation results agree with the measurement results under each operational condition, the loss models are considered valid. This verification method is valid only when all the three loss models are accurate and the estimated values exactly match the measured values. If any discrepancy exists, it is difficult to tell which loss model causes the problem.

5.2. Dielectric losses

The high permittivity dielectrics tend to be ferro-electric and exhibit similar hysteresis loops as the ferro-magnetic ceramics [57]. The shape and size of the hysteresis loops are functions of electrical field, frequency and temperature. Since the design and implementation of the integrated module are based on the in-house ceramics, it is possible to setup experiments measure the loss characteristics of those dielectric materials [55]. However, the loss of some dielectric materials, such as NP0 and N1250 is too small to be accurately measured under general operational conditions. Thus, to estimate the dielectric loss in these materials, the equivalent loss factor (\( \tan \delta \)) is applied. The loss in the dielectric is calculated as

\[
P_d = \omega C V^2_{ac-coupled} \tan \delta
\] (5-1)

This empirical equation is independent of the geometry parameters of the structure. It assumes that the fringing effect is negligible; the structure is homogenous; the electrical filed distribution is uniform and the voltage is constant along the winding length direction. In a integrated LLCT module, the voltage across the resonant capacitor is not measurable. Thus, the rms value obtained from the circuit simulation with the same operational conditions is applied for the dielectric loss calculation.
5.3. Core losses

With the assumption of sinusoidal excitation and uniform magnetic flux distribution inside the core, the core loss density can be calculated by the empirical equation, Steinmetz equation (SE) [58]. The only requirement is that the material constants are known or can be extracted from the manufacturer’s datasheet. The terms $\alpha$ and $\beta$ in equation (5-2) are within the ranges: $1<\alpha<3$ and $2<\beta<3$ for ferrite. The material characteristics are also temperature dependent. The Steinmetz equation can be rewritten to include quadratic temperature dependence in °C [59].

$$P_v = kf^\alpha B^\beta$$  \hspace{1cm} (5-2)

$$P_v = kf^\alpha B^\beta (ct_0 - ct_1T + ct_2T^2)$$  \hspace{1cm} (5-3)

Since the operating temperature is normally fixed within a small range and the temperature does not influence the calculated value much, the simple equation (5-2) is widely used in practical designs as it yields quick and accurate estimation. However, this equation is valid for sinusoidal excitation only.

To overcome this limitation, some improved approaches have been developed based on the Steinmetz equation [60-62]. Noting the core loss is directly related to the magnetization velocity and that in turn is proportional to the rate-of-change of the induction $dB/dt$, the modified Steinmetz equation (MSE) is proposed (5-4) [63]. Determining the average $dB/dt$ over one cycle and normalizing it with respect to a sinusoidal waveform yields the relationship expressed in terms of an equivalent frequency as shown in (5-5).

$$P_v = C_m f_{eq}^{\alpha-1} B^\beta f_r$$  \hspace{1cm} (5-4)

$$f_{eq} = \frac{2}{\Delta B^2 \pi^2} \int_0^\pi \left( \frac{dB}{dt} \right)^2 dt$$  \hspace{1cm} (5-5)

However, the MSE approach is not valid for cases with DC bias. Another disadvantage of MSE is that the manner in which MSE includes the dependence of loss on $dB/dt$ does not specifically match the frequency dependence of the Steinmetz equation. This makes the MSE conflict SE in some specific cases and results in the anomalies in the behavior of the loss predicted by the MSE. Mathematically consistent with the Steinmetz equation,
the Generalized Steinmetz equation (GSE) (5-6) was proposed by J. Li to overcome the limitations of MSE [64]. In the verification experiment estimating the loss in a toroidal core, the GSE provides the accurate estimation.

\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha \left| B(t) \right|^{\beta-\alpha} dt
\]  
(5-6)

\[
k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} \cos^\alpha \sin^{\beta-\alpha} d\theta}
\]  
(5-7)

The GSE also has its disadvantages and is not accurate at some cases. Moreover, the GSE is criticized by that “The dependence of the instantaneous loss in GSE in only the instantaneous parameters” is in fact a problem in practice. To avoid the problem, the iGSE was proposed by K. Venkatachalam [65].

\[
P_{v_{-iGSE}} = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt
\]  
(5-8)

where

\[
k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} \cos^\alpha \sin^{\beta-\alpha} d\theta}
\]  
(5-9)

With further mathematically derivation, the iGSE is compared with the MSE. The equation in (5-8) (5-9) is reformulated as

\[
P_{v_{-iGSE}} = \frac{k}{T} \cdot \frac{\left(\Delta B\right)^{\beta-\alpha} \cdot \int_0^T \left| \frac{dB}{dt} \right|^\alpha dt}{\int_0^{2\pi} \cos^\alpha \sin^{\beta-\alpha} d\theta}
\]  
(5-10)

While the MSE in (5-4) (5-5) can be reformed to

\[
P_{v_{-MSE}} = \frac{k}{T} \cdot \frac{\left(\Delta B\right)^{\beta-2} \cdot \left[ \frac{dB}{dt} \right]^2 dt}{2^{\beta-1} \pi^{\alpha-1} \int_0^{2\pi} \cos^\alpha \sin^{\beta-\alpha} d\theta}
\]  
(5-11)

Compare the above two equations, it is obviously that when \( \alpha = 2 \), they are exactly the same. Thus, the MSE can be regarded as a specific form of iGSE with the fixed value of parameter \( \alpha \).
In this research, iGSE is applied for the core loss estimation. To apply iGSE, the functions of $dB/dt$ and $B(t)$ are needed. With the parameter values provided in the design specifications, the circuit simulation is made to give the voltage and current waveforms exerted on the LLCT module. The data acquired from the waveforms is processed by Fourier expansion. The first 80 items of the Fourier series are apply to reconstruct the voltage and current waveforms as continuous functions of time, $V(t)$ and $I(t)$. By applying Ampere’s law, the magnetic flux $B(t)$ can be derived from the current function $I(t)$ as

$$B(t) = \frac{n_p}{T_{gap} + \frac{I_{mlL}}{\mu_0 \mu_r \mu_0}} \cdot I(t), \quad (5-12)$$

The function of $dB/dt$ can be obtained from the differential of the magnetic flux function $B(t)$. The function of $dB/dt$ can also be obtained from the voltage function directly by applying Faraday’s law as

$$\frac{dB(t)}{dt} = \frac{1}{n_p \cdot A_{core}} \cdot V(t). \quad (5-13)$$

Then the function $B(t)$ can be obtained by making integral of equation (5-13).

The functions $B(t)$ obtained from the current waveform and the voltage waveform are slightly different. Figure 5-1 shows the calculated B(t) curved with the data extracted from in-circuit measured waveforms. The difference is due to the assumption made when applying Ampere’s law to calculate $B(t)$. The equation (5-12) assumes that the permeability of the core material is independent of the magnetic field. It is not the truth in practical case. Thus, the $dB/dt$ and $B(t)$ obtained from the voltage waveforms are regarded more accurate and used in the core loss estimation.

Since the integrated LLCT module consists of resonator in series with the primary winding of the transformer, the voltage across the primary winding can not be measured in the experiment. Thus, the voltage across the secondary winding is used to calculate the magnetic flux density.

Since the geometry parameter $k_i$ in (5-9) is independent of time $t$, it can be extracted from the integral as constants. In the design program, the integrals of the voltage and current are calculated in advance and replaced by two coefficients.
Chapter 5: Electromagnetic Loss Modeling for the Stacked Structure

5.4. Conduction losses

Conduction losses increase dramatically with frequency due to eddy current effects. The published research work about predicting conduction losses can be traced back to 1940 by Bennet [66]. In 1966, Dowell developed the one-dimensional model to compute ac resistance and winding inductance of a winding portion for sinusoidal excitation [67]. Dowell’s model was extended by the following research work [68] and commonly referred to as Dowell method to be distinguished from the Ferreira method [69]. Venkatraman applied Dowell’s results to non-sinusoidal excitation using Fourier analysis [70]. Carsten normalized Venkatraman’s results using resistance of a conductor one skin depth in height [71]. In 1978, Jongsma developed an extensive algorithm for design of transformers using an approximation based on Dowell’s results [72]. In 1979, Perry analyzed a general N-layer solenoid and derived power-loss expressions for each layer [73]. In 1987, Vandelac & Ziogas incorporated aspects of Dowell, Jongsma, Carsten and Venkatraman to extend Perry’s analysis [74]. This approach is widely applied to predict the conduction losses in round-wire windings by replacing the round conductors with
square conductors of the same cross-sectional area and then substituting a conductor foil for the square conductors in the same layer, and resulting analytical solution of the one-dimensional model.

The Ferreira method is to apply the well known Bessel functions solution. Ferreira indicated that orthogonality exists between skin effect and proximity effect when the applied magnetic field due to other conductors (proximity field) is assumed to be uniform over the conductor cross section [75]. If the conductors have an axis of symmetry and the applied field is uniform and parallel to the symmetric axis, the calculation equation of conduction losses can be mathematically decoupled into two parts, representing the losses due to skin effect and proximity effect respectively.

In the planar integrated structures, the electromagnetic field can be approximated as one-dimensional field. The external magnetic field of a planar winding conductor can be induced by the current in other conductor or the displacement current; also can be the fringing field close to the air gap of a core. Due to the large copper area for providing the resonant capacitance, the current density of the displacement current is much smaller than that of the conduction current. The influence of the displacement current on the magnetic field distribution is neglected. To simplify the calculation, the fringing effect at the area close to the air gap and the terminals is also neglected. Since the width of the copper layer is much larger than the thickness, it is assumed that the electromagnetic field on the cross section of the integrated module is of one-dimensional field distribution as discussed in chapter 4.

The orthogonality principle is applied for the planar integrated structure to separate the calculation of skin effect loss and proximity effect loss. The requirement is that the current in a given conductor does not significantly influence the overall magnetic field around the conductor. For a structure with a large number of conductors, this is generally the case. Since the field distribution in the primary winding is different from that in the secondary windings, the conduction losses in the two windings are calculated separately.

**5.4.1. Conduction loss in the primary winding**

The following assumptions are made to simplify the conductor loss analysis:
One-dimensional magnetic distribution, which means that the magnetic field is perpendicular to the conduction current and parallel to the conductor surface.

- Sinusoidal current excitation. For non-sinusoidal current, the analysis can be extended by Fourier expansion.
- Homogenous structure and constant material properties.

With the above assumptions, an infinite long strip conductor with the width much larger than the thickness is studied. When the conductor is carrying ac current $I$ (peak value), the skin effect loss per length of the conductor is given by:

$$
P_{s} = \frac{w}{2\sigma} \int_{0}^{h} |J_{x}|^{2} |dy| = \frac{I^{2}}{4w\sigma\delta} \frac{\sin\nu + \sin\nu}{\cosh\nu - \cos\nu} 
$$

where $\sigma$ is the conductor conductivity, $h$ is the conductor height, $w$ is the conductor width, $\delta$ is the skin effect and

$$
\nu = \frac{h}{\delta}. 
$$

For the applied external magnetic field, $H_{s}$, the proximity effect loss per length is given by:

$$
P_{p} = \frac{w}{2\sigma} \int_{0}^{h} |J_{x}|^{2} |dy| = \frac{wH_{s}^{2}}{\sigma\delta} \frac{\sin\nu - \sin\nu}{\cosh\nu + \cos\nu}. 
$$

In more general cases, the magnetic field around the conductor is not uniform. Assuming the boundary condition of a non-uniform magnetic filed around a strip conductor carrying current $I$ is:

$$
H_{s1} = \frac{(k-1)I}{w} 
$$

$$
H_{s2} = \frac{KI}{w} 
$$

where $k$ is any real number.

For this un-uniform surface magnetic field proximity problem, Ferreira has proved that it is equivalent to the problem of the conductor is placed in a uniform magnetic field [76], where
\[ H_s = \frac{H_{s1} + H_{s2}}{2} = \frac{(2k-1)I}{2w}. \] (5-19)

Considering the un-uniform current distribution along the \( x \)-axis, the current in each conductor is a function of the horizontal position, \( I(x) \). Also, the magnetic field at the top and bottom boundaries of each conductor is a function of the horizontal position, \( H_s(x) \), which can be calculated through \( I(x) \) by applying ampere’s law. Thus, the total skin effect loss and proximity effect loss in the primary winding are given as

\[
P_{\text{skin}} = \int_0^w I^2(x) \cdot \frac{\sinh v + \sin v}{\cosh v - \cos v} \, dx, \quad (5-20)
\]

\[
P_{\text{prox}} = \int_0^w \frac{H_s^2(x)w}{\sigma \delta} \cdot \frac{\sinh v - \sin v}{\cosh v + \cos v} \, dx. \quad (5-21)
\]

The functions \( I(x) \) and \( H_s(x) \) are linear and can be derived according the structure and interconnections. The calculation of conduction loss in each layer of the primary winding with the structure shown in Figure 4-5 is illustrated in the following. Figure 4-5 shows a four-turn primary winding has eight LC cells connected in parallel and is excited with current \( I \) at the two terminals. Assuming linear current distribution along the longitudinal direction, the current amplitude distribution in each conduction layer is a linear function of distance \( x \), determined by the terminal values. With the known current in each conductor layer, the magnetomotive force in each conductor layer is the average of the magnetomotive force on the top and bottom surfaces of the conductor layer and can be expressed as:

\[
M_{a(1,2)}(x) = I \cdot \frac{1}{16l} x
\]

\[
M_{a(1,1)}(x) = I \cdot \left( \frac{1}{2} + \frac{1}{16l} x \right)
\]

\[
M_{a(2,2)}(x) = I \cdot \left( \frac{9}{8} + \frac{1}{16l} x \right)
\]

\[
M_{a(2,1)}(x) = I \cdot \left( \frac{13}{8} + \frac{1}{16l} x \right)
\]

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\[ M_{a(3,2)}(x) = I \left( \frac{9}{4} + \frac{1}{16l} x \right) \]

\[ M_{a(3,1)}(x) = I \left( \frac{11}{4} + \frac{1}{16l} x \right) \]

\[ M_{a(4,2)}(x) = I \left( \frac{27}{8} + \frac{1}{16l} x \right) \]

\[ M_{a(4,1)}(x) = I \left( \frac{31}{8} + \frac{1}{16l} x \right) \]

\[ M_{b(1,2)}(x) = I \left( \frac{1}{8} - \frac{1}{16l} x \right) \]

\[ M_{b(1,1)}(x) = I \left( \frac{5}{8} - \frac{1}{16l} x \right) \]

\[ M_{b(2,2)}(x) = I \left( \frac{5}{4} - \frac{1}{16l} x \right) \]

\[ M_{b(2,1)}(x) = I \left( \frac{7}{4} - \frac{1}{16l} x \right) \]

\[ M_{b(3,2)}(x) = I \left( \frac{19}{8} - \frac{1}{16l} x \right) \]

\[ M_{b(3,1)}(x) = I \left( \frac{23}{8} - \frac{1}{16l} x \right) \]

\[ M_{b(4,2)}(x) = I \left( \frac{7}{2} - \frac{1}{16l} x \right) \]

\[ M_{b(4,1)}(x) = I \left( 4 - \frac{1}{16l} x \right) \] \hspace{1cm} (5-22)

For the applied structure, the magnetic field in side can be regarded zero, thus the magnetic field in each conductor layer can be expressed as

\[ H_{a,b(i,j)}(x) = \frac{M_{a,b(i,j)}(x)}{w_{win}} \] \hspace{1cm} (5-23)

where
Thus, the proximity effect loss along the conductor length is expressed as

\[ P_{p(i,j)}(x) = I^2 \cdot \frac{1}{h_{cu} \cdot w_{win} \cdot \sigma} \cdot \nu \cdot G \cdot M^2_{a(i,j)}(x) \] (5-27)

where

\[ G = \frac{\sinh \nu - \sin \nu}{\cosh \nu + \cos \nu} \] (5-28)

To obtain the overall proximity effect loss of a conductor, the integral along the cell length is applied:

\[ P_{p(i,j)} = I^2 \cdot \frac{1}{h_{cu} \cdot w_{win} \cdot \sigma} \cdot \nu \cdot G \cdot \int_0^l M^2_{a(i,j)}(x)dx = I^2 \cdot R_{dc} \cdot \nu \cdot G \cdot k_{ij} \] (5-29)

where

\[ k_{ij} = a^2_y + \frac{1}{2} a_y cl + \frac{1}{3} c^2 l^2 \] (5-30)

by taking the approximation of \( w_{cu} \approx w_{win} \)

\[ R_{dc} = \frac{l}{h_{cu} \cdot w_{cu} \cdot \sigma} \] (5-31)

Similarly, the skin effect loss can be expressed as

\[ P_{s(i,j)}(x) = \frac{I(x)^2}{4 \cdot w_{win} \cdot \sigma \cdot \delta} \cdot \frac{\sinh \nu - \sin \nu}{\cosh \nu + \cos \nu} = I(x)^2 \cdot R_{dc} \cdot \nu \cdot F \] (5-32)

where

\[ F = \frac{\sinh \nu - \sin \nu}{\cosh \nu - \cos \nu} \] (5-33)

The overall skin effect loss of a conductor is

\[ P_{s(i,j)} = \int_0^l I(x)^2 \cdot R_{dc} \cdot \nu \cdot F \cdot dx = I_e^2 \cdot R_{dc} \cdot \nu \cdot F \] (5-34)
where

\[ I_c = \sqrt{\frac{I_a^2 + I_b^2 + I_d^2}{3}} \]  

(5-35)

\( I_a \) and \( I_b \) refer to the current at the two terminals of the conductor.

### 5.4.2. Conduction loss in the secondary winding

In the secondary windings, the current is uniform along the winding length. Thus, both the current distribution and the magnetic field distribution stay constant along the x-axis.

With the center-tapped structure, the secondary winding consists of two independent windings. Each winding operates during half of the operational period. In terms of that, the two windings are exactly the same. The skin-effect loss in the two windings is the same and can be calculated with (5-20). Since each winding operates for only half a period, the skin-effect loss is half of the calculated value.

In the stacked structure, the two copper layers of the secondary winding are in different positions. Thus, the magnetic field exerted on the two winding layers is different. As shown in Figure 5-2, in the first half-period, the inner copper layer carries current, while the net current in the outer copper layer is zero. The magneto-motive force on the outer layer is almost zero, and then the magnetic field exerted to the outer copper layer is zero. That means there is no proximity effect loss in the outer layer. In this half period, the conduction losses in the two conduction layers are expressed as the following.

For the outer layer (layer 1),

\[ P_{t_{x-1}} = 0 \]  

(5-36)

\[ P_{t_{p-1}} = 0 \]  

(5-37)

For the inner layer (layer 2),

\[ P_{2s_{-1}} = \frac{1}{4} I_s^2 \cdot R_{dc} \cdot v \cdot F \]  

(5-38)

\[ P_{2p_{-1}} = \frac{1}{4} I_s^2 \cdot R_{dc} \cdot v \cdot F \]  

(5-39)

In the second half-period, the outer layer carries current. Though the net current in the inner layer is zero, the magnetic field exerted on the inner copper layer reaches the
maximum value. During this half period, the conduction losses in the two conduction layers are expressed as the following.

For the outer layer (layer 1),

\[ P_{1s\_2} = \frac{1}{4} I_s^2 \cdot R_{dc} \cdot V \cdot F \]  
(5-36)

\[ P_{1p\_2} = \frac{1}{4} I_s^2 \cdot R_{dc} \cdot V \cdot F \]  
(5-37)

For the inner layer (layer 2),

\[ P_{2s\_1} = 0 \]  
(5-38)

\[ P_{2p\_2} = I_s^2 \cdot R_{dc} \cdot V \cdot F \]  
(5-39)

Thus, the total conduction loss of each conduction layer in one complete period is

\[ P_{ls} = \frac{1}{8} I_s^2 \cdot R_{dc} \cdot V \cdot F \]  
(5-40)

\[ P_{lp} = \frac{1}{8} I_s^2 \cdot R_{dc} \cdot V \cdot G \]  
(5-41)

\[ P_{2s} = \frac{1}{8} I_s^2 \cdot R_{dc} \cdot V \cdot F \]  
(5-42)

\[ P_{2p} = \frac{5}{8} I_s^2 \cdot R_{dc} \cdot V \cdot G \]  
(5-43)

The equations show that the proximity effect loss in the inner layer is larger than the outer layer. In cases non-sinusoidal excitation is applied, the total conduction loss is equal to the sum of the conduction loss contributed by each harmonics.
5.4.3. Factors influencing the conduction loss

Both the harmonics and the copper thickness influence the total conduction loss. Based on an example structure the influence from the two factors is numerically analyzed.

It is assumed that a stacked structure with eight cells in the primary winding and two turns in the secondary winding operates at 1 MHz. The cell width is 10 mm, and the cell length is 60 mm. The copper thickness is equal to the skin depth of copper at 1 MHz. The space between two adjacent copper layers in the primary winding is twice the skin depth. The winding layout is shown in Figure 5-3.

The first seven harmonics of the current waveforms were extracted from the circuit simulation results and applied in the calculation of conduction loss. Figure 5-4 shows the percentage that each harmonic contributes to the conduction loss. It shows that in the LLC resonant converter, the conduction loss induced by the fundamental component is dominant.

Figure 5-5 shows the conduction loss in each layer. The inner layer of the secondary winding contributes most of the conduction loss, about 22%. The loss in the secondary winding is about 78% of the total conduction loss.

By iterating the loss calculation with the varying copper thickness, the optimum copper thickness for the primary and secondary winding is obtained. As shown in Fig. 5-6, when the copper thickness is about 60% of the skin depth, the conduction loss in both the primary and secondary winding is at its minimum; when the copper thickness is 3 times of the skin depth, the conduction loss in the primary winding is at its maximum. It also
indicates that in a one-dimensional magnetic field, increasing the copper thickness will finally reach a constant conduction loss. This is because the values of $G$ in (5-28) and $F$ in (5-33) approach to constants, when the ratio of copper thickness to skin depth increases.

![Diagram of winding layout](image)

Figure 5-3 Winding layout to numerical analysis of the conduction loss breakdown

![Graph of percentage of power loss vs. frequency](image)

Figure 5-4 Percentage of conduction loss induced by harmonics

![Graph of percentage of loss vs. layer number](image)

Figure 5-5 Conduction loss in each layer
5.5. **Loss measurement**

Since it is difficult to separate the loss in different parts of the integrated module and measure respectively, the practical method to verify the loss estimation is to measure the total loss of the module and compare the result to the sum of the calculated values for each experimental condition. Several approaches are discussed in the Appendix E for the measurement of in-circuit power loss in an integrated module.

**Case study**

Based on an integrated LLCT module with the symmetric single layer structure, the loss calculation results are compared with the experimental results measured by average power method [77].

As shown in the experimental setup in Figure 5-7, three thermo-couples were used to monitor the surface and inside temperature of the module respectively. All measurements were made at the thermal steady state. The voltage and current at the primary side of the LLCT module were recorded as shown in Figure 5-8, and used to calculate the input power with the average power method. Similarly, the output power of the LCT was measured. By subtracting the output power from the input power, the instantaneous
power loss of the LLCT was obtained. A series of experiments were made under conditions that the converter input voltage keeps 30V, while the load resistance varied to change the operating current. Corresponding to each case, the waveforms were measured and the power loss was calculated. The calculated breakdown of the LLCT loss is shown in Figure5-9. The winding loss increases significantly with the increasing current. The increase in current leads to the rise of the ac-coupled voltage. Thus the dielectric loss goes up when the current increases.

Figure 5-10 shows the power loss versus the operating current curves. The calculation results are consistent with the experimental results within the acceptable error range.
Chapter 5: Electromagnetic Loss Modeling for the Stacked Structure

Figure 5-8 the acquired voltage and current waveforms for the LLCT module with SSL structure

Figure 5-9 calculated losses distribution of the LLCT module with SSL structure
The difference between the measurement results and the calculation results is mainly due to three reasons.

**Measurement error:**

The measurement error is due to many factors. The error of the power measurement consists of the error due to the amplitude measurement and the error due to the phase measurement. With the resistive load, the relative error of the power measurement due to the phase shift error is about 1%. With the 7-bit A/D conversion, the relative error in the amplitude measurement is 1%, which leads to the relative error of the power measurement of 2%. Other measurement error can be induced during the experiment procedure. For example, the varying temperature changes the offset level of the probes. This kind of error is difficult to estimate and ignored. Combining the amplitude measurement error and phase measurement error, the estimated error of the power measurement is more than 3%. The total loss of the LLCT module is obtained by subtracting the output power from the input power. When the power measurement has the error of 3%, the error of the measured LLCT module total loss is 55%, when the LLCT efficiency is 90%. The detailed analysis of the measurement error of the electrical method is addressed in Appendix B.
Calculation error:

Due to the leakage inductance, the flux density distribution inside the ferrite core is not uniform. The ferrite core surrounding the primary winding carries more flux than that surrounding the secondary winding, as shown in Figure 5-11.

![Figure 5-11 Flux distribution inside the ferrite core](image)

With given parameters of the LLCT module, the flux in the primary side is 12% larger than that in the secondary side. It results that the core loss density in the primary side is 30% larger than that in the secondary side.

Ignoring the copper loss in interconnections leads to a smaller calculation result of the winding loss. By simply assuming that the interconnections have the same power loss density as the cells inside the core, the copper loss is proportional to the copper length. Thus, the power loss in the interconnection is about 14.6% of that in the windings.

In the integrated LLCT module with the symmetric single layer structure, the assumption of one-dimensional magnetic filed is not valid. This leads to an underestimation of the conduction loss in the calculation applying the previously proposed conduction loss model.

5.6. Summary

The losses in an integrated LLCT module are categorized into three parts according to the three different materials applied. The loss models for the three types of losses are discussed in this section. The estimation of loss in dielectric material is based on imperial equations and the material properties obtained from the manufacturer’s datasheet. The estimation of core loss is based the improved generalized Steinmetz equation, which is an
extension from the Steinmetz equation. Concerning changing rate and magnitude of flux density as the determining factors of core loss, the iGSE takes the form consistent with Steinmetz equation and can process non-sinusoidal waveforms. The iGSE assumes uniform distribution of magnetic flux in the core. The estimation of conduction loss is based on the orthogonal principle and the assumptions of one-dimensional electromagnetic field distribution along the winding width direction and linear current distribution along the winding length direction. Since it is not possible to measure the in-circuit losses in the three kinds of materials separately, the total loss is be measured and compared with the estimation value. If the estimation values agree with the measured values at various operational conditions, the loss models are considered valid. The electrical method and thermal method are introduced to measure the in-circuit loss. Applied in a specific case, the electrical method provides results consistent with the estimation results. However, the measurement error existing in the electrical method is large. The improve method for the in-circuit loss measurement of the integrated LLCT module is proposed and detailed in Appendix E.
Chapter 6: Thermal Management of Integrated Passive Modules

6.1. Thermal constraint in integrated power passives

Associated with the increasing power density, loss density increases. Although the loss density in the passive component is much lower than that in power semiconductor devices, the lower thermal conductivities of most materials applied in passive components leads to high temperature gradient inside the component. The thermal issues limit the increase of power density. The maximum allowed temperature inside an integrated module is one of the most important design criteria.

Although the surface temperature of a passive module can be controlled by the external cooling conditions, the temperature rise inside the module is determined by the loss distribution, the material properties and the structure. For a practical design, the structure is basically determined by the considerations of electromagnetic function, and the materials are generally selected for their electromagnetic characteristics to meet the required parameter values. When the structure and materials are fixed, the loss distribution is determined by the circuit operation, which is normally also fixed. Thus, it is difficult to improve the thermal performance of an existing design without changing the structure or materials. The most common approach to reduce the inside temperature gradient is to increase the dimensions, which reduces the loss density and increase the heat dissipation area, however, sacrifices the power density. It is clear that in cases where the thermal constraint is the dominant design criterion, the heat dissipating ability of a module is directly related to the power density. In other words, by improving the heat dissipating ability, the power density can be improved. Being aware of this, the heat extraction technology is developed.

6.2. Thermal performance improvement with heat extraction

6.2.1. Heat extraction theory

Passive integrated power electronics modules (IPEMs) containing ceramics as a large part of its volume typically has a low loss density, as well as low average thermal
conductivity. These structures now prove to be a favorable initial application area for the concept of embedded heat extractors. To prevent self-heating of the extractors by the excitation of the high magnetic and electric field strengths at the high frequency of operation in the passive IPEM, the embedded heat extracting material has to be electromagnetically inactive. As shown in Figure 6-1, aluminum nitride is chosen and the embedded material oriented parallel to the magnetic field in the ferrite. It is clear that an optimum volume of the heat extracting material reduces the total electromagnetic energy processed, but more efficient heat removal allows higher electromagnetic stresses and higher losses per unit volume in the remaining material. Figure 6-2 shows how the core loss increases with the increasing volume occupied by the heat extractors. Figure 6-3 illustrates the application of the heat extractor improves the capability of higher power density.

![Figure 6-1 Heat extraction in two directions](image1)

![Figure 6-2 Core loss vs. relative volume of heat extractors](image2)
In cases where the thermal constraint is the main design limitation, the heat extraction theory is suitable and will improve the power density. In practical designs, a tradeoff exists between high efficiency and high power density; this can be controlled by different criteria.

### 6.2.2. Theoretical analysis based FEM

The factors influencing the thermal performance of the module with embedded heat extractors were investigated by J. Dirker using a three-dimensional model calculated with finite element method [78].

In his research, both geometry factors and thermal factors are considered.

*Geometrical considerations for the embedded heat extractors:*

- Relative volume
- Physical size
- Cross-sectional shape of inserts
- Spacing offset
- Depth (third dimension length)

*Thermal considerations for the embedded heat extractors:*

- Volumetric heat-generation in the heat-generating material [W/m³]
- Thermal conductivity of the heat-generating material [W/mK]
- Thermal conductivity of the heat-extracting material [W/mK]
• Thermal contact resistance between heat extractor and heat-generating medium [m²K/W]
• Ambient temperature [K]

Based on the FEM calculation results, the following conclusions are drawn:

• For heat extractors with small ratio of the cross-sectional area to the length, the advantage of optimizing the heat extraction geometry becomes less. In other words, the cross-sectional shape does influence the thermal performance much in these cases.

• If the fraction of the domain occupied by the heat extractor (the relative volume) is below or in the regain of 10%, the optimum cooling insert geometry is that of a continuous flat plate.

• The thermal contact resistance on the heat flow path affects the overall thermal performance much.

By applying the continuous flat plate as the structure of the heat extractors, shown in Figure 6-4, the two-dimensional thermal field can be used to investigate the performance [79].

(a) Heat extractor as a continuous flat plate
(b) Two-dimensional representative domain

Figure 6-4 the heat extractor structure for power electronics applications and its representative domain

Obtained from the calculation with practical material properties and the measured thermal contact resistance, the power density improvement due to the heat extractors occupying 10% of the volume is shown in Figure 6-5. About 200% power density improvement is expected.

![Figure 6-5 Power density improvement vs. thermal contact resistance](image)

Figure 6-5 Power density improvement vs. thermal contact resistance
6.2.3. Experimental verification

Experiment is set up to verify the calculation result (refer to Appendix C). Five layers of ferrite with a height of 4.5 mm and a width 5 mm are stacked together to form a closed magnetic loop. Four ferrite sections are used for each level and placed tightly in contact with each other to reduce the influence of the air gaps that might have caused uneven magnetic field distribution. Aluminium nitride slices with a width of 5 mm and a height of 0.5mm are used as heat extraction sections when required. In total, 6 aluminium nitride layers can be introduced into the ferrite stack when needed.

The magnetic material loop structure is placed into contact on two sides with identical aluminium heat sinks of which the temperature are measured with 2 embedded thermal couples each. A 12 V DC fan at a fixed position relative to the heat sinks is used to cool the heat sink surfaces. Horizontal bolts are used to maintain pressure between the magnetic loop structure and the heat sinks. Care is taken to apply the same amount of force in all experiments.

In cases where aluminium nitride layers are present, good thermal contact is maintained between the aluminium nitride and ferrite layers by applying uniform pressure from above. Pre-weighed mass pieces are placed on to the set-up for this purpose. All surfaces except those of the heat sinks are thermally insulated to resemble adiabatic boundaries as closely as possible. The heat transfer between the test section and the other ferrite cores is ignored.

On the opposite side of the magnetic loop, litz wire (100/44) is wound around the magnetic core assembly forming a winding with 9 (or 10) turns. The whole structure functioned as an inductor. Excited with a sinusoidal waveform, the induced magnetic field uniformly generated heat within the ferrite core. A mica capacitor, chosen for its low loss characteristics, is connected with the inductor to form a series resonator. Operating at the resonant frequency, high voltage across the inductance and large flux density in the core is obtained. To effectively heat the ferrite, the operating frequency was set 1MHz. The experimental setup is shown in Figure 6-6.

By keeping the external cooling condition unchanged, the temperature on the heat sink surface is regarded a monotonic function of the heat flow at steady state. Since the power
loss generated in the core is not accurately measured, the heat sink temperature is taken as the reference corresponding to the power loss. Varying the input power yields different heat sink temperature and the maximum temperature in core at steady state. By irritating the experiment with and without the heat extractors, the thermal performance improvement is illustrated with the curves shown in Figure 6-7.

Figure 6-6 Experiment to verify the calculation result about the power density improvement by heat extactor occupying 10% volume
By applying the theoretical linear relationship between the maximum core temperature different and the heat sink temperature, the capability of power density increase of 187% is obtained [80].

Considering the possible error induced during the experiment procedure, the experimental results match the calculation results well. The application of heat extractors is promising.

### 6.3. Power density improvement with heat extraction

To apply heat extractors in power electronics module, the heat extractor plate should be parallel to the magnetic flux to avoid influencing the electromagnetic performance. The structure shown in Figure 6-8 can embed heat extractors without change the electromagnetic performance, but the inside filed distribution changes due to the redistribution of the magnetic flux. The induced fringing effect may even influence the conduction loss. If the heat extractors occupy a small relative volume, the change of the magnetic field is neglected.

The embedding of heat extractors also changes the distribution of the thermal field. Originally regarded as a uniform heat source, the core is now several individual heat sources.
sources with higher heat generation. Assuming the losses in the conductor and dielectric keep unchanged, the thermal performance of the integrated passive module with embedded heat extractors can be investigated in simulation.

The thermal performance in the integrated module with the embedded heat extractors is influenced by several factors. To investigate how heat extractors influence the performance of the module, a thermal model is built based on the structure shown in Figure 6-8. Several assumptions are made to simplify the model. In the windings, thin dielectric layers and thin copper layers interleave and stack tightly within an encapsulant package. The winding and the encapsulant together are regarded as a uniform heat-generating block with the same thermal conductivity as the dielectric material, 0.2 W/mK. This approximation changes the temperature gradient inside the winding block, but has only a slight influence on the module performance as it relates to heat extractors, because heat extractors do not contact the inner part of the winding block.

The ferrite core is regarded as a uniform heat-generating medium. The thermal contact resistance between the core and the winding block is assumed to be the same as that between the heat extractors and the winding block, and is termed as the internal thermal contact resistance, $R_{in}$. The thermal contact resistance between the core and the heat sink is assumed to be the same as that between the heat extractors and the heat sink, and is termed as the external thermal contact resistance, $R_{ext}$.

As shown in Figure 6-8, the module length is much larger than its width and height. Thus, a two-dimensional heat transfer can be assumed on the cross section. Figure 6-9 shows the diagram of heat paths on the cross section. In the module, there are two heat
sources: the winding block and the magnetic core. By applying the superposition principle, the amounts of heat flux generated by these two heat sources can be studied separately, and the results can be combined linearly. For the heat generated inside the winding block, both the core and heat extractors are conduction channels, through which heat can be conducted to the heat sink. The internal and external thermal contact resistance and the equivalent thermal resistance of the combination of the core and heat extractors are the barriers on this heat path. For heat generated in the core, the winding block functions as a conductor connecting the core and heat extractors. The internal thermal contact resistance functions twice on one of the heat paths. Because of the different heat distribution, the equivalent thermal resistance of the combination of core and heat extractors is different from that for heat coming from inside the winding block. The equivalent structures of both cases are shown in Figure 6-9, where the core and the winding block are replaced by equivalent spot heat sources, and the corresponding equivalent thermal resistance is applied.

The relative volume of heat extractors influences both the core loss and the thermal resistance on the heat paths, while the interfacial thermal resistance between the core and heat extractors, $R_{Fe-AlN}$, does not have much impact on the thermal performance. The interfacial thermal resistance becomes less important in cases where the core thickness is much smaller than the width and height of the module, because a small core thickness leads to a small area of contact with the heat extractors. Most of heat directly dissipates to the heat sink without passing the heat extractors. The amount of heat conveyed between the core and heat extractors is also related to the heat-generation distribution and detailed boundary conditions.

In some applications, the loss in the winding block will be much higher than that in the core. If the core thickness is much smaller than the cooling area contacting the heat sink--for example, if a planar core is applied--the thermal resistance between the core and heat extractors, $R_{CE}$, is so large that the heat conduction between the core and heat extractors is negligible. For such cases, it can be assumed that no heat transfer between the core and the heat extractors, and the model in Figure 6-10 (a) can be solely applied. Assuming the value of $R_{CE}$ is infinite, the overall equivalent thermal resistance of the module can be expressed by the following:
where $k_n$ is the ratio of the heat-extractor thermal conductivity to the core thermal conductivity; $R_s$ is the sum of the internal and external thermal resistances, $R_{\text{core}}$ is the thermal resistance of the core without embedded heat extractors; and $\alpha$ is the relative volume of heat extractors. The value of overall equivalent thermal resistance $R_o$ varies within the range $R_s + \frac{R_{\text{core}}}{k_n} \leq R_o \leq R_s + R_{\text{core}}$. If thermal adhesive with high thermal conductivity is applied in the internal and external interface, the value of $R_s$ can be small; then, the overall equivalent thermal resistance is approximately inversely proportionate to the relative volume of the heat extractors.

In some applications, the core loss is the dominant loss of the module, and the core thickness is comparable to the width and height of the module. Then, the model in the Figure 6-10 (b) can be solely applied. Ignoring the heat passing through the winding block, the investigations for this model are carried out with numerical methods as stated in the previous section.

However, in most power electronics applications, with the optimal design goal of high power density, the core loss is comparable to the winding loss and the dielectric loss, and core thickness is not negligible. For these cases, such as the integrated L-L-C-T module presented in Section III, numerical methods must be applied in the investigation.

![Figure 6-9 Diagram of heat pathes on cross section](image-url)
Chapter 6: Thermal Management of Integrated Passive Modules

(a) for removing heat in the winding block and (b) for extracting heat from the core.

Figure 6-10 Equivalent thermal structure

The above analysis shows that the thermal performance and internal temperature rise of a specific integrated LLCT module with embedded heat extractors is a combined result of several coupling factors. For a physical prototype, the internal temperature is influenced greatly by the loss distribution.

Based on the structure of the prototype shown in Figure 6-11, a thermal model is built in the simulation software, I-DEAS.

Figure 6-11 Prototype structure used to build thermal model

The simulation model consists of nine core layers and ten aluminum nitride layers. These two kinds of layers interleave with each other. By changing the thickness of the layers, the relative volume of heat extractors can be controlled. The thermal resistance on the interfaces is determined by materials and processing technologies. The internal
thermal contact resistance is estimated by calculating the thermal resistance between the winding block and the inner surface of the core and heat extractors. The material in between can be either air (1) or the encapsulant (2) with a thermal conductivity of 1.5 K/mW. The material in between the heat sink and the outer surface of the module can be either thermal adhesive (3) or a gap pad (4). The interfacial thermal resistance between the core and heat extractors is obtained from the measurement results (see Appendix D). By changing the surface roughness and the applied pressure, the interfacial thermal resistance can be modified. The values of the thermal contact resistance and the corresponding interfacial thermal resistance are listed in Table 3. In the simulation, the temperature on the heat sink surface is fixed at 25°C. The two terminal surfaces are set as thermal isolation.

![Figure 6-12 Thermal model in I-DEAS](image)

Since the varying relative volume of the heat extractor changes the total loss as well as the loss distribution in the core, the module with the best thermal-handling ability may not be the one with the minimum temperature rise. Thus, the overall equivalent thermal resistance is not representative of the performance of a module. The maximum temperature difference between the inside of the module and the heat sink surface, $\Delta T_{\text{max}}$,
is accommodated to represent the performance of the module. The lower maximum temperature difference refers to better performance and also a larger power-processing capability.

Four factors influencing the performance of modules with heat extractors—the internal and external thermal contact resistances, the interfacial thermal resistance between the core and the heat extractors, and the relative volume of the heat extractors—were set as four respective variables in four different simulation groups.

For each simulation group, the maximum temperature difference in the module without a heat extractor was first simulated. This result was compared with the other results in the group. The module performance was then presented with normalized values of the maximum temperature difference.

Simulations were based on two sets of materials, as listed below. Figure 6-13 shows that by applying a small number of heat extractors, the maximum temperature difference decreases. The optimal value of the relative volume of heat extractors varies with different interface thermal properties. For the module presented, with the heat extractors occupying 15% of the total volume, the maximum temperature difference could be reduced to 85% with efficiency reduction of about 1%, as shown in Figure 6-14. In other words, with the application of heat extractors, the power throughput and the power density of the module can increase to 118% while the maximum temperature remains unchanged, assuming that the increased power loss is proportionately distributed into the winding block and the core.

To eliminate the influence of the internal and external contact thermal resistances, their values are set as $2 \times 10^{-4}$ m$^2$K/m. The relative volume of heat extractors is 10%. The maximum temperature difference in the module without a heat extractor is set as the comparison reference. Figure 6-15 shows that the interfacial thermal resistance between the core and heat extractors has only a slight influence on the thermal performance of this module.

The relative volume of heat extractors for this simulation is set at 10%, and the interfacial thermal resistance between the core and heat extractors is $2 \times 10^{-4}$ m$^2$K/W. With the changing values of $R_{in}$ and $R_{ext}$, the maximum temperature difference is obtained for
modules with and without heat extractors, respectively. The two sets of values are compared to demonstrate the extent to which the maximum temperature difference could be reduced by applying the heat extractors. Figure 6-16 shows that the smaller the contact thermal resistance, the greater that improvement contributed by heat extractors.

A comprehensive design approach including the experimental verification is presented in the next chapter.

Figure 6-13 Normalized maximum temperature difference

Figure 6-14 Efficiency vs. relative volume of heat extractors
Figure 6-15 Normalized maximum temperature difference vs. the thermal resistance between the core and heat extractors

Figure 6-16 Normalized maximum temperature difference vs. internal and external resistances

6.4. Summary

In high loss density integrated module, the thermal constraint limits the increase of power density. The improvement on the heat dissipation will lower inside temperature rise and reduce the maximum temperature inside the module, and further increase the power density without overheat the module. Based on this concept, the heat extraction technology is introduced to power electronics components. By replacing part of the low thermal conductivity material with high thermal conductivity electromagnetic inactive material in the integrated passive module, the heat dissipation ability of the module is
improved. Though reduction of effective electromagnetic material leads to higher power loss, the temperature rise inside the module reduces due to the improved heat dissipation ability.

The heat extraction concept was verified by calculation and experiment in a sole heat generating medium, the magnetic core. From the experimental investigation it is found that the presence of aluminum nitride heat extraction layers decreases the maximum temperature rise within the ferrite core material. The experimentally obtained performance improvement compared well with the theoretical expectation.

Based on the success, the concept is expanded into the integrated passive modules. The theoretical analysis and simulation results show promising results of the application of heat extraction. For the specific LLCT module, the application of heat extractors (occupying 10% of the original core volume) is expected to reduce the internal temperature rise to 85% of the original value, according to the simulation results with the normal interfacial conditions. The experimental verification will be presented in the next chapter.
Chapter 7: Experimental Evaluation

7.1. Prototype design

The design approach of integrated passives with the stacked structure has been proposed in Chapter 4. The loss models for magnetic, dielectric and conductive materials have been detailed in Chapter 5. The thermal management and the induced benefit of power density improvement have been presented in Chapter 6. Based on the above research, prototypes are designed and implemented for a specific application, and tested to verify the previous work.

The integrated LLCT module is designed for a 1 MHz 500W front end converter in a DPS testbed. The converter is to transfer 400V input to 48V output. The power stage has half bridge structure and thus the voltage exerted to the LLCT module is 200 V. The integrated LLCT module provides the combination of a series resonator and a transformer with low magnetizing inductance and center-tapped secondary windings. The electrical parameter values of the LLCT module are based on the design of the converter, as listed in Table 7-1 [81].

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant Inductance, $L_r$</td>
<td>0.4 uH</td>
</tr>
<tr>
<td>Resonant Capacitance, $C_r$</td>
<td>50 nF</td>
</tr>
<tr>
<td>Magnetizing Inductance, $L_M$</td>
<td>20 uH</td>
</tr>
<tr>
<td>Turns ratio, $n_p : n_s : n_s$</td>
<td>4:1:1</td>
</tr>
<tr>
<td>Resonant Frequency, $f_r$</td>
<td>1.1 MHz</td>
</tr>
</tbody>
</table>

The detailed current and voltage waveforms are obtained from circuit simulation. Figure 7-1 shows the simulation diagram, in which ideal models of switches and diodes are applied. The voltage and current waveforms are shown in Figure 7-2. The secondary current is represented by the combination of current in the two center-tapped windings. The voltage across secondary windings is used to calculate magnetic flux density. The Fourier components of the current waveforms, shown in Figure 7-3, are used to estimate...
conduction loss. The voltage across the resonant capacitance is obtained for dielectric loss estimation. The values of the electrical parameters are listed in Table 7-2.

Figure 7-1 Circuit diagram of the 1MHz LLC resonant converter

(a) Primary current, $I_{pri}$ and voltage across resonant capacitor $V_{Cr}$
Chapter 7: Experimental Evaluation

(b) Voltage and current on one of the center-tapped secondary winding

Figure 7-2 Simulation waveforms

(a) Fourier components of primary current

(b) Fourier components of secondary current (reversing the current in one of the windings)

Figure 7-3 Fourier components of $I_{pri}$ and $I_{sec}$
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Table 7-2 Electrical parameter values from circuit simulation

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating Frequency, $f$</td>
<td>1 MHz</td>
</tr>
<tr>
<td>Voltage across secondary windings (dc), $V_{sec}$</td>
<td>96 V</td>
</tr>
<tr>
<td>Voltage across resonant capacitor (rms), $V_{Cr}$</td>
<td>10.6 V</td>
</tr>
<tr>
<td>Primary current (rms), $I_{pri_rms}$</td>
<td>3.12 A</td>
</tr>
<tr>
<td>Primary current, 1$^{st}$ harmonic(peak), $I_{pri_1st}$</td>
<td>4.32 A</td>
</tr>
<tr>
<td>Primary current, 3$^{rd}$ harmonic(peak), $I_{pri_3rd}$</td>
<td>0.787 A</td>
</tr>
<tr>
<td>Primary current, 5$^{th}$ harmonic(peak), $I_{pri_5th}$</td>
<td>0.345 A</td>
</tr>
<tr>
<td>Primary current, 7$^{th}$ harmonic(peak), $I_{pri_7th}$</td>
<td>0.178 A</td>
</tr>
<tr>
<td>Secondary current (rms), $I_{sec_rms}$</td>
<td>12.0 A</td>
</tr>
<tr>
<td>Secondary current, 1$^{st}$ harmonic(peak), $I_{sec_1st}$</td>
<td>16.7 A</td>
</tr>
<tr>
<td>Secondary current, 3$^{rd}$ harmonic(peak), $I_{sec_3rd}$</td>
<td>2.73 A</td>
</tr>
<tr>
<td>Secondary current, 5$^{th}$ harmonic(peak), $I_{sec_5th}$</td>
<td>1.11 A</td>
</tr>
<tr>
<td>Secondary current, 7$^{th}$ harmonic(peak), $I_{sec_7th}$</td>
<td>0.562 A</td>
</tr>
</tbody>
</table>

As stated in Chapter 4, the turns number of the LLCT module is normally set as the minimum integer multiple of the turns ratio. In this case, there are four turns in primary winding and two turns in the secondary winding. Due to the operation frequency, the core material is chosen as 3F4 with the permeability of 900. The leakage layer material is made by ferrite polymer composite (FPC) films with the permeability of 8, provided by EPCOS. The ceramic dielectric material is N1250 with permittivity of 174. Corresponding to the used materials, the processing technologies would be electroplating, photo-masking, chemical etching, laser cutting and soldering. The dimension tolerance is set 0.5 mm. The insulation material is 0.2mm Kapton tape. According to the processing technology and the practical dimensional tolerance, the minimum distance between two adjacent conductors is set 0.5mm for insulation (refer to Appendix D). The material properties and dimensional tolerances are listed in Table 7-3.

With the known material properties and fixed dimensional tolerances, there are only two independent design variables, the magnetic flux $B$ and the aspect ratio of the LC cell
\( k_a \). The aspect ratio of the LC cells relates to the dimensions of copper layer and thus the average current density. In the design calculation, the flux density \( B \) has 20 values varying from 5mT to 200mT; the cell aspect ratio \( k_a \) has 20 values varying from 0.005 to 0.5. Four practical issues are considered as the design constraints.

1. \( L_{m\text{,cal}} < L_{m\text{,req}} \). Since there are only four turns in the primary winding, the maximum obtainable magnetizing inductance would be small. If the calculated maximum magnetizing inductance is smaller than the required value, the design is void. Since the magnetizing inductance is adjusted by changing the air gap length, when the calculated air gap is smaller than the practical value of the minimum air gap determined by processing technologies, the design is valid.

2. \( L_{int} > L_r \). If the inherent leakage inductance is larger than the required resonant inductance, the design is void.

3. \( B_{leakage} > B_{sat} \). If the leakage flux in the leakage layer is larger than the saturation flux density of the leakage material, the design is void.

4. \( W_{core} < W_{core\text{,min}} \). If the calculated core thickness is smaller than the minimum core thickness that can be machined, the design is void.

### Table 7-3 Properties of materials applied in the integrated LLCT module

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core material</td>
<td>3F4 (900)</td>
</tr>
<tr>
<td>Dielectric material</td>
<td>N1250 (174)</td>
</tr>
<tr>
<td>Leakage layer material</td>
<td>FPC (8)</td>
</tr>
<tr>
<td>Heat extractor material</td>
<td>AlN</td>
</tr>
<tr>
<td>Conductor material</td>
<td>Copper</td>
</tr>
<tr>
<td>Insulation material</td>
<td>Kapton tape, 0.2 mm</td>
</tr>
<tr>
<td>Core saturation flux density</td>
<td>0.2 T</td>
</tr>
<tr>
<td>Leakage material saturation flux density</td>
<td>0.2 T</td>
</tr>
<tr>
<td>Thickness of dielectric material</td>
<td>0.15 mm</td>
</tr>
<tr>
<td>Dimension tolerance</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>Minimum core thickness</td>
<td>1.5 mm</td>
</tr>
</tbody>
</table>
The design maps are presented with respect to several characteristics. The design map with respect to power density is calculated without applying any design constraint and shown in Figure 7-4. It shows that increasing flux density and reducing aspect ratio lead to the increase of power density. The possible maximum power density is 1029 W/in$^3$.

Figure 7-4 Design map with respect to power density without constraints
However, the designs without constraints may not be all valid. By applying the four constraints, the invalid designs are filtered. Within the 400 calculated designs, there are 2 designs violating constraint 1, 260 designs violating constraint 2 and 3 respectively, and 253 designs violating constraint 4. As shown in Figure 7-5, constraint 2 and constraint 3 lead to the same valid design range. When the winding height increases, the inherent leakage inductance approaches the required resonant inductance and the required height of the additional leakage layer is reduced to a small value, which leads to saturation with even a small magnetic flux density. In the current application case, the dimensional increment between two continuous calculation steps is larger than the difference between the boundary dimensions of constraint 2 and 3. Thus, the applications of constraint 2 and constraint 3 get the same results of the valid design range. If the dimensional increment reduces to the small value, constraint 2 and constraint 3 should identify different valid design ranges. Design results under constraint 4 are shown in Figure 7-5 (c). By comparing the design results with the three design constraints, it shows that constraint 2 and 3 is the dominant constraints for the design specifications. Here, the thermal issue is not considered.

Figure 7-6 shows the design maps with respect to power density, loss, inherent leakage and core thickness, as well as the contour of the power density. With all constraints applied, the maximum obtainable power density reduces to 740 W/in$^3$. Finally, the design with flux density of 47 mT and LC cell aspect ratio of 0.167 is selected for implementation. The selection is based on several considerations. To better satisfy the assumption of one-dimensional electromagnetic field distribution stated in Chapter 4, the small cell aspect ratio is required. As shown in the design map, the selected design has the inherent leakage inductance close to the resonant inductance. By adjusting the dimensions during assembly, the inherent leakage inductance can meet the required value of the resonant inductance, and thus the additional leakage layer is not needed in the module. The power density of the selected design is about 524 W/in$^3$, which is larger than the benchmark of 480 W/in$^3$ [17]. Compared to the possible maximum power density shown in the design map, the selected design is conservative but having space to experimentally achieve higher power density by increasing throughput power in the in-circuit test. The estimated total loss of the selected design is about 5.94 W, including
conduction loss of 0.94 W, core loss of 4.84 W and dielectric loss of 0.16 W. It corresponds to an efficiency of 99%, which is acceptable in practical applications. The core loss is estimated by Steinmetz equation. The conduction loss is estimated under the assumption of one-dimensional electromagnetic and doesn’t take harmonic into account. The simplified loss models underestimate the loss. Considering the low thermal conductivity of ferrite, about 5 W/mK, the amount of loss would be able to create sufficient temperature gradient inside the module to verify the heat extraction theory (refer to chapter 6).

The dimensions obtained from the selected design are rounded to practical values. The detailed dimensions of each part of the integrated LLCT module are listed in Table 7-4. The copper thickness is obtained from the optimal design with minimum conduction loss, as stated in Chapter 5. To verify the heat extraction theory, heat extractors are applied in the LLCT module. Based on the selected design, 10% of the core volume will be replaced by the heat extractors. Heat extractors have the thickness of 0.5mm and the same cross-sectional dimensions.

Two constructed prototypes are presented in Figure 7-7. Prototype I is without heat extractors and prototype II is with heat extractors embedded in the core (refer to chapter 6, the volume ratio between heat extractors and cores is 1:9). The implementation of the prototypes is presented in Appendix D. Prototype I is implemented and evaluated firstly, and then the magnetic core is replaced by the core with embedded heat extractors, while the winding keeps unchanged to construct prototype II. Besides the evaluation of electrical performance, the internal temperature rise of the two prototypes is compared to evaluate the improvement of thermal performance by the application of heat extractors.
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Constraint 1

Invalid design due to constraint 1

(a) Design map obtained with design constraint 1 only

Constraint 2

Invalid design due to constraint 2

(b) Design map obtained with design constraint 2 only
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(c) Design map obtained with design constraint 3 only

(d) Design map obtained with design constraint 4 only
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(e) Design map obtained with all design constraints

Figure 7-5 Designs obtained with various design constraints

(a) Design map with respect to power density
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(b) Contour of the design with respect to power density

(c) Design map with respect to loss
(d) Design map with respect to inherent leakage inductance

(e) Design map with respect to the core thickness

Figure 7-6 Design maps with all constraints
Table 7-4 Design parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core width</td>
<td>2 mm</td>
</tr>
<tr>
<td>Cell width</td>
<td>10 mm</td>
</tr>
<tr>
<td>Cell length</td>
<td>60 mm</td>
</tr>
<tr>
<td>Copper width</td>
<td>9.5 mm</td>
</tr>
<tr>
<td>Copper thickness in primary</td>
<td>40 μm</td>
</tr>
<tr>
<td>Copper thickness in secondary</td>
<td>55 μm</td>
</tr>
<tr>
<td>Winding window height</td>
<td>4 mm</td>
</tr>
<tr>
<td>Winding window width</td>
<td>11 mm</td>
</tr>
<tr>
<td>Module width</td>
<td>15 mm</td>
</tr>
<tr>
<td>Module length</td>
<td>60 mm</td>
</tr>
<tr>
<td>Module height</td>
<td>16 mm</td>
</tr>
<tr>
<td>Power density</td>
<td>524 W/in³</td>
</tr>
</tbody>
</table>

7.2. Evaluation of construction and design

The design maps with respect to each of the four constraints demonstrated in Figure 7-5 show that the stacked structure has least limitation from the design constraint 1 and largest limitation from the design constraint 2 and 3. Since the turn’s number in the primary is small, only four, the magnetizing inductance has a small value. This feature is suitable to the application in the LLCT module, which requires small magnetizing inductance. The inherent leakage inductance leads to the largest design limitation, because the stacked LC cells provide large space for the leakage flux. The minimum acceptable core thickness also leads to certain design limitation. With the improve material and processing technologies, thinner core may be valid and the valid design range can be extended.

The frequency response curve is obtained by small signal test to verify electromagnetic performance of the prototypes. The resonant inductance and capacitance is fixed after the prototype is built, while the magnetizing inductance is adjustable through the air gap. If the secondary winding is open, the impedance measured in the primary is related to the
magnetizing inductance only, as shown in Figure 7-8. The first resonance is due to the series resonance of the magnetizing inductance and the resonant capacitance. The second resonance is due to the parallel resonance of the magnetizing inductance and the parallel parasitic capacitance reflected to the primary winding. With one of the two center-tapped secondary windings shorted, the impedance curve measured in the primary side is related to the leakage inductance between the primary winding and the corresponding secondary winding. From the frequency response curves shown in Figure 7-9 and 7-10, it is observed that the obtained inherent leakage inductance is different in different secondary windings. It is due to the asymmetry of the multi-turn windings implemented in the stacked structure. The secondary winding has two turns. The inner turn and the outer turn are in different positions and provide different paths for leakage flux. Due to the same reason, the parasitic capacitance between each of the two secondary winding turns and the primary winding is different. The frequency response curves of prototype II are shown in Figure 11, 12 and 13. The design values and measured values of the electrical parameters are listed in Table 7-5. The magnetizing inductance can be adjusted by a change in air gap length to meet the requirement. The inherent leakage inductance is sensitive to the dimensional tolerance induced during the processing procedure. It is also sensitive to the lead inductance in the secondary side. Because the turn’s ratio is 4:1, a small lead inductance in the secondary side could be a large value after it is reflected to the primary side.

(a) Prototype I without heat extractor
(b) Prototype II with heat extractors

Figure 7-7 Prototypes

Figure 7-8 Frequency-response curve of prototype I measured at the primary with the secondary open
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Figure 7-9 Frequency-response curve of prototype I measured at the primary with one of the center-tapped winding (terminals X and Y) in the secondary short-circuit

Figure 7-10 Frequency-response curve of prototype I measured at the primary with one of the center-tapped winding (terminals Y and Z) in the secondary short-circuit
Figure 7-11 Frequency-response curve of prototype II measured at the primary with the secondary open

Figure 7-12 Frequency-response curve of prototype II measured at the primary with one of the center-tapped winding (terminals X and Y) in the secondary short-circuit
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Figure 7-13 Frequency-response curve of prototype II measured at the primary with one of the center-tapped winding (terminals Y and Z) in the secondary short-circuit

Table 7-5 Design values of electrical parameters and experimental results

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Design values</th>
<th>Prototype I</th>
<th>Prototype II</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant capacitance, $C_r$</td>
<td>50 nF</td>
<td>47.1 nF</td>
<td>45.3 nF</td>
</tr>
<tr>
<td>Magnetizing inductance, $L_M$</td>
<td>20 μH</td>
<td>14.1 μH</td>
<td>17.3 μH</td>
</tr>
<tr>
<td>Resonant inductance, $L_r$</td>
<td>0.4 μH</td>
<td>447 μH (inner)</td>
<td>315 μH (inner)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>493 μH (outer)</td>
<td>353 μH (outer)</td>
</tr>
<tr>
<td>Parasitic capacitance</td>
<td>-</td>
<td>30.6 pF</td>
<td>32.5 pF</td>
</tr>
</tbody>
</table>

7.3. Experimental setup and results

The purpose of the experimental setup is to determine the in-circuit electrical, loss and thermal characteristics of the prototypes. A 1 MHz 500W LLC resonant converter is constructed to evaluate the prototypes under specific conditions. The schematic diagram of this converter is shown in Figure 7-1 and the schematic of the experimental setup is shown in Figure 7-14.
To calculate in-circuit conduction loss, long leads are used in the secondary side for installing the current probe to measure the current in the secondary winding. The resulting lead inductance is reflected to the primary side and has a large value due to the turns ratio of 4. With the long lead in the secondary side, the overall resonant inductance increases appreciably. To keep the resonant frequency at around 1 MHz, a MICA capacitor (20 nF) is in series with the primary winding. To obtain appropriate converter operation, the magnetizing inductance is adjusted to 15 \( \mu \)H. The measured in-circuit waveforms will be used to calculate the losses and to verify the loss models, illustrated in chapter 5. The in-circuit loss measurement is made by a thermal method with high accuracy (the estimated measurement error is 4.8%, refer to Appendix E). The thermal condition is represented by the monitored internal temperature rise under the specific circuit operational conditions. The shape of the prototypes is a hexahedron with four flat side surfaces and two end surfaces, where terminals locate. Heat sinks are attached on the four side surfaces to cool down the module, as illustrated in Figure 7-14 (b). Four heat flux sensors are inserted between the module surfaces and the heat sinks. With the two ends surfaces covered by adiabatic materials, it is regarded that all heat dissipates through heat flux sensors to heat sinks. Thus, the total loss can be measured at thermal steady state.

(a) Electrical and converter loss measurement
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Figure 7-14 Schematic of experimental test setup for the integrated LLCT prototypes showing electrical, loss and temperature measurements

The inside maximum temperature is measured by the thermocouple placed in the center of the core. The difference between the maximum inside temperature and the average heat sink temperature is referred to as the internal temperature rise, which is an indicator of the overall thermal conditions of the prototypes and also an importance factor of power density identification. To eliminate the variation of temperature on the four heat sinks, the heat sinks have the same material and dimensions. The prototype is placed in a wind tunnel to obtain balanced external cooling condition. Since the usage of the wind tunnel increase the distance between the prototype and converter circuit, planar leads are applied to eliminate the induced lead inductance. However, the application of planar leads increases the intra-winding capacitance, which leads to high frequency resonances on the waveforms. Thus, the wind tunnel shown in Figure 7-15 is used only for experiments of power density identification, where the balanced external cooling condition is desired. The experimental setup is shown in Figure 7-16.

With the measurement setup illustrated in Figure 7-14 (a), the waveforms are measured at terminals of prototype I operating at 834 kHz, and shown in Figure 7-17. As observed, high frequency resonances are observed on the current and voltage waveforms. It is due to the intra-winding capacitance induced by the planar leads. The secondary current differs in different half operation period. This is because the difference of the inherent leakage inductance in the secondary windings causes different equivalent resonant inductance in different half periods. The operation of the 1 MHz LLC resonant current is sensitive to small variance of the resonant inductance (several dozens of micro henrys). In this practical case, the unbalanced inherent leakage inductance in the secondary
windings is compensated by the inductance induced by the external leads in the secondary side.

Figure 7-15 Experimental layout for power density identification of the integrated LLCT prototypes

Figure 7-16 Experimental setup for evaluating the integrated LLCT prototype
With the modified resonant inductance, prototype I and II are tested respectively at around 955 kHz. The waveforms are shown in Figure 7-18. The presented secondary current waveform has been processed by reversing the original waveform for a half period. The current waveforms in Figure 7-18 show that the current is evenly distributed into the two secondary windings.

Figure 7-17 Original waveforms measured from prototype I without modified external leads

Waveforms of 480W output

(a) Input and output waveforms of measured at prototype I
7.4. **Experimental and calculation data**

The experimental results demonstrated in the above, including the small signal test and in-circuit test with rated output power, show evidence that both prototype I and II function well under the rated operational conditions with the designed performance.

To investigate the characteristics of the stacked structure, more experiments are made to obtain practical data for the analysis of the accuracy of loss models, the power density and the improvement brought by the heat extraction technology.

Five experiments are made with prototype I under different circuit operational conditions. In experiment 1, 2 and 3, the testing LLC converter has the constant load resistance of 5 ohms and various output voltage. In experiment 4 and 5, the testing converter has the constant load resistance of 2.5 ohms and various output voltage. The frequency in the five experiments is controlled to be close to 950 kHz. Since the devices in the control circuit is temperature dependent, the variance of frequency is within a range of 50 kHz. The output power of the set of experiments varies from 383 W to 695 W.
The input and output waveforms of the prototype are recorded for loss estimation, while the total loss in the prototype is measured with thermal method. The internal temperature rise of the prototype is monitored for each experiment at thermal steady state as an indicator of the thermal conditions. The details of operational conditions in each experiment are listed in Table 7-6. The power density and corresponding thermal conditions of each experiment are listed in Table 7-7. The estimated of loss breakdown and the measured total loss for each experiment are listed in Table 7-8.

To investigate the influence of the application of heat extractors, a set of experiments are made with prototype II under the similar operational conditions with the experiments for prototype I. In practice, the operation conditions of two experiments are hardly to be the same, because the operation of the LLC resonant converter is sensitive to the circuit parameters and the frequency, while the device performance is sensitive the temperature. Nevertheless, if the corresponding frequency and magnitude of voltage and current of two experiments is close to each other, the generated loss and loss distribution in the prototype differ little, and thus the prototype can be regarded to be tested under the same operational conditions and the performance of prototypes can be compared. The details of operational conditions in each experiment with prototype II are listed in Table 7-9. The power density and corresponding thermal conditions of each experiment with prototype II are listed in Table 7-10. The details of loss estimation for prototype II in each experiment are listed in Table 7-11.

As shown in Tables 7-7 and 7-10, the monitored internal temperature rise of the prototypes is of small value. This implies that higher throughput power can be exerted to the prototypes to obtain higher power density without violating the thermal constraint. A set of experiments are specially made with prototype II to investigate the upper limit of the power density of prototype II. There are totally four experiments. In the first three experiments the output voltage of the converter is fixed at 48 V, while the load resistant changes from 5 ohms to 2.5 ohms and the output power increases from 454 W to 900 W. In the four experiment, the load resistant keeps at 2.5 ohms, while the output voltage increases to 51 W and the output power reaches 1008 W. The operational conditions of the set of experiments are listed in Table 7-12m and the power density, efficiency and
thermal conditions are listed in Table 7-13. The primary side waveforms in the four experiments are shown in Figure 7-19.

Table 7-6 Operation conditions for prototype I

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>f (kHz)</td>
<td>960</td>
<td>960</td>
<td>950</td>
<td>990</td>
<td>997</td>
</tr>
<tr>
<td>V_in (V)</td>
<td>302.0</td>
<td>326.4</td>
<td>353.7</td>
<td>285.0</td>
<td>302.7</td>
</tr>
<tr>
<td>I_in (A)</td>
<td>1.356</td>
<td>1.458</td>
<td>1.594</td>
<td>2.331</td>
<td>2.469</td>
</tr>
<tr>
<td>P_in (W)</td>
<td>409.5</td>
<td>475.9</td>
<td>563.8</td>
<td>664.4</td>
<td>747.4</td>
</tr>
<tr>
<td>V_out (V)</td>
<td>44.08</td>
<td>47.64</td>
<td>51.99</td>
<td>39.45</td>
<td>41.91</td>
</tr>
<tr>
<td>I_out (A)</td>
<td>8.70</td>
<td>9.37</td>
<td>10.20</td>
<td>15.63</td>
<td>16.57</td>
</tr>
<tr>
<td>P_out (W)</td>
<td>383.5</td>
<td>446.4</td>
<td>530.3</td>
<td>616.6</td>
<td>694.4</td>
</tr>
<tr>
<td>Loss_{converter} (W)</td>
<td>26.0</td>
<td>29.5</td>
<td>33.5</td>
<td>47.8</td>
<td>52.9</td>
</tr>
<tr>
<td>η_{converter} (%)</td>
<td>93.7%</td>
<td>93.8%</td>
<td>94.1%</td>
<td>92.8%</td>
<td>92.9%</td>
</tr>
</tbody>
</table>

Table 7-7 Power density and thermal conditions of prototype I

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Heatsink temperature (°C)</td>
<td>32.58</td>
<td>34.65</td>
<td>36.53</td>
<td>34.33</td>
<td>35.85</td>
</tr>
<tr>
<td>Temperature gradient (°C)</td>
<td>10.65</td>
<td>12.53</td>
<td>15.32</td>
<td>11.49</td>
<td>13.09</td>
</tr>
<tr>
<td>Measured loss (W)</td>
<td>10.187</td>
<td>12.672</td>
<td>15.256</td>
<td>11.734</td>
<td>13.450</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>97.3%</td>
<td>97.2%</td>
<td>97.1%</td>
<td>98.1%</td>
<td>98.1%</td>
</tr>
<tr>
<td>Power density (W/in³)</td>
<td>436.4</td>
<td>508.0</td>
<td>603.5</td>
<td>701.7</td>
<td>790.3</td>
</tr>
</tbody>
</table>
### Table 7-8 Calculated losses and measured total loss of prototype I

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core loss (W)</td>
<td>3.945</td>
<td>4.853</td>
<td>6.265</td>
<td>3.02</td>
<td>3.62</td>
</tr>
<tr>
<td>Primary winding loss (W)</td>
<td>0.579</td>
<td>0.669</td>
<td>0.787</td>
<td>1.231</td>
<td>1.371</td>
</tr>
<tr>
<td>Secondary winding loss (W)</td>
<td>0.612</td>
<td>0.703</td>
<td>0.843</td>
<td>1.823</td>
<td>2.043</td>
</tr>
<tr>
<td>Loss in interconnections (W)</td>
<td>0.200</td>
<td>0.230</td>
<td>0.273</td>
<td>0.534</td>
<td>0.597</td>
</tr>
<tr>
<td>Conduction loss (W)</td>
<td>1.389</td>
<td>1.602</td>
<td>1.904</td>
<td>3.588</td>
<td>4.011</td>
</tr>
<tr>
<td>Dielectric loss (W)</td>
<td>0.199</td>
<td>0.240</td>
<td>0.278</td>
<td>0.572</td>
<td>0.671</td>
</tr>
<tr>
<td>Total calculated loss (W)</td>
<td>5.533</td>
<td>6.695</td>
<td>8.447</td>
<td>7.18</td>
<td>8.302</td>
</tr>
<tr>
<td>Measured total loss (W)</td>
<td>10.187</td>
<td>12.672</td>
<td>15.256</td>
<td>11.734</td>
<td>13.450</td>
</tr>
</tbody>
</table>

### Table 7-9 Operational conditions for prototype II

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>f (kHz)</td>
<td>952</td>
<td>952</td>
<td>951</td>
<td>951</td>
<td>951</td>
</tr>
<tr>
<td>V_in (V)</td>
<td>302.9</td>
<td>328.5</td>
<td>351.4</td>
<td>277.4</td>
<td>294.0</td>
</tr>
<tr>
<td>I_in (A)</td>
<td>1.400</td>
<td>1.534</td>
<td>1.640</td>
<td>2.420</td>
<td>2.568</td>
</tr>
<tr>
<td>P_in (W)</td>
<td>424.1</td>
<td>503.9</td>
<td>576.3</td>
<td>671.3</td>
<td>755.0</td>
</tr>
<tr>
<td>V_out (V)</td>
<td>44.89</td>
<td>48.74</td>
<td>52.22</td>
<td>39.54</td>
<td>41.99</td>
</tr>
<tr>
<td>I_out (A)</td>
<td>8.84</td>
<td>9.70</td>
<td>10.37</td>
<td>15.63</td>
<td>16.58</td>
</tr>
<tr>
<td>P_out (W)</td>
<td>396.8</td>
<td>472.8</td>
<td>541.5</td>
<td>618.0</td>
<td>696.2</td>
</tr>
<tr>
<td>Loss_{con} (W)</td>
<td>27.2</td>
<td>31.1</td>
<td>34.8</td>
<td>53.3</td>
<td>5838</td>
</tr>
<tr>
<td>η_{converter} (%)</td>
<td>93.6%</td>
<td>93.8%</td>
<td>94.0%</td>
<td>92.1%</td>
<td>92.2%</td>
</tr>
</tbody>
</table>
### Chapter 7: Experimental Evaluation

Table 7-10 Power density and thermal conditions of prototype II

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Heatsink temperature (°C)</td>
<td>34.446</td>
<td>37.567</td>
<td>39.637</td>
<td>37.137</td>
<td>38.629</td>
</tr>
<tr>
<td>Temperature gradient (°C)</td>
<td>7.79</td>
<td>9.27</td>
<td>10.67</td>
<td>7.75</td>
<td>9.06</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>97.1%</td>
<td>97.0%</td>
<td>97.0%</td>
<td>97.9%</td>
<td>97.9%</td>
</tr>
<tr>
<td>Power density (W/in$^3$)</td>
<td>451.6</td>
<td>538.0</td>
<td>616.3</td>
<td>703.3</td>
<td>792.3</td>
</tr>
</tbody>
</table>

Table 7-11 Calculated losses and measured total loss of prototype II

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core loss (W)</td>
<td>4.865</td>
<td>6.126</td>
<td>7.481</td>
<td>3.640</td>
<td>4.339</td>
</tr>
<tr>
<td>Primary winding loss (W)</td>
<td>0.615</td>
<td>0.730</td>
<td>0.834</td>
<td>1.332</td>
<td>1.492</td>
</tr>
<tr>
<td>Secondary winding loss (W)</td>
<td>0.631</td>
<td>0.761</td>
<td>0.875</td>
<td>1.933</td>
<td>2.157</td>
</tr>
<tr>
<td>Loss in interconnections (W)</td>
<td>0.209</td>
<td>0.249</td>
<td>0.285</td>
<td>0.566</td>
<td>0.632</td>
</tr>
<tr>
<td>Conduction loss (W)</td>
<td>1.454</td>
<td>1.741</td>
<td>1.994</td>
<td>3.831</td>
<td>4.281</td>
</tr>
<tr>
<td>Dielectric loss (W)</td>
<td>0.199</td>
<td>0.240</td>
<td>0.278</td>
<td>0.572</td>
<td>0.671</td>
</tr>
</tbody>
</table>
## Chapter 7: Experimental Evaluation

### Table 7-12 Operational conditions for prototype II in high power density experiments

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f$ (kHz)</td>
<td>938</td>
<td>925</td>
<td>954</td>
<td>920</td>
</tr>
<tr>
<td>$V_{in}$ (V)</td>
<td>353.4</td>
<td>361.3</td>
<td>367.1</td>
<td>385.5</td>
</tr>
<tr>
<td>$I_{in}$ (A)</td>
<td>1.388</td>
<td>2.070</td>
<td>2.633</td>
<td>2.827</td>
</tr>
<tr>
<td>$P_{in}$ (W)</td>
<td>490.5</td>
<td>747.9</td>
<td>966.6</td>
<td>1089.8</td>
</tr>
<tr>
<td>$V_{out}$ (V)</td>
<td>47.82</td>
<td>48.44</td>
<td>48.11</td>
<td>51.00</td>
</tr>
<tr>
<td>$I_{out}$ (A)</td>
<td>9.51</td>
<td>14.38</td>
<td>18.71</td>
<td>19.77</td>
</tr>
<tr>
<td>$P_{out}$ (W)</td>
<td>454.8</td>
<td>696.6</td>
<td>900.1</td>
<td>1008.3</td>
</tr>
<tr>
<td>Loss$_{converter}$ (W)</td>
<td>35.75</td>
<td>51.32</td>
<td>66.43</td>
<td>81.54</td>
</tr>
<tr>
<td>$\eta_{converter}$ (%)</td>
<td>92.71</td>
<td>93.14</td>
<td>93.13</td>
<td>92.52</td>
</tr>
</tbody>
</table>

### Table 7-13 Power density and thermal conditions of prototype II in high power density experiment

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Heatsink temperature ($^\circ$C)</td>
<td>37.534</td>
<td>40.611</td>
<td>42.464</td>
<td>45.701</td>
</tr>
<tr>
<td>Temperature gradient ($^\circ$C)</td>
<td>9.901</td>
<td>11.476</td>
<td>12.425</td>
<td>14.388</td>
</tr>
<tr>
<td>Measured loss (W)</td>
<td>13.873</td>
<td>16.566</td>
<td>18.392</td>
<td>21.935</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>96.95</td>
<td>97.62</td>
<td>97.96</td>
<td>97.82</td>
</tr>
<tr>
<td>Power density (W/in$^3$)</td>
<td>517.5</td>
<td>792.7</td>
<td>1024.3</td>
<td>1147.4</td>
</tr>
</tbody>
</table>
Chapter 7: Experimental Evaluation

7.5. Evaluation of the experimental works

The experimental works are evaluated on four aspects: the electromagnetic performance, the accuracy of the loss models, the influence of the application of heat extractors and the power density.

7.5.1. Evaluation with respect to electromagnetic parameters

The experimental results show that the integrated structure can provide the required electrical performance under the operational conditions. The small signal test and the in-circuit test results provide evidences that the electromagnetic design model can be implemented with sufficient accuracy to realize the desired equivalent circuit parameters. Small variation in the leakage inductance is mainly due to the assembly tolerance and can be tuned internally and externally if required. This is accomplished by slightly tuning the winding positions (the insulation distance between windings) and the length of the external leads.

When the stacked structure is applied to implement multi-turn winding, the structure is inherently asymmetric. The inherent leakage inductance in the center-tapped secondary windings is uneven. In general designs, the overall resonant inductance composites the
inherent leakage inductance and the additional leakage inductance created by the leakage layer, and the additional leakage inductance dominant, thus the unbalance is not significant. In some cases, no or small additional leakage inductance is applied, the inherent leakage inductance is close to but slightly less than the required resonant inductance. By adjusting the length and position of the external leads in the secondary side, the overall resonant inductance of both secondary windings can be adjusted to the desired value. Thus, the structural asymmetry does not exert a crucial constrain on the application of the stacked structure in implementing the integrated LLCT module.

7.5.2. Evaluation of the loss models

The estimated total loss in each experiment of prototype I and II are compared with the measured values and shown in Figures 7-20 and 7-21. The relative error of the total loss with respect to the measured value is shown in percentage in Figure 7-22. The calculated total loss is less than the measured value in all of the ten experiments, but the calculated values are in the same trend with the measured values. The relative error in the ten experiments is within a certain range. For prototype I, the relative error is within the range from 38% to 47% and has the average value of 42%, while for prototype II, the relative error is within the range from 36% to 45% and has the average value of 40%. For each experiment, the relative error of prototype II is slightly larger than that of prototype I.

To analyze the loss breakdown, the percentage of the three calculated losses to the total calculated loss is shown in Figures 7-23 and 7-24 respectively for prototype I and II. With different load resistance, the ratio between voltage and current differs, and thus the calculated loss distribution in different materials differs too. According to the calculation results, the loss distribution for the two prototypes differs little. With load resistance of 5 ohms, the calculated core loss contributes more than 75% of the calculated total loss. With load resistance of 2.5 ohms, the percentage of core loss reduces to 45%.
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Figure 7-20 Comparison of the calculated and measured values of the total loss of prototype I

Figure 7-21 Comparison of the calculated and measured values of the total loss of prototype II
Figure 7-22 Relative error of the total calculated loss with respect to the measured value in the two sets of experiments with prototype I and II respectively.

Figure 7-23 Percentage of different losses to the total loss, calculated values, prototype I.
Since the total loss of the prototype is measured by thermal method with high accuracy, the measured values are more reliable more the calculation results. There are several reasons leads to the discrepancy between the measured values and the calculated values.

As stated in Chapter 5, the core loss is calculated by the improved generalized Steinmetz equation (iGSE) with the in-circuit voltage waveforms across the secondary windings. The iGSE is an improved empirical equation and assumes uniform magnetic flux distribution in the core, which is not feasible for a planar core, especially when the core thickness is small. The influence of the leakage flux to the core loss is ignored in core loss calculation. This leads to underestimation of the core loss.

The calculation of conduction loss is based on the assumption of one dimensional electromagnetic field. Under this assumption, the fringing effect and the end effect are ignored. It assumes that the current density distribution is uniform along the conductor width direction. This assumption underestimates the conduction loss. The current concentration in the leads should be more significant but not taken into account. This also leads to underestimation. The copper thickness obtained by electroplating could be much different from the design value, while the conduction loss is sensitive to the copper thickness.
With the conduction loss model, the percentage of the conduction losses contributed by windings and interconnections to the total conduction loss is shown in Figures 7-25. The calculation results show that the percentage of the conduction loss in the secondary winding to the total conduction loss increases when the output current increase. In all the five experiments, the conduction loss in the interconnections occupies a constant percentage, about 14%, in the total conduction loss.

![Figure 7-25 Calculated conduction loss distribution in prototype I](image)

**7.5.3. Evaluation of the improvement by heat extractors**

To evaluate the influence of the embedded heat extractors to the prototype, the testing conditions of the prototypes need to be identified. Though the operational conditions for prototype II are controlled to be similar to those for prototype I, it is hard to make the operational conditions in two independent experiments exactly the same. Nevertheless, if the difference of operational conditions in the two experiments makes little influence on the prototype performance and can be ignored, we regard the two experiments have the same operational conditions and will induce the same results on the same prototype. Then if we replace the prototype in one of the experiments by another prototype, the performance of the two prototypes can be compared as they are tested under the same operational conditions. To clearly show the difference of the operational conditions between the two sets of experiments, the parameter values are listed in Table 7-14.
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The Steinmetz equation (5-2) is applied to evaluate the sensitivity of the core loss to the variance of frequency and voltage.

\[ P_v = k f^\alpha B^\beta = Gf^{\alpha - \beta} \]  

(7-1)

The variance of the frequency \( \Delta f \) lead to the percentage change of the core loss of

\[ \left( \frac{\Delta P_v}{P_v} \right)_{\Delta f} = \left( \frac{P_v(f) \cdot \Delta f}{P_v(f)} \right) = \frac{\left( \alpha - \beta \right) \cdot \Delta f}{f} \]  

(7-2)

For core material 3F4, \( \alpha \) is 1.75 and \( \beta \) is 2.9; the maximum frequency variance shown in Table 7-14 is 46 kHz; the average frequency is about 970 kHz. Thus, the percentage of core loss change with respect to the frequency variance is about 5.5%.

Similarly, the percentage change of the core loss respect to the voltage variance is calculated with

\[ \left( \frac{\Delta P_v}{P_v} \right)_{\Delta V} = \frac{\beta \cdot \Delta V}{V} \]  

(7-3)

From Table 7-14, the maximum variance of output voltage is 0.81 V at the output voltage of 48V, and then the percentage of core loss change with respect to the voltage variance is about 4.9%.

The percentage change of the conduction loss respect to the current variance is calculated with

\[ \left( \frac{\Delta P_{\text{Cu}}}{P_{\text{Cu}}} \right)_{\Delta I} = \frac{2 \cdot \Delta I}{I} \]  

(7-4)

From Table 7-14, the maximum variance of output current is 0.33 A at the output current of 9.5A, and then the percentage of conduction loss change with respect to the output current variance is about 6.9%.

To obtain more accurate comparison of the two sets of operational conditions, both sets of operational conditions are used to calculate the total loss and loss distribution with respect to prototype I with the loss models developed in Chapter 5, respectively. If the difference between the two sets of calculation results is small, it can be assumed that the two sets of operational conditions are the same and the difference of prototype
performance under the two sets of operational conditions is mainly due to the different structural characteristics of the prototypes. Calculation results for the two sets of operational conditions are compared one by one as shown in Figure 7-26. The percentage difference of the calculation results are shown in Figure 7-27. The largest difference exists in the calculated conduction loss in experiment 2, where the percentage difference is 8.68 %, but the amount of the conduction loss is small in this experiment. The percentage difference of the calculated total loss is within a small range, less than 4.5%.

Table 7-14 Comparison of the two sets of operational conditions

<table>
<thead>
<tr>
<th>Experiment</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>f (kHz)</td>
<td>Prototype I</td>
<td>960</td>
<td>960</td>
<td>950</td>
<td>990</td>
</tr>
<tr>
<td></td>
<td>Prototype II</td>
<td>952</td>
<td>952</td>
<td>951</td>
<td>951</td>
</tr>
<tr>
<td>V_in (A)</td>
<td>Prototype I</td>
<td>302.0</td>
<td>326.4</td>
<td>353.7</td>
<td>285.0</td>
</tr>
<tr>
<td></td>
<td>Prototype II</td>
<td>302.9</td>
<td>328.5</td>
<td>351.4</td>
<td>277.4</td>
</tr>
<tr>
<td>I_in (A)</td>
<td>Prototype I</td>
<td>1.356</td>
<td>1.458</td>
<td>1.594</td>
<td>2.331</td>
</tr>
<tr>
<td></td>
<td>Prototype II</td>
<td>1.400</td>
<td>1.534</td>
<td>1.640</td>
<td>2.420</td>
</tr>
<tr>
<td>V_out (V)</td>
<td>Prototype I</td>
<td>44.08</td>
<td>47.64</td>
<td>51.99</td>
<td>39.45</td>
</tr>
<tr>
<td></td>
<td>Prototype II</td>
<td>44.89</td>
<td>48.74</td>
<td>52.22</td>
<td>39.54</td>
</tr>
<tr>
<td>I_out (A)</td>
<td>Prototype I</td>
<td>8.70</td>
<td>9.37</td>
<td>10.20</td>
<td>15.63</td>
</tr>
<tr>
<td></td>
<td>Prototype II</td>
<td>8.84</td>
<td>9.70</td>
<td>10.37</td>
<td>15.63</td>
</tr>
</tbody>
</table>

Figure 7-26 Calculation results for prototype I with the operation conditions exerted on prototype II, compared with the original calculation results of prototype I
The small difference between the calculated total loss with the two sets of operational conditions for prototype I indicates that the two sets of operational conditions can be regarded as the same with a small tolerance. Keeping in mind that the difference of the two sets of operational conditions varies within a small range of 4.5%, the experimental results of the two sets of experiments can be compared. The measured total loss, efficiency and temperature performance of prototype I and prototype II are shown in Figures 7-28 ~ 7-30 respectively for comparison. The percentage error refers to the difference between values obtained from the two prototypes with respect to the values obtained from prototype I in percentage.

The influence of the applied heat extractors can be summarized in the following. The total loss of prototype II is larger than that of prototype I, and the total loss increases in percentage is within the range from 5.6% to 14.8% in the five comparison experiments. The efficiency of prototype II is slightly less than that of prototype I. The efficiency reduction is less than 0.3% in the five comparison experiments. The internal temperature rise in prototype II is much lower than that in prototype I. The percentage of temperature rise reduction with respect to the temperature rise in prototype I is within the range from 32.8% to 40% in the five experiments. That means the application of heat extractors
(10% of the original core volume) reduces the internal temperature rise to about 67.8% of the original value. By maintaining the same temperature rise with the prototype I, the throughput power of prototype II can be 132% of that of prototype I, if all other conditions are unchanged. The presented value 132% is just a rough estimation, because the electrical and thermal conditions will change in practical cases. In addition, when the thermal conditions change, the material properties will change and make the results more unpredictable.

A method to evaluate the different performance of two prototypes is presented in [82, 83]. In this method, the heat sink temperature is fixed as the reference temperature and the maximum inside temperature is maintained at a certain value in the comparison experiments by controlling the input power. At thermal steady states, the input power and power density of the two experiments is comparison and represented as the performance comparison of the two prototypes. A similar method is to fix the maximum inside temperature as the reference and maintain the heat sink temperature by controlling the input power. To use this method, the external cooling conditions should be unchanged. However, these two methods are time-consuming and not adopted in this research.

![Figure 7-28](image-url)  
Figure 7-28 The total losses measured on prototypes I and II in the two sets experiments respectively as well as the loss increase in prototype II with respect to that in prototype I in percentage
Figure 7-29 The efficiency measured on prototypes I and II in the two sets experiments respectively and the efficiency reduction in prototype II compared to that in prototype I.

Figure 7-30 The internal temperature rise measured on prototypes I and II in the two sets experiments respectively and the reduction of temperature rise in percentage with respect to the temperature rise in prototype I.
7.5.4. Evaluation on the aspect of power density

In this research, the power density of a power electronics module refers to the ratio of the output power to the overall volume of the module. For a practical module, different output power relates to different power density, also leads to different efficiency and internal temperature gradient. Regarding power density as one of the characteristics of the component, the identification of power density requires information including the output power, efficiency and the internal temperature rise, as well as the operation conditions, which determines the loss distribution and thus influences the internal temperature rise. For the integrated LLCT prototypes, the output power listed in Table 7-7, 7-10 and 7-13 is the dc output power of the testing converter. The real output power of the prototype should be the sum of the loss in the rectifier and the converter output power. Since the rectifier applied is the silicon carbide schottky diode D20S30 from Infineon technologies with the forward voltage 1.5 V, the accurate values of the prototype power density would be dozens of wattage per cubic inches more than the values listed in the tables.

The experimental results shown in section 7.4 indicates that prototype I, which is implemented with stacked structure, reaches the power density of 790 W/in³ with the efficiency of 98.1% and internal temperature gradient of 13.09 °C. The testing converter functions at the frequency of 997 kHz with the output dc voltage 41.91 V and current 16.57 A. The small internal temperature rise indicates higher power density can be obtained in this prototype without violating the thermal constraint.

The experimental results shown section 7.4 indicates that prototype II, which has embedded heat extractors occupying 10% of the core volume, reaches the power density of 1147 W/in³ with the efficiency of 97.8% and internal temperature rise of 14.39 °C. The testing converter functions at the frequency of 920 kHz with the output dc voltage 51.0 V and current 19.8 A. The small internal temperature rise indicates higher power density can be obtained in this prototype without violating the thermal constraint.

Though the thermal conditions allow pushing the power density to higher level, the prototype may suffer lower efficiency with higher output power. Figures 7-31 and 7-32 show the total loss and internal temperature gradient increases quickly with the increase
of output power. As shown in Figure 7-33, in these experiments, the efficiency reaches the maximum, when the output power is 900 W with the output dc voltage 48.11 V and current 18.7 A.

![Graph showing LLCT Loss vs. Output Power](image1)

**Figure 7-31** Prototype II performance: loss vs. output power

![Graph showing Internal Temperature Gradient vs. Output Power](image2)

**Figure 7-32** Prototype II performance: internal temperature gradient vs. output power
7.5.5. Summary

From the above design and experimental results, the following analysis and conclusions are drawn.

- Four practical constraints limit the valid design range for the stacked structure when it is applied to implement the integrated LLCT module. As one of the general limitations of the planar integrated structures, the limited turns number lead to small inductance of the structure. Since the LLC resonant converter requires a small magnetizing inductance, the limited turns number does not impact the valid design range. Another drawback of the integrated structure is that the integrated winding is normally thicker than conventional discrete winding. The thicker integrated windings occupy more space in the winding window, which leads to larger inherent leakage inductance. The inherent leakage inductance is the major design constraint of the high frequency integrated LLCT module, which applies the leakage inductance as the resonant inductance and requires a small value. The limitation of the processing technology also constraints the design range through the minimum acceptable core thickness.

- The two-turn center-tapped secondary winding structure implemented with the stacked structure is an asymmetric structure. The different positions of the two winding turns lead to different inherent leakage inductance with respect to the
primary winding. In general designs, the unbalanced inherent leakage inductance is of small difference and can be compensated by the external lead inductance.

- Two prototypes are built based on the selected design. The small signal test and in-circuit test results show the integrated prototypes provide desired parameters and electrical performance and verifies the validity of the implementation technologies.

- The total in-circuit loss of the prototypes is measured with thermal method with the estimated measurement error of 4.8% (refer to Appendix E). The obtained measurement results provide reliable values of the in-circuit total loss of the prototype.

- Large discrepancy exists between the estimated total loss and the measured value in all experiments. Within the ten experiments, the relative error with respect to the measured values is within the range from 36% to 47%. Though the loss models underestimate losses, the calculated values of the total loss are in the same trend with the measurement results.

- The underestimation of the core loss may be due to failing to satisfy the assumption required by the empirical model, iGSE. The magnetic flux in the thin planar core is not uniform as required by the model. And the empirical model itself generates certain estimation error. The neglect of the leakage flux contribution to the core loss is another reason of underestimating the core loss.

- The conduction loss model is based on the assumption of one-dimensional electromagnetic field distribution along the conductor width direction and the linear current distribution along the length direction. It is not the truth in real prototypes. The fringing effect and edge effect neglected in the model actually induce eddy current and condition loss, which is not taken into account. The copper thickness of the primary winding is controlled by the electroplating technology and may vary greatly from the design value. Whereas, the research stated in Chapter 5 indicates the conduction loss is sensitive to the copper thickness.

- The conduction loss in the interconnections is estimated with the skin effect loss only. The calculation results show that the loss in the interconnection contributes little, about 14%, to the total conduction loss.
The dielectric loss is estimated by the empirical equation with voltage across the resonant capacitor obtained from simulation results. The calculation results show that the dielectric loss contributes (3.5\% - 8.1\%) of the total loss. Nevertheless, the applied ceramic material N1250 is experimentally proven to have very low loss factor by large signal test results [56]. The empirical loss model assumes the loss factor as 1\% and may overestimate the dielectric loss.

The application of heat extractors (occupying 10\% of the original core volume) increases total loss and slightly reduces the efficiency with a very small amount, less than 0.3\% for all the ten experiments. On the other hand, the embedded heat extractors greatly reduce the internal temperature rise to 68\% or less of the original value. The improved thermal performance allows higher power density without violating the thermal constraint.

Associated with the efficiency and internal temperature gradient, the power density is regarded as one of the characteristics representing the performance of practical modules. The prototype implemented with the stacked structure without heat extractors reaches the power density of 790 W/in3 with the efficiency of 98.1\% and internal temperature rise of 13.09 °C. The prototype with heat extractors reaches the power density of 1147 W/in3 with the efficiency of 97.8\% and internal temperature rise of 14.39 °C. Higher power density can be obtained in both prototypes, but the determination of the practical value of power density is always a tradeoff between the available space and the required efficiency, and limited by external cooling conditions.
Chapter 8: Conclusions and Future Work

8.1. Conclusions

The study as presented consists of the electromagnetic modeling, structural study, electromagnetic design, loss modeling, thermal management, design evaluation and experimental evaluation of the whole process. The conclusions drawn for each of these sections are summarized below.

8.1.1. Electromagnetic modeling

The previous work of electromagnetic modeling of integrated passive is summarized in Chapter 2. The simplified design model is presented as the background knowledge. The asymmetric performance of integrated structures is observed and investigated. The electromagnetic modeling of inhomogeneous integrated structure is proposed in Chapter 2 and Appendix A by form of differential equations. The knowledge developed in this study is applied to understand the asymmetric feature of the multi-turn stacked structure proposed in Chapter 3. The awareness of asymmetric performance of integrated passive is importance to the future research and applications of integrated passives, especially when the frequency increases and the impact of parasitic parameters becomes more and more significant.

8.1.2. Structural study

In Chapter 3, the drawback of the planar spiral winding structure is studied and two alternative structures, the symmetric single layer structure and the stacked structure are proposed for improvement. Structural study shows that the stacked structure is of the highest potential to minimize both the overall volume and the losses. The interconnection structure for the stacked structure is proposed with the implementation details and the corresponding processing technologies as well as the experimental verification of the validity.
8.1.3. Electromagnetic design

In Chapter 4, a practical design approach is proposed to investigate the limitation of the stacked structure and to develop the optimal design. Loss models are included in the design approach. Design maps with respect to the characteristics of the integrated module are developed to investigate the relationship between the interested characteristics and the design variables, and to easily select the optimal design by making tradeoffs. The proposed design approach can be extended to the other integrated structures.

8.1.4. Electromagnetic loss modeling

The electromagnetic modeling, performed in Chapter 5, is based on the electromagnetic modeling developed in Chapter 4 and modified to include the energy distribution within the structure. The resultant model estimates conduction losses (both skin and proximity effects), dielectric losses and core losses – all for non-sinusoidal electromagnetic fields with closed-form expressions. The developed conduction loss model is extended and applied to identify the optimal conductor thickness for the design process.

8.1.5. Thermal management

In Chapter 6, the heat extraction technology is developed with the theoretical calculation and the experimental verification on the sole-heat-generation-medium component. With the obtained success, heat extraction technology is applied to integrated passive modules. Analytical study and numerical simulation with the stacked structure shows that the power density limited by thermal constraints can be extended by improving the thermal performance with heat extractors.

8.1.6. Experimental evaluation

Utilizing the presented design evaluation tool, two prototypes were constructed with and without embedded heat extractors respectively for experimental evaluation. All are constructed according to the general processing technologies described in Appendix D.

The experiments are made with a 1 MHz LLC resonant converter. A measurement structure is developed to measure the in-circuit loss of prototype by monitoring the heat flux at thermal steady states, as described in Appendix E.
Two sets of experiments with similar operational conditions are made to evaluate the electromagnetic performance, the efficiency and the thermal performance. Evaluation of the electromagnetic characteristics of the prototypes compares favorably with the design criteria. Variations are within the acceptable range. In-circuit loss measurement shows high efficiency of the prototypes, more than 97%, for all the experiments. Compared with the measured values, the loss models underestimate the losses. The relative error reaches 46% of the measured values. Nevertheless, the calculated values are in the same trend of the measured values. The variance of the relative error is within 10%.

The prototype without heat extractors reaches the power density of 603 W/in\(^3\) (36.8 W/cm\(^3\)) with the efficiency of 97.1% and the internal temperature gradient of 15.32 °C at the rated converter output power (530 W). The maximum power density obtained in the experiments for the prototype is 790 W/in\(^3\) (48 W/cm\(^3\)) with the efficiency of 98.1% and internal temperature gradient of 13.09 °C at the output power of 694 W. The data indicate that the operational condition influences the loss distribution and thermal performance of the prototype.

The application of heat extractors (occupying 10% of the original core volume) increases the total loss to 110% (±5%) of the original values, and leads to a slight reduction on the efficiency, 0.3%, but greatly improves the thermal performance by reducing the internal temperature rise to 64% (±4%) of the original values in the experiments. The prototype has the power density of 616 W/in\(^3\) (37 W/cm\(^3\)) with the efficiency of 97.0% and the internal temperature gradient of 10.67 °C at the rated output power (541 W). Taking the advantage of the good thermal performance, the highest power density obtained in the additional experiments is 1147 W/in\(^3\) (70 W/cm\(^3\)) with the efficiency of 97.8% and the internal temperature gradient of 14.79 °C at the converter output power of 1008 W. The loss internal temperature gradient implies that higher power density can be obtained without violating the thermal constraint.

8.2. Future research

- The loss modeling needs be improved to better estimate the total loss and loss distribution. By setting up comparison experiments with fixed voltage or current...
amplitude respectively, the core loss model and conduction loss model can be calibrated with the measured total loss for the specific prototype.

- With the modified design approach and improved loss models, the thermal management needs to get involved to provide the optimal design with heat extractors by making tradeoff between the efficiency and the internal temperature gradient.

- Finally, with the increase of power density and frequency, the parasitic parameters will get involved in the operation performance. Better modeling of the higher order resonances within the structure and possible design modification to suppress should be investigated. Conversely, the energy storage elements that cause these higher order resonances may be applied in a constructive way.
Appendix A: Study of Asymmetric Integrated Structures

A.1. Introduction

In recent years, electromagnetic integration has been developed and applied in the field of power electronics. More and more structures have been proposed to implement integrated passive modules. These structures include barrel wound structures [41], planar transmission line structures [25], planar spiral winding structures [84], the stacked structure [85], and so on. All of these proposed structures are regarded as symmetric structures by default, due to the homogeneous construction. In most cases, symmetric performance is observed in modules made with these structures, because the processing technologies provide high symmetry for these simple structures. However, with the proposal of a magnet wire winding structure [47], asymmetric performance is observed. Further research shows that almost all practical integrated passive modules, no matter how simple they are, to some extent, are asymmetric, due to the dimensional tolerances induced in the processing procedure. For integrated passives with complicated structures, the asymmetry is often inherent. With the developing applications of integrated passives, the understanding of its asymmetric feature becomes more and more important.

A.2. Observation of asymmetric performance

The simplest integrated structure consists of two conductors isolated with a dielectric material in between as shown in Figure A-1.

![Figure A-1: The simplest integrated structure, LC cell](image)

The asymmetry was firstly observed in an integrated passive with magnet wire winding structure. This is an integrated LC series resonator. Two magnet wire windings are wound around a core and insulated by Kapton tape. The terminals a, b, c, d shown in Figure A-2 has the same relationship with each other as shown in Figure A-1. As shown in Figure A-2 (b), the two windings are interleaved layer by layer. Both windings have
the same numbers of turns and layers. The frequency response curve measured from terminal pair a & d and that from b & c are shown in Figure A-3.

As we observed, the first and second resonant frequencies in the two measurements are different. Consequently, the equivalent lumped parameter values are different. That means two different performances can be obtained in one unit by simply changing the excitation terminal pairs. On the other hand, the specific performance is associated with the specific terminal pairs of the unit. Different terminal pairs can not be exchanged. This unit is not symmetric.

The asymmetry of this unit can be understood intuitively. Because it is a multi-layer structure and the winding layers of different windings are interleaved layer by layer, one winding occupies an inner position and the other an outer position. The inner and outer positions correspond to different distance between the winding layers and the core as well as different space for passing the magnetic flux. It is obvious that the inner layer has a different inductance from the outer layer, and the inter-winding capacitance between two adjacent layer pairs is different for different layers. Actually, a multi-layer structure is not a symmetric; the inner terminals can not be exchanged with the outer terminals.

The question is what the extent of the asymmetry is. Read from Figure A-3, the first resonant frequency obtained from the two terminals pairs is quite different. We may suspect that the large discrepancy is due to the uncertainty of position of the magnet wire windings made by hand. If the integrated passive is made with other technologies, where the structure dimensions can be controlled accurately, will the discrepancy be eliminated?

A transformer made by printed circuit board technology was tested as a resonator to obtain the terminal performance on different terminal pairs, since PCB technology can produce windings with very tight tolerances. As shown in Figure A-4 (a), the transformer has a planar core and windings made in a piece of PCB. The winding structure is shown in Figure A-4 (b). A transformer is a structure with two isolated conductors and can be regarded as a resonator, though the performance may not be as good as the specially designed resonator. Similar to the magnet wire winding resonator, the terminals of the PCB resonator are designated in Figure. As observed from the frequency response curves in Figure A-5, a large discrepancy exists. The asymmetry inherently lies in the structure, not the processing technology.
Appendix A: Study of Asymmetric Integrated Structures

(a) An integrated LC series resonator with magnet-wire winding

(b) Cross-sectional view of the winding structure

Figure A-2 An integrated LC resonator with magnet wire windings

Equivalent parameters:

L = 348 μH
C = 308 pF
R = 13 Ω

(a) Frequency response curve measured on terminals A and D

(b) Frequency response curve measured on terminals B and C

Figure A-3 Frequency response curves of the integrated resonator with magnet-wire winding structure

(a) Integrated series resonator with PCB windings

(b) Cross-sectional view of half of the winding window

Figure A-4 An integrated LC resonator with PCB windings

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A.3. **Definition of asymmetric structure**

To further investigate the asymmetric feature of the integrated structures, definitions and criteria for the asymmetric structure are proposed in this section. The considerations are based on the simplest integrated structure, two-isolated-conductor structure with four terminals.

### A.3.1. Definition

For general applications in the field of power electronics, the passive components, except EMI filters, function at low frequencies (less than several mega hertz). For these applications, only the performance around the first two resonances of the integrated structure is of concern. For the low frequency applications, the integrated structures with excitation terminals on different conductors at different ends provide higher efficiency of the space usage. Thus, in the following discussion, only the performance on terminal pairs A & D, and that on B & C is considered. The definition of asymmetric structure is based on the interchangeability of the two terminal pairs. For a two-conductor integrated structure as shown in Figure 1, if the two terminal pairs A & D and B & C can be interchanged without changing the terminal performance, the integration structure is regarded a symmetric structure.

A two-conductor structure can be regarded as a system with distributed parameters, such as distributed inductance L, capacitance C, resistance R, and admittance G.
Appendix A: Study of Asymmetric Integrated Structures

Obviously, all homogeneous structures are symmetric structures due to the uniform distribution of electrical parameters, which is guaranteed by the uniform dimensions and material distribution, as shown in Figure A-6 (a). However, an inhomogeneous structure is not necessarily asymmetric.

According to the positions of the four terminals, there exist two possible central lines. The vertical central line corresponds to the position relationship between terminals A and C, and terminals B and D. The horizontal central line corresponds to the position relationship between terminals A and B, and terminals C and D. For an arbitrary two-conductor integrated structure, if the distribution of electrical parameters is symmetric about any of the two central lines, the structure is symmetric.

![Homogeneous structure with two symmetric axes](image1)

![Symmetric inhomogeneous structure](image2)

![Structure symmetric about the vertical central line](image3)

![Structure symmetric to the horizontal central line](image4)

Figure A-6 Symmetric structures

Figure A-6 (b), (c) and (d) show the three situations of the symmetric inhomogeneous integrated structures. In the diagrams, the shape of conductors represents the dimension distribution and shading represents the distribution of the material properties. The excitation current path is shown with arrows. For integrated structures having at least one central line, the current path obtained from different excitation terminals is the same, thus the electrical performance on different excitation terminals is the same.

It is easy to find the central line in some integrated structures having single turn or single layer of winding. In complicated structures, such as multi-layer structures, normally there is no such a central line.
Appendix A: Study of Asymmetric Integrated Structures

A.3.2. Symmetric performance in inhomogeneous structures

The criteria of a symmetric inhomogeneous integrated structure were verified by two samples made with magnet wire windings. The first sample is to verify the symmetry due to the existence of the vertical central line between the two conductors. As shown in Figure A-7 (a), the resonator consists of two one-layer windings. The winding structure is shown in Figure A-7 (b). The inner layer is made of magnet wire AWG #30, and has 57 turns. The outer layer is made of magnet wire AWG #26, and has 29 turns. The adjacent turns contact each other closely. The winding in each layer fully occupies the winding window. The structure is symmetric about the vertical central line, but not the horizontal central line. Read from the measured frequency response curves, we can see the performance on different terminal pairs is very similar. Considering the position tolerance induced by the magnet wire, the structure can be regarded as symmetric.

The second sample is to verify the symmetry due to the existence of the horizontal central line. As shown in Figure A-8 (a), two magnet wires, AWG #30, are twisted and then wounded around the bobbin to form multi-layer windings. Since the multi-layer winding was wound randomly, there is no vertical central line in this structure, but the horizontal central line exists at every point between the two wires along the winding length, because the two wires are twisted. The measured performance, shown in Figure A-8 (b) and (c), verify the symmetry.

Comparing the performance shown in Figure A-8 and A-9 with that shown in Figure A-5, we can see that the asymmetry lies inherently in the structure, instead of in the processing technologies.

(a) 3D view of the integrated structure symmetric about the horizontal central line
Appendix A: Study of Asymmetric Integrated Structures

(b) Performance on terminals A and D

(c) Performance on terminals B and C

Figure A-7 An integrated structure – symmetric about the horizontal central line

(a) 3D view of the integrated structure

(b) Cross-sectional view of the winding structure

(c) Performance on terminals A and D

(d) Performance on terminals B and C

Figure A-8 An integrated structure – symmetric about the vertical central line

A.4. Simplified model

Based on the above discussion, a simplified model of asymmetric integrated structures is proposed, as shown in Figure A-9. The model consists of both the horizontal and vertical central lines. Since the capacitance between the two conductors in a two-conductor structure is always symmetric about the horizontal central line, the vertical distribution of the capacitance is not considered in the model.
Appendix A: Study of Asymmetric Integrated Structures

The simplified model has sufficient parameter to represent the asymmetric characteristic of integrated structures at low frequencies. For any asymmetric integrated module, the equivalent lumped parameters of the model can be extracted according to the distribution of parameters, the current, and the voltage. The parameter values are calculated with the energy method – the energy stored in the equivalent lumped parameters should be equal to the integral of the energy distributed in the corresponding part of the structure. In a theoretical study, by varying the parameter values in the model, the characteristics of the asymmetric structure can be investigated easily through circuit simulation.

The circuit diagram shown in Figure A-10 is used to simulate the performance of asymmetric structures with known lumped parameter values. The magnetic coupling between the two conductors is taken into account and represented by the coupling coefficients M1 and M2. The simulation is based on a set of estimated values listed in Table A-1. The results are shown in Figure A-11 and Table A-2. It shows that the first and third resonant frequency obtained on the two excitation terminal pairs a & d and b & c are quite different, while the second resonant frequency varies little.

The simplified model carries the constructional features of asymmetric structures, and can be used to investigate the factors influencing the extent of asymmetry. However, in practical cases, it is difficult to identify the central line on the physical component.
Appendix A: Study of Asymmetric Integrated Structures

Figure A-10 Circuit diagram of the simplified model

Figure A-11 Simulated frequency response curves

Table A-1 Parameters for simulation

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>L&lt;sub&gt;1&lt;/sub&gt;, L&lt;sub&gt;4&lt;/sub&gt;</td>
<td>2 mH</td>
</tr>
<tr>
<td>L&lt;sub&gt;2&lt;/sub&gt;, L&lt;sub&gt;3&lt;/sub&gt;</td>
<td>0.5 mH</td>
</tr>
<tr>
<td>R&lt;sub&gt;1&lt;/sub&gt;, R&lt;sub&gt;4&lt;/sub&gt;</td>
<td>2 ohms</td>
</tr>
<tr>
<td>R&lt;sub&gt;2&lt;/sub&gt;, R&lt;sub&gt;3&lt;/sub&gt;</td>
<td>0.5 ohms</td>
</tr>
<tr>
<td>C&lt;sub&gt;1&lt;/sub&gt;, C&lt;sub&gt;2&lt;/sub&gt;, C&lt;sub&gt;3&lt;/sub&gt;</td>
<td>50 pF</td>
</tr>
<tr>
<td>G&lt;sub&gt;1&lt;/sub&gt;, G&lt;sub&gt;2&lt;/sub&gt;, G&lt;sub&gt;3&lt;/sub&gt;</td>
<td>10&lt;sup&gt;-8&lt;/sup&gt; ohm&lt;sup&gt;-1&lt;/sup&gt;</td>
</tr>
<tr>
<td>M&lt;sub&gt;1&lt;/sub&gt;, M&lt;sub&gt;2&lt;/sub&gt;</td>
<td>0.9</td>
</tr>
</tbody>
</table>
Table A-2 Simulation results

<table>
<thead>
<tr>
<th>Terminals</th>
<th>Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>AD</td>
<td>BC</td>
</tr>
<tr>
<td>1\textsuperscript{st} resonance</td>
<td>247.7 kHz</td>
</tr>
<tr>
<td>2\textsuperscript{nd} resonance</td>
<td>1.47 MHz</td>
</tr>
<tr>
<td>3\textsuperscript{rd} resonance</td>
<td>2.64 MHz</td>
</tr>
</tbody>
</table>

A.5. Generalized model

A generalized model is proposed to investigate the asymmetric structure performance with known distribution of the parameters. In the generalized model of a homogeneous two-conductor structure, the parameter distribution is regarded as uniform, and all parameters have constant values along the structure [28]. In the generalized model of an inhomogeneous structure, the parameter values are functions of the positions.

Setting the wave propagation direction as the $x$ axis, the parameters can be represented as $L(x)$, $C(x)$, $R(x)$ and $G(x)$. The lumped-parameter model of an infinitesimal element in the integrated structure is shown in Figure A-12. The top conductor is labeled as conductor 1 and the bottom one is labeled as conductor 2. The model assumes current flows into terminal a and b and out of terminals c and d. The voltage across terminals a and c is denoted by $V_1$, the voltage across terminals b and d is denoted by $V_2$, and the voltage across terminals a and b is denoted by V. Values of parameters, current and voltage are all functions of $x$.

Applying Kirchhoff’s voltage and current laws yields,

\begin{align*}
\frac{d}{dx} I_1(x) &= -(G(x) + j\omega C(x))V(x) \\
\frac{d}{dx} I_2(x) &= (G(x) + j\omega C(x))V(x) \\
\frac{d}{dx} V_1(x) &= [R_1(x) + j\omega L_1(x)]I_1(x) + j\omega M(x)I_2(x) \\
\frac{d}{dx} V_2(x) &= [R_2(x) + j\omega L_2(x)]I_2(x) + j\omega M(x)I_1(x)
\end{align*}

By defining:
From (A-1), we have,
\[ \frac{d}{dx} I_{1-2}(x) = y(x)V(x) \] (A-6)

From (A-2) and Kirchhoff's voltage law, we have
\[ \frac{d}{dx} V(x) = R_1(x)I_1(x) - R_2(x)I_2(x) + j\omega L_1(x)I_1(x) \\ - j\omega L_2(x)I_2(x) - j\omega M(x)(I_1(x) - I_2(x)) \] (A-7)

Applying the concepts of differential mode and common mode into the parameter distribution, we define:

\[ \begin{align*}
R(x) &= \frac{1}{2}(R_1(x) + R_2(x)) \\
\Delta R(x) &= \frac{1}{2}(R_1(x) - R_2(x)) \\
L(x) &= \frac{1}{2}(L_1(x) + L_2(x)) \\
\Delta L(x) &= \frac{1}{2}(L_1(x) - L_2(x))
\end{align*} \] (A-8)

Thus, we have

\[ \begin{align*}
R_1(x) &= R(x) + \Delta R(x) \\
R_2(x) &= R(x) - \Delta R(x) \\
L_1(x) &= L(x) + \Delta L(x) \\
L_2(x) &= L(x) - \Delta L(x)
\end{align*} \] (A-9)

Also define

\[ \begin{align*}
Z(x) &= 2[R(x) + j\omega(L(x) - M(x))] \\
\Delta Z(x) &= 2(\Delta R(x) + j\omega \Delta L(x))
\end{align*} \] (A-10)

Combining (A-7)(A-8)(A-9)(A-10) yields
\[ \frac{d}{dx} V(x) = Z(x)I_{1-2}(x) + \Delta Z(x)I_0 \] (A-11)
Equations (A-6) and (A-13) are the two basic differential equations describing the performance of an inhomogeneous structure. To obtain the solution, they are combined together as

\[
\begin{align*}
- \frac{d}{dx} V(x) &= Z(x)I_{1-2}(x) + \Delta Z(x)I_0 \\
- \frac{d}{dx} I_{1-2}(x) &= y(x)V(x)
\end{align*}
\]  

(A-14)

The general solution is dependent on the detailed functions of \(Z(x)\), \(\Delta Z(x)\) and \(y(x)\). If the functions \(Z(x)\) and \(y(x)\) have constant values respectively and \(\Delta Z(x)\) is equal to zero, the described structure is degraded to a homogeneous structure. Thus, the homogeneous structure is a special class of the inhomogeneous structures.

As a specific case, all parameter values change linearly along the \(x\) direction. Under this condition, equation (A-14) can be simplified.

The parameter distribution functions can be represented as

\[
\begin{align*}
R_1(x) &= R_{10} + r_1 \cdot x \\
R_2(x) &= R_{20} + r_2 \cdot x \\
L_1(x) &= L_{10} + l_1 \cdot x \\
L_2(x) &= L_{20} + l_2 \cdot x \\
C_1(x) &= C_{10} + c_1 \cdot x \\
C_2(x) &= C_{20} + c_2 \cdot x \\
G_1(x) &= G_{10} + g_1 \cdot x \\
G_2(x) &= G_{20} + g_2 \cdot x \\
M(x) &= M_0 \sqrt{L_1(x)L_2(x)}
\end{align*}
\]  

(A-15)

Equ. (A-14) is simplified as

\[
\begin{align*}
- \frac{d}{dx} V(x) &= (A_z + B_z \cdot x)I_{1-2}(x) + (A_{\Delta Z} + B_{\Delta Z} \cdot x)I_0 \\
- \frac{d}{dx} I_{1-2}(x) &= (A_y + B_y \cdot x)V(x)
\end{align*}
\]  

(A-16)

where \(A_z, B_z, A_{\Delta Z}, B_{\Delta Z}, A_y, B_y\), are all constants.

The differential equations describing the generalized inhomogeneous integrated structure are presented above, but the analytical solution has not been derived yet.
Appendix A: Study of Asymmetric Integrated Structures

Though the assumption of the linear parameter distribution simplifies the differential equations, the analytical solution has not been obtained by the author yet. For the research of a specific case with unknown parameter distributions, the numerical method can be used to obtain the solution.

![Figure A-12 Equivalent circuit for an infinitesimal length of an inhomogeneous structure](image)

**A.6. Summary**

The asymmetric performance of integrated power passives is reported in this paper. The asymmetry of the performance is due to the non-uniform distribution of parameters, which is determined by the dimensions, material properties and the structures. Experimental results show that the dominant factor influencing the extent of asymmetry is the structure, instead of the tolerance of dimension and material properties induced in the processing procedure. This implies that this class of structures can be deliberately designed to achieve specific functions.

Based on the applications in power electronics, the definition and criteria of the asymmetric integrated structure is proposed. Any two-conductor integrated structure can have symmetric terminal performance, if the parameter distribution is symmetric to either the vertical or the horizontal central line. This proposition is verified by experimental results.
The simplified lumped-parameter model and the corresponding circuit are proposed to investigate the low frequency performance of the integrated structures. A generalized model of the inhomogeneous integrated structure is proposed in form of differential equations.

With more and more applications of integrated power passives, the awareness and understanding of the asymmetric features of integrated structure is important for design and application. Though there is no direct application of the asymmetric feature at present, the fact that one unit can provide two different performances may be attractive in some applications in the future. Furthermore, the asymmetry in a structure allows use of the same component for multiple functions, simply by interchanging terminals. This can be an important trial in simplifying system assembly in integrated systems.
Appendix B: Study of Inductor Loss Measurement with Average Power Method

B.1. Introduction

The in-circuit loss measurement is important for the evaluation of the performance of power electronics components. The electrical and thermal measurement methods are discussed in the section. The analytical analysis of the measurement error of the electrical measurement is presented.

B.2. Theory of the Average Power Method

The average power loss in the inductor can be obtained by calculating the instantaneous power in one cycle, as expressed in the equation:

\[ P_{\text{loss}} = \frac{1}{T} \int_{0}^{T} v(t) \cdot i(t) \, dt \]  \hspace{1cm} (B-1)

This method requires the measurement of the voltage and current with high accuracy on both the magnitude and the phase shift. The measuring equipment and the experimental setup influence the measurement accuracy.

B.2.1. Instantaneous Power Measurement on a Resistor

When a sinusoidal excitation is applied to a resistor, the measured voltage and current waveforms can be expressed as the following:

\[ v(t) = V \sin(\omega t) \] \hspace{1cm} (B-2)

\[ i(t) = I \sin(\omega t - \sigma) \] \hspace{1cm} (B-3)

where \( \omega \) is the frequency of the exciting waveform, \( \sigma \) is the induced phase angle due to the error in the measurement.

Ideally, the obtained waveforms should exactly represent the real analog signals with the same curve shape and phase. However, in practical case, the measured waveforms are approximations of the real waveforms, due to the measurement error induced in sampling. Measurement errors are induced to both the magnitude and the phase of the waveforms. The comparison of the real signals and the measured waveforms are shown in Figure B-1.
B.2.2. Instantaneous Power Measurement on an Inductor

For an inductor, when a sinusoidal excitation is applied, the voltage and current waveforms can be expressed as:

\[ v(t) = V \sin(\omega t) \quad \text{(B-4)} \]

\[ i(t) = I \sin(\omega t - \frac{\pi}{2} + (\delta + \sigma)) = -I \cos(\omega t + \delta + \sigma) \quad \text{(B-5)} \]

where \( \omega \) is the frequency of the exciting waveform, \( \sigma \) is the induced phase angle due to the measurement error, \( \delta \) is the loss angle due to the power loss generated in the inductor.

The comparison of the real waveforms and the measured waveforms on an inductor are shown in Figure B-2.
Appendix B: Study of Inductor Loss Measurement with Average Power Method

The measurement error is mainly related to the sampling rate and the A/D conversion resolution of the oscilloscope. In present digital oscilloscopes, the sampling rate is much higher than the resolution of the A/D conversion. For a Tektronix TDS 7054, the A/D resolution is 8 bits, which represents 128 scales, while sampling rate can reach 5GS/s.

B.3. Measurement Error due to the Equipment

B.3.1. Measurement Error on the Waveform Magnitude:

When considering the measurement error on the curve shape, we assume the error due to sampling can be ignored. With the A/D resolution of \( n \) bits, the maximum induced percentage error in the measured magnitude can be calculated as:

\[
err(n) = \frac{1}{2^n} \times 100\%
\]  

(B-6)

It leads to the error in the power loss measurement in percentage, which can be calculated as

\[
error(n) = \max\left( \frac{1 - \left(1 - \frac{1}{2^n}\right)^2}{1} \times 100\%, \frac{1 - \left(1 - \frac{1}{2^n}\right)^2}{1} \times 100\% \right)
\]  

(B-7)

Table B-1 lists the percentage error versus the number of the bits of the A/D conversion.

<table>
<thead>
<tr>
<th>Number of the effective bits in A/D conversion, ( n )</th>
<th>Induced percentage error on the waveform magnitude</th>
<th>Induced percentage error in power loss measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td>7 bits</td>
<td>0.7813 %</td>
<td>1.5686 %</td>
</tr>
<tr>
<td>8 bits</td>
<td>0.3906 %</td>
<td>0.7828 %</td>
</tr>
<tr>
<td>9 bits</td>
<td>0.1953 %</td>
<td>0.3910 %</td>
</tr>
<tr>
<td>10 bits</td>
<td>0.0977 %</td>
<td>0.1954 %</td>
</tr>
<tr>
<td>11 bits</td>
<td>0.0488 %</td>
<td>0.0977 %</td>
</tr>
<tr>
<td>12 bits</td>
<td>0.0244 %</td>
<td>0.0488 %</td>
</tr>
<tr>
<td>13 bits</td>
<td>0.0122 %</td>
<td>0.0244 %</td>
</tr>
</tbody>
</table>
### Appendix B: Study of Inductor Loss Measurement with Average Power Method

<table>
<thead>
<tr>
<th>Resolution</th>
<th>Error 1</th>
<th>Error 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>14 bits</td>
<td>0.0061%</td>
<td>0.0122%</td>
</tr>
<tr>
<td>15 bits</td>
<td>0.0031%</td>
<td>0.0061%</td>
</tr>
</tbody>
</table>

From Table B-1, we observe, no matter in the application of a resistor or an inductor, the measurement error on the waveform magnitude doesn’t influence the measurement accuracy of the power loss much. The maximum percentage error is less than 1.6%. Thus, this error is not considered in our following analysis.

#### B.3.2. Measurement error on the phase shift:

To determine the phase of a waveform, it needs to determine the point that the waveform crosses the x axial. The accuracy of calculating the phase of the measured waveform is determined by two factors: the A/D resolution and the sampling rate. Figure B-3 and B-4 show how the waveforms are acquired.

![Figure B-3 Acquisition of waveforms](image)

Figure B-3 Acquisition of waveforms
B.3.3. Error in phase measurement induced by sampling rate only:

With the oscilloscope, TDS7054, the sampling rate can reach 5GS/s, which corresponds to 20,000 sampling points per cycle for a 250 kHz waveform. It implies that the minimum phase angle the oscilloscope can identify is \( \sigma = \frac{360^\circ}{20000} \approx 0.018^\circ \). Table B-2 shows how the sampling rate influences the measurement error of the phase.

Table B-2 Induced phase angle by the sample rate

<table>
<thead>
<tr>
<th>Signal period</th>
<th>Operating frequency</th>
<th>Number of sampled points per cycle</th>
<th>Induced measurement error due to the sampling rate (°)</th>
</tr>
</thead>
<tbody>
<tr>
<td>40 ns</td>
<td>25 MHz</td>
<td>200</td>
<td>1.8°</td>
</tr>
<tr>
<td>80 ns</td>
<td>12.5 MHz</td>
<td>400</td>
<td>0.9°</td>
</tr>
<tr>
<td>400 ns</td>
<td>2.5 MHz</td>
<td>2,000</td>
<td>0.18°</td>
</tr>
<tr>
<td>2 μs</td>
<td>500 kHz</td>
<td>10,000</td>
<td>0.036°</td>
</tr>
<tr>
<td>4 μs</td>
<td>250 kHz</td>
<td>20,000</td>
<td>0.018°</td>
</tr>
<tr>
<td>10 μs</td>
<td>100 kHz</td>
<td>50,000</td>
<td>0.072°</td>
</tr>
</tbody>
</table>
B.3.4. Error in phase measurement induced by A/D conversion only:

The resolution of A/D conversion (sample mode) is 8 bits (effective bits 6.8), which is corresponding to 128 scales. Increasing the sampling rate doesn’t improve the accuracy, as shown in Figure B-4 (c). To determine the crossing point, the measured data are filtered using average method.

The maximum error induced by A/D conversion is $\frac{1}{2}LSD$, which may move the waveform up or down in the range of $\frac{1}{2}LSD$ of the full measurement scale, and then induce extra phase shift, as shown in Figure B-5.

With n-bit A/D resolution, the induced error in the phase measurement can be calculated as $\sigma = \arcsin\left(\frac{1}{2^{n+1}}\right)$.
Table B-3  Induced error in phase measurement through the A/D conversion

<table>
<thead>
<tr>
<th>Number of the effective bits in A/D conversion, n</th>
<th>Induced extra phase shift due to the error of A/D conversion (°)</th>
</tr>
</thead>
<tbody>
<tr>
<td>7 bits</td>
<td>0.4442 °</td>
</tr>
<tr>
<td>8 bits</td>
<td>0.2229 °</td>
</tr>
<tr>
<td>9 bits</td>
<td>0.1117 °</td>
</tr>
<tr>
<td>10 bits</td>
<td>0.0559 °</td>
</tr>
<tr>
<td>11 bits</td>
<td>0.0280 °</td>
</tr>
<tr>
<td>12 bits</td>
<td>0.0140 °</td>
</tr>
<tr>
<td>13 bits</td>
<td>0.0070 °</td>
</tr>
<tr>
<td>14 bits</td>
<td>0.0035 °</td>
</tr>
<tr>
<td>15 bits</td>
<td>0.0017 °</td>
</tr>
</tbody>
</table>

**B.3.5. Measurement error of the phase**

To find out the maximum possible induced phase shift in the measurement, the influence of both the sample rate and the A/D conversion resolution are to be considered.

For oscilloscope TDS 7054, “Hi Res acquisition mode” provides additional resolution bits, as shown in Table B-4.

Table B-4  Additional resolution bits

<table>
<thead>
<tr>
<th>Time base speed</th>
<th>Bits of resolution</th>
<th>Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>40 ns and faster</td>
<td>8 bits</td>
<td>&gt;550 Mhz</td>
</tr>
<tr>
<td>80 ns to 200 ns</td>
<td>9 bits</td>
<td>&gt;110 Mhz</td>
</tr>
<tr>
<td>400 ns to 1μs</td>
<td>10 bits</td>
<td>&gt;22 MHz</td>
</tr>
<tr>
<td>2μs to 4μs</td>
<td>11 bits</td>
<td>&gt;5.5 MHz</td>
</tr>
<tr>
<td>10μs to 20μs</td>
<td>12 bits</td>
<td>&gt;1.1 MHz</td>
</tr>
<tr>
<td>40μs</td>
<td>13 bits</td>
<td>&gt;550 kHz</td>
</tr>
</tbody>
</table>
Appendix B: Study of Inductor Loss Measurement with Average Power Method

<table>
<thead>
<tr>
<th>Signal and input conditions</th>
<th>Effective bits</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 MHz, 9.2 div, 2GS/s sample rate, Sample acquisition mode</td>
<td>6.8 bits</td>
</tr>
<tr>
<td>1 MHz, 9.2 div, 10MS/s sample rate, Hi Res acquisition mode</td>
<td>8.7 bits</td>
</tr>
<tr>
<td>500 MHz, 6.2 div, 2GS/s sample rate, Sample acquisition mode</td>
<td>6.8 bits</td>
</tr>
</tbody>
</table>

Table B-5 Channel input and vertical specifications

Through the Hi Res acquisition mode can improve high A/D conversion resolution, in this mode the maximum sampling rate is 1.25GS/s.

The above data show that the A/D conversion resolution is dominant in determining the accuracy of the position of the crossing points. The maximum error induced from the acquisition is about 0.45°. Here we ignore errors induced from other resources, such as the asynchronism between different channels.

**B.4. Calculated power loss vs. of the phase shift**

**B.4.1. Measurement of Power Loss on a Resistor**

From equation (B-2) and (B-3), the power loss on a resistor without the induced phase shift can be represented with the instantaneous power as

\[
P_{loss} = \frac{VI}{2} \cos(\alpha)
\]

(B-8)

\(\alpha\) is the phase shift between the voltage and current waveform, which should be zero. Take the induced phase shift due to the measurement error as the disturbance of \(\alpha\),

\[
\sigma = \Delta \alpha
\]

(B-9)

Due to the error of phase measurement, the induced error on the calculated power loss is

\[
\Delta P_{loss} = -\frac{VI}{2} \sin(\alpha) \cdot \Delta \alpha = -\frac{VI}{2} \sin(\alpha) \cdot \sigma
\]

(B-10)

The percentage error of the power loss measurement is
\[
\frac{\Delta P_{\text{loss}}}{P_{\text{loss}}} = -\tan(\alpha) \cdot \sigma \rightarrow 0 \quad (\alpha \rightarrow 0, \sigma \rightarrow 0) \quad (B-11)
\]

It shows that the induced phase shift has little impact on the accuracy of the power loss measurement on a resistor.

**B.4.2. Measurement of Power Loss on an Inductor**

From equation (B-4) and (B-5), the power loss on an inductor without the induced phase shift can be calculated with the instantaneous power as

\[
P_{\text{loss}} = \frac{VI}{2} \sin(\delta) \quad (B-12)
\]

where \(\delta\) is the loss angle caused by the loss. At certain operating frequency, \(\omega\), it can be expressed as

\[
\tan(\delta) = \frac{R}{\omega L} \quad (B-13)
\]

where \(R\) is the equivalent resistance and \(L\) is the inductance.

The error in power loss measurement versus the varying of the loss angle is

\[
\Delta P_{\text{loss}} = \frac{VI}{2} \cos(\delta) \cdot \Delta \delta \quad (B-14)
\]

Take the induced phase shift \(\sigma\) as the disturbance of the loss angle:

\[
\sigma = \Delta \delta \quad (B-15)
\]

Thus

\[
\Delta P_{\text{loss}} = \frac{VI}{2} \cos(\delta) \cdot \sigma \quad (B-16)
\]

The percentage error of the power loss measurement is

\[
\frac{\Delta P_{\text{loss}}}{P_{\text{loss}}} = \frac{\sigma}{\tan(\delta)} \approx \frac{\sigma}{\delta} \quad (\sigma, \delta \rightarrow 0) \quad (B-17)
\]

It shows that when the loss angle is small, the induced phase shift has a great impact on the accuracy of the power loss measurement.
Appendix B: Study of Inductor Loss Measurement with Average Power Method

The comparison of power loss measurement on an inductor and a resistor is shown in Figure B-6. To a resistor, the phase shift between voltage and current waveforms is small (ideal zero), thus the influence of the induced phase shift on the power loss measurement accuracy is small. To an inductor, the phase shift between voltage and current waveforms is about one fourth of the period, (ideally 90°), thus the influence of the induced phase shift on the power loss measurement accuracy is large.

It shows that the phase shift between the voltage and current waveforms is the dominant factor influencing the accuracy of power loss measurement.

**B.4.3. Relative Error on Power Loss measurement vs. Magnitude and Phase Shift**

Consider a general case with sinusoidal excitation as shown in Figure B-7. The voltage and current waveforms can be expressed as:

\[
\begin{align*}
    v(t) &= V_o \sin(\omega t) \\
    i(t) &= I_o \sin(\omega t - \alpha)
\end{align*}
\]  

where the \( V_o, I_o \) are the magnitude of the voltage and current waveforms, respectively; \( \alpha \) is the phase shift between the voltage and current waveforms, representing the power loss in the DUT.

The power loss in the DUT can be calculated as
Appendix B: Study of Inductor Loss Measurement with Average Power Method

\[ P_{\text{loss}} = \frac{V_o I_o}{2} \cos(\alpha) \]  \hspace{1cm} (B-19)

\[ \frac{\Delta P_{\text{loss}}}{P_{\text{loss}}} = \frac{\Delta V_o}{V_o} + \frac{\Delta I_o}{I_o} + \tan(\alpha) \cdot \Delta \alpha \]  \hspace{1cm} (B-20)

where \( \Delta V_o, \Delta I_o, \Delta \alpha \) are the disturbance of the measured values \( V_o, I_o, \alpha \), respectively, which is due to the measurement error.

Setting the magnitude measurement error as \( \Delta m \), \( m = V_o, I_o \), the curve of the power loss measurement error versus the magnitude measurement error can be shown in Figure B-8.

---

Figure B-7 Waveforms in a general DUT

Figure B-8 The power loss measurement error vs. the magnitude measurement error
The curve of the power loss measurement error versus the phase shift is shown in Figure B-9.

![Relative error vs. Phase shift](image)

Figure B-9 The power loss measurement error vs. the phase shift measurement error

### B.5. Case study

#### B.5.1. The scale of the loss angle of an inductor

Take a real inductor a study case. The specifications of the inductor are listed below.

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductance, L</td>
<td>$170 \mu$H</td>
</tr>
<tr>
<td>Estimated equivalent resistance, R</td>
<td>4 Ω</td>
</tr>
<tr>
<td>Operating frequency, f</td>
<td>250kHz</td>
</tr>
<tr>
<td>Loss angle, $\delta$</td>
<td>0.66°</td>
</tr>
</tbody>
</table>
B.5.2. The scale of the induced phase shift in the measurement

From observation on the measured data, the induced phase shift is about 1°. The time interval between the crossing points of two waveforms is about 10ns, which is corresponding to the phase shift 0.9° at the operating frequency of 250kHz.

B.5.3. The influence of the lead inductance on the Power Loss Measurement

It’s reasonable to regard the lead as a source inducing extra inductance.

When the circuit operates at high frequencies, the lead inductance influences the value of $\sigma$ on a resistor. Assuming the lead inductance is 300nH and the resistance is 4 Ω. When the operating frequency is 250kHz, the phase shift due to the lead inductor is

$$\sigma = \arctg\left(\frac{\omega L}{R}\right) = 1.074^\circ$$ \hspace{1cm} (B-21)

The induced relative error on the calculated power loss is

$$\text{error} = \frac{P_{\text{loss}} - \Delta P_{\text{loss}}}{P_{\text{loss}}} \times 100\% = \frac{\frac{VI}{2} \cos(0^\circ) - \frac{VI}{2} \cos(1.074^\circ)}{\frac{VI}{2} \cos(0^\circ)} \times 100\% = 0.018\% \hspace{1cm} (B-22)$$

Figure B-10 shows how the induced phase shift influences the relative error of power loss measurement on a resistor.

It shows that in the case that the ideal phase shift is zero, the relative error of the measured power loss less than 100%. In many applications, the impedance of the lead inductance is much smaller than the resistance. It ensures the relative error on power loss measurement keep small. Thus, the lead inductance has a small impact on the accuracy of the power loss measurement and can be ignored.

If the DUT is an inductor, the lead inductance should be much less than the inductor inductance, thus the influence can also be ignored.

However, if the DUT is a capacitor, the lead inductance may impact the power loss measurement accuracy.
Appendix B: Study of Inductor Loss Measurement with Average Power Method

Summary:

1. It’s feasible to use the average power method to measure the power loss on a circuit with small phase shift.
2. Large error exists when using the average power method to measure the power loss on a circuit with large phase shift.

B.5.4. Discussion:

The error induced in the measurement, as discussed in the above, is due to the system problems. The systematic error is indeterminate but stable under certain condition. Thus, all the measurement results, obtained under certain condition, have the same systematic error. Since we can never completely remove the systematic error, we are not able to get the approximation of the true value by averaging the results of several measurements. Unfortunately, under some measurement conditions, the systematic error is too large for an acceptable measurement.

It’s not feasible to directly measure the power loss on an in-circuit inductor using average power method. However, if the inductor can be taken off from the circuit and measured with a sinusoidal excitation, we can add a capacitor to form a LCR resonator working close to the resonant point, then change the phase shift from 90° to a value close to 0° and have a highly accurate measurement. Ignoring the loss of the capacitor, the loss measured can be regarded as the loss of the inductor.

Figure B-10 Relative error of power loss measurement vs. the phase shift error
Appendix B: Study of Inductor Loss Measurement with Average Power Method

The average method can be applied to measure the power loss of an LCT module, which is connected to a resistor as the load. The phase shift between the voltage and current waveforms is small, due to the large load resistance. Thus both the input and output power of the LCT module can be measured accurately. By substituting the output power from the input power, we can obtain the loss the LCT. The rate of the loss on the LCT to the loss on the resistor influences the measurement accuracy. The larger the rate, the higher the accuracy is.

As an alternative, thermal method can be applied to measure the power loss in a component with the in-circuit excitation. Figure B-11 and B-12 show one of the thermal measurement setups. In this setup, a steady status temperature change of 1.5°C represents to a power loss increase of 0.1W. The setup can measure the component power loss within 2W. The measurement accuracy is mainly related to the sensitivity of the thermal couple. Keeping the external thermal environment same for both the calibration and measurement period is important for an accurate measurement.
Appendix C: Design of Experiment to Verify the Calculation of Cooling Structure

C.1. Introduction

Heat extractors have been proposed to improve the thermal-conduction ability of integrated power electronics modules (IPEMs). High thermal conductivity materials, such as aluminum nitride, are inserted into the cores of the IPEMs. Though the reduction of the ferrite core increases the loss density, the heat extractors provide good heat removal channels, which increase the thermal handling ability as well as the capability of loss density [78].

A two dimensional thermal model has been investigated with numerical methods to illustrate the loss density capability improvement of the IPEMs with embedded heat extractors [86]. According to the calculation results, the continuous flat plate is the optimal cross-sectional shape of the heat extractors, when the relative volume occupied by the heat extractors is less than 10%.

The numerical calculation is based on several ideal simplifications of the real case. To verify the calculation results, the experimental setup is expected to satisfy those assumptions for a two dimensional thermal model.

- Heat is generated uniformly inside the core.
- Heat dissipates through only the two surfaces perpendicular to the heat extraction direction.
- No heat flows along the width direction of the heat extractor.

Now, an experimental setup is needed to verify the calculation results.

Three questions are considered.

1. How to uniformly heat the tested sample. The structure of the sample and the application of magnetic field are to be designed.
2. How to measure the heat generated inside the ferrite core. Calculation and experimental measurement are needed.
3. How to measure the temperature distribution and the maximum temperature inside the ferrite.
C.2. Sample Structure

C.2.1. Heat Extractor

A heat extractor is a long strip made of high thermal conductivity material. The length usually is much larger than the width and height. It can be a cylinder or rectangle or any other shapes, depending on the detailed design requirement. Aluminum nitride with the thermal conductivity of 170 W/mK is a proper heat extractor material.

![Heat Extractor Diagram](image1)

Figure C-1 Heat extractor – long, narrow, thin, high thermal conductivity – aluminum nitride

C.2.2. Sample Structure

The tested sample is an object physically combining ferrite core and heat extractor. The heat extractor can be strips inserted inside the ferrite, as shown in Figure C-2, or the thin layers stacked with ferrite layers.

If the heat extractors are strips inserted inside ferrite core, there are two possible situations. The magnetic flux can be in parallel with the heat extractors, as shown in Figure C-3. The magnetic flux can also be perpendicular to heat extractors surface, as shown in Figure C-4. It’s clear that when the magnetic flux is in parallel with the heat
extractors, the magnetic field has little impact from the heat extractors. It’s a suitable structure.

If the heat extractors are thin layers, there is only one feasible situation, where the magnetic flux is in parallel with the heat extractors, as shown in Figure C-7.
C.3. Experimental Temperature measurement

Main temperatures of interest are the maximum temperature within the magnetic core and the temperatures on the exposed side surfaces of the inserted aluminum nitride strips. Locations of these are indicated in the below shown figures.
C.4. Construction

The magnetic core and cooling insert combination should be constructed in such a way as to inhibit magnetic flux as little as possible. For this reason, the cooling structures are placed in such a way as to run parallel to the flux lines. In order to insert the cooling structures and temperature measuring sensors, it would be necessary to construct the core in layers as shown in the figure below.
C.5. Heat Generation

To uniformly generate heat inside a ferrite core, a uniform magnetic field is needed. An inductor is the simplest example of heat generation inside a ferrite core.

C.5.1. Heat generation in a real inductor

If we take the whole inductor as the study object, as shown in Figure C-12, we may get a uniform heat generator by ignoring the no uniform magnetic field at the corners. However, if the winding covers the whole ferrite core, the heat extractors may lack contact area with the heat sink.

The heat extractors can be located perpendicular to the magnetic flux, as shown in Figure C-13, or parallel to the magnetic flux, as shown in Figure C-14. Obviously, the structure in Figure C-14 is more suitable for both electromagnetic performance and the assembly.
### C.6. A ferrite core with the round winding

![Diagram of round winding core](image)

It’s difficult to generate sufficient high frequency magnetic flux in a open magnetic loop.

#### C.6.1. A ferrite core with rectangular winding

![Diagram of rectangular winding core](image)

The magnetic field applied from external windings is uniform. However the open magnetic loop has large magnetic reluctance. It’s no way to create high magnetic flux density, as 0.1T with the alternating frequency 500kHz, which requires an 8A current in the windings.

#### C.6.2. A ferrite core with excitation support

A C type magnetic core is needed to form a closed loop with the tested sample and generate sufficient magnetic flux in the tested sample. Ignoring the magnetic flux concentration at corners, the magnetic filed in the tested sample is regarded uniform. Using thermal insulation material to separate the tested sample and the magnetic core and eliminate the heat leakage through the magnetic core. This structure is applied in the experimental setup.
The sample is designed long in order to eliminate the influence of the terminals where the flux density is not uniform.

High permeability material is applied to reduce the leakage flux.

Thermal insulation material eliminates heat transfer between the sample and the excitation loop.

Figure C-17 A ferrite core with excitation support

### C.7. Core loss calculation

With the input sinusoidal waveform, the Steinmetz equation can be used to calculate the core loss.

\[
\overline{P} = k f^\alpha \hat{B}^\beta V_{\text{core}}
\]  

(C-1)

\[f\] – frequency;

\[\hat{B}\] – peak magnetic flux amplitude;

\[V_{\text{core}}\] – core volume;

\[k, \alpha, \beta\] – material constants.

Considering the physical properties change with temperature, the equation can be modified with thermal coefficients as

\[
\overline{P} = k f^\alpha \hat{B}^\beta V_{\text{core}} \left( c_0 t_0 + c_1 T + c_2 T^2 \right)
\]

(C-2)

where \(c_0, c_1\) and \(c_2\) are constants from the manufactures data sheet.

By measuring the voltage across the inductor, the core loss can be obtained.

### C.8. Core loss measurement

Alternatively, by recording the input voltage, current and the phase, we can calculate the total power loss, including the core loss and the winding loss. Ignoring the capacitor loss, we obtain the inductor loss. Applying proper litz wire can eliminate the winding loss to a very loss value; thus, the measured power loss can be regarded as the core loss. The power loss in the tested sample can be calculated with volume ratio.

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C.8.1. Measurement of temperature

The temperature in the ferrite cores can be measured by thermo-couples. A hole is drilled on the ferrite core and a thermo-couple is placed. It measures the maximum temperature of the ferrite core, by locating the thermo-couple in the central of the core.

![Diameter of the hole is 1mm](image1.png)

Figure C-18 Diameter of the hole is 1mm

![5 ferrite core layers (5mm) and 6 aluminum nitride layers (0.5mm)](image2.png)

Figure C-19 5 ferrite core layers (5mm) and 6 aluminum nitride layers (0.5mm)

C.8.2. Experiment setup

The tested sample is composed of five ferrite core layers (4.5mm) and six aluminum nitride layers (0.5mm) stacked. The width of both the ferrite core and the aluminum nitride is 5mm. The sample is covered with thermal insulation material, which ensures heat transfers along one dimension. Thermal insulation material also covers the surface of the winding to prevent heat transfer between the winding and the sample. The input voltage, current and phase are measured. The maximum temperature and the temperature at terminals are measured.
Appendix C: Design of Experiment to Verify the Calculation of Cooling Structure

Figure C-20 Diagram of the experimental Setup

(a) Supporting frame    (b) with cover    (c) with heat sinks on sides

Figure C-21 Structure of the supporting frame

Figure C-22 Pictures of the practical supporting frame
C.8.3. Comparison experiment

To compare the performance of samples with and without heat extractors, the comparison experiment is designed. Two samples are built with or without heat extractors respectively, sharing the same core dimensions and external cooling conditions. By input different power, the thermal performance of the two samples can be illustrated with the maximum temperature difference between the inside of core and the heat sink surface. Thus, the thermal performance improvement by the heat extractors can be studied.
C.8.4. Measurement of contact thermal resistance

The contact thermal resistance between the ferrite core and the AlN substrate influences the cooling effect greatly. To make accurate calculation, the contact thermal resistance are to be measured for specific ferrite cores and AlN substrates. The contact thermal resistance is related to surface roughness and flatness, and the pressure exerted on the surface. With the normal commercial product of ferrite core and aluminum nitride, the measured interficial thermal resistance is shown in Figure C-24.

![Figure C-24 Measured interfacial thermal resistance of the commercial ferrite between aluminum nitride](image)

C.8.5. Design of electrical parameters

The core dimensions are listed below:

Cross-sectional area: \( A_{core} = 5 \times (4.5 \times 5) \times 10^{-6} \text{mm}^2 \)

Mean magnetic path length: \( MML = 2 \times (62 + 31) \times 10^{-3} \text{mm} \)

Core volume: \( V_{core} = MML \times A_{core} \text{mm}^3 \)

Designed power loss density: \( P_{LD} = 10^6 \text{W} \cdot \text{m}^{-3} \)

Designed power loss: \( P_{LD} = 20.9 \text{W} \)

According to datasheet, for 3F3 material at 1 MHz, the required magnetic flux density: \( B_D = 0.05 \text{T} \)

Set the turn’s number: \( n = 10 \)
Appendix C: Design of Experiment to Verify the Calculation of Cooling Structure

Required voltage across the inductor: \( V_o = nB_D A_{core} \cdot (2\pi f) = 353V \)

The measured inductance: \( L_r = 122 \times 10^{-6} \text{H} \)

The measured air gap in the magnetic loop: \( \text{gap} = 28 \times 10^{-6} \text{mm} \)

Required capacitance in series: \( C_r = 208 \times 10^{-12} \text{F} \)

The core loss functions as an equivalent resistor in parallel with inductor:

\[
R_{e} = \frac{V^2}{P_{D}} = 5.96k\Omega
\]

The value of the equivalent resistance is influenced by the inductance voltage:

\[
R_e = k \cdot V_L^{2-\alpha} \quad (k \text{ is a constant, } \alpha \text{ is the material constant with typicals value larger than 2}).
\]

Setting the voltage across the L-C series resonator is the input voltage and the inductor voltage is the output voltage, the transfer functions and gain can be obtained. The influence of the equivalent resistance on the overall impedance, the quality factor Q and the gain are shown in Figure C-25.

![Figure C-25 influence of the equivalent resistance to the resonator](image)

The sinusoidal excitation is provided by a power amplifier, which has the minimum output match impedance of 12.5Ω and the maximum output match impedance of 3000Ω. To obtain sufficient output power, the overall resistance of the L-C series resonator should be designed within the range.

As observed in Figure C-25 (c), the larger equivalent resistance, the smaller the overall resistance. When the amplifier starts to output, the voltage rises up from zero. At this
moment, the overall resistance of the resonator is much smaller than the required value. It leads to an un-stable output of the amplifier. To solve this problem, a 6Ω resistor is in series with the resonator.

To obtain enough voltage across the inductor, several factors need to be considered. First, the overall impedance should be large enough to maintain the stable output of the amplifier. Second, quality factor Q should be as large as possible to obtain sufficient inductor voltage with relatively low loss on other components. Third, the capacitance of the commercial capacitors is not accurate. Fourth, the air gap in the inductor is difficult to accurately control and the temperature rise also influence the inductance. However, the quality factor is sensitive to both the capacitance and inductance, when the excitation frequency is close to the resonant frequency. Calculation results are listed below to illustrate the influence of these factors to the inductor voltage.

![Figure C-26 Overall impedance vs. C](image1.png) ![Figure C-27 Quality factor vs. C](image2.png)
The influence of temperature rise on the inductance is not considered in the design. According to the datasheet of 3F3 material, when the temperature is 25°C, the initial permeability is 2000; when the temperature is 75°C, the initial permeability is 3000. In the practical measurement, since large magnetic field is applied, the amplitude permeability doesn’t change as much as the initial permeability. As observed, the temperature change doesn’t impact the circuit parameters much.

In the real circuit, the series resistance is 6 ohms; the output voltage of the amplifier is about 100V; the quality factor is about 3 to 4; the excitation frequency is a little smaller than the resonant frequency to obtain a stable excitation waveform. The ferrite core generated heat successfully.

However, due to the measurement error, the measured core loss using electrical measurement has big difference from the calculated heat using the thermal method. The amount of heat calculated with thermal method is regarded more accurate.
Appendix D: Construction Process

D.1. Introduction

The construction process for the integrated LLCT structure is based on a number of standard mass production processes that are presently used in power electronics industries. Although the construction process does not directly affect the electromagnetic modeling and performance of the integrated structures, the availability of processing technologies for the selected materials determines the feasibility of a specific design. The dimensional tolerance induced in the processing procedure restricts the range of designs that are realizable. The standardization of processing technology helps to reduce the cost of the mass production of the integrated power passive modules, which is important to the practical application of the integrated modules. The processing technologies applied for the implementation of planar integrated passive modules have been well addressed in [51]. In this section, the discussion focuses on the illustration of the step-by-step processing procedure to fabricate an integrated LLCT module with the stacked structure.

D.2. Construction of the LC cell

The LC cell is the fundamental element of the stacked structure. The implementation of the LC cell is to metalize the conductor, normally copper, on the desired dielectric substrate with the required dimensions. This is accomplished by the metallization of copper layer on the dielectric material and the removal of undesired copper. The dielectric material used in this design is the N1250 ceramic with the dimension of 120mm x 60 mm x 0.15 mm. The detailed processing procedure is illustrated in the following associated with the diagram.

D.2.1. Cleaning

The ceramic must be thoroughly cleaned to ensure the best adhesion between the metalized layers to the ceramic. So far, two methods are used to do the cleaning as listed in Table D-1.
Table D-1 Two cleaning methods of the ceramic substrate

<table>
<thead>
<tr>
<th>Method for cleaning</th>
<th>Material and equipment</th>
<th>Operation</th>
<th>Time</th>
</tr>
</thead>
<tbody>
<tr>
<td>Method 1</td>
<td>Acetone and propanol</td>
<td>soak</td>
<td>4 hours</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Forced air dry</td>
<td></td>
</tr>
<tr>
<td>Method 2</td>
<td>Acetone</td>
<td>soak</td>
<td>30 min</td>
</tr>
<tr>
<td></td>
<td>5% NaOH solution</td>
<td>soak</td>
<td>3 hours</td>
</tr>
<tr>
<td></td>
<td>5% H₂SO₄ solution</td>
<td>soak</td>
<td>1 hour</td>
</tr>
<tr>
<td></td>
<td>D-I water</td>
<td>rinse</td>
<td>5 min</td>
</tr>
<tr>
<td></td>
<td>Force air dry</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

D.2.2. Sputtering

RF sputtering is used to deposit both the titanium adhesion layer and copper seed layer in an argon atmosphere. The titanium layer is firstly sputtered on the substrate surface, and then the copper layer is sputtered on the titanium layer. After that, the substrate is turned over and the whole process is repeated on the other side. The procedure and parameters are summarized in Table D-2.

Table D-2 Sputtering process parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Time</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Titanium adhesion layer</td>
<td>15 min @ 100W</td>
<td>300 ± 50 nm</td>
</tr>
<tr>
<td>Copper seed layer</td>
<td>25 min @ 100W</td>
<td>500 ± 100 nm</td>
</tr>
</tbody>
</table>

D.2.3. Electroplating

The thickness of the seed copper layer is normally less than the required thickness. The electroplating is used to increase the copper layer thickness. According to the optimal design stated in Chapter 6, the copper thickness should be about 60% (within the range [40%, 100%]) of the skin depth. At 1 MHz, the skin depth of copper is 66μm. The
desired copper thickness is 40μm. So, if the thickness is within the range from 26.4 to 66μm, it is acceptable.

It is difficult to identify such a small thickness with equipment during electroplating procedure. The best way to control the thickness of the electroplated layer is to control the electroplating current and time. As illustrated in Figure D-1, In the electroplating procedure, copper ion with valence +2 captures 2 electrons, returns to copper, and deposit on the sample, which is connected to the negative electrode of the external voltage source. It can be expressed with the equation:

\[ \text{Cu}^{2+} + 2e^- = \text{Cu} \]  

(D-1)

Figure D-1 The mechanism of electroplating

Table D-3 lists the data related to copper electroplating. Assume that all electrons provided in the cathode are captured by copper ion. The maximum electroplating rate can be defined as the electroplated copper thickness created with unit current density in unit time. If the applied current density is 1 A/cm\(^2\) and the electroplating time is 1 minute, the electroplated copper thickness would be 22.2μm, as calculated in the following.

\[
\left[\frac{A}{cm^2}\right] \times \left[\frac{cm^2}{1cm^2}\right] \times \left[\frac{1min}{60s}\right] = \frac{63.57 g/mole}{1.6 \times 10^{-19} C/electron \times 2 \times \frac{electron}{atom} \times 6.02 \times 10^{23} \frac{atoms}{mole}} \times \frac{1g/cm^3}{8.92 g/cm^3} \times \frac{1\mu m}{10^4 \mu m} = 22.22 \mu m
\]

Thus, the electroplated copper thickness can be calculated by the equation

\[ h_{cu} = k \cdot J \cdot T \]  

(D-2)

where \(h_{cu}\) is the copper thickness with the dimension [μm], \(J\) is the current density with the dimension [A/cm\(^2\)], \(T\) is electroplating time with dimension [min], \(k\) is the maximum electroplating rate with the dimension [μm / A cm\(^2\) min]
Appendix D: Construction Process

Table D-3 Data related to electroplating

<table>
<thead>
<tr>
<th>Atomic Weight of copper</th>
<th>63.57</th>
</tr>
</thead>
<tbody>
<tr>
<td>Valence of copper ion</td>
<td>2</td>
</tr>
<tr>
<td>Density of copper</td>
<td>$8.92 \times 10^6 \text{ g/m}^3$</td>
</tr>
<tr>
<td>Avogadro’s number</td>
<td>$6.02 \times 10^{23} \text{ atoms/mole}$</td>
</tr>
<tr>
<td>Electron charge</td>
<td>$-1.6 \times 10^{-19} \text{ C}$</td>
</tr>
</tbody>
</table>

However, the maximum electroplating rate may be quite different from the real value. To accurately control the copper thickness of the LC cells, experiments have been setup to measure the real electroplating rate. The experimental details are listed in Table D-4. It needs to be noted that the electroplating rate may be influenced by many factors. The measured value is available only for the specific experimental environment, which is the same with that for electroplating of the substrates of LC cells.

Table D-4 Experimental data for the measurement of the real electroplating rate

<table>
<thead>
<tr>
<th>Temperature</th>
<th>25°C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electroplating solution</td>
<td>Copper sulfate (saturated)</td>
</tr>
<tr>
<td>Sample material</td>
<td>Single layer PCB</td>
</tr>
<tr>
<td>Copper dimension</td>
<td>58 mm x 30 mm</td>
</tr>
<tr>
<td>Current density</td>
<td>$0.02 \text{ A/cm}^2$</td>
</tr>
<tr>
<td>Average electroplating current</td>
<td>0.355 A</td>
</tr>
<tr>
<td>Electroplating time</td>
<td>50 min</td>
</tr>
<tr>
<td>Estimated thickness</td>
<td>22.7 μm</td>
</tr>
<tr>
<td>Initial thickness (center)</td>
<td>1.518 mm</td>
</tr>
<tr>
<td>Final thickness (center)</td>
<td>1.530 mm</td>
</tr>
<tr>
<td>Thickness of electroplated copper</td>
<td>12 μm</td>
</tr>
<tr>
<td>Real electroplating rate</td>
<td>$11.8 \text{ μm/A cm}^2 \cdot \text{min}$</td>
</tr>
</tbody>
</table>
The electroplated copper layer has non-uniform thickness. The edge is always thicker than the central part as illustrated in Figure D-2. For best results, the plating should be done at a low current density, for example 0.02 A/cm\(^2\). This is due to the increased voltage drop drop across the plating surface at high plating current. Providing uniform electrical field between the copper target and the sample would help to eliminate the unevenness of the thickness.

![Figure D-2 Cross-section of electroplated substrate](image)

When producing the LC cells, the substrate is electroplated on both sides at the same time. The experiment data are listed in Table D-5.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total copper area on one side (including leads)</td>
<td>28.2 cm(^2)</td>
</tr>
<tr>
<td>Total electroplating current (two sides) (1)</td>
<td>~ 1.2 A</td>
</tr>
<tr>
<td>Electroplating time (1)</td>
<td>90 min</td>
</tr>
<tr>
<td>Total electroplating current (two sides) (2)</td>
<td>~ 0.6 A</td>
</tr>
<tr>
<td>Electroplating time (2)</td>
<td>30 min</td>
</tr>
<tr>
<td>Initial thickness (center)</td>
<td>0.158 mm</td>
</tr>
<tr>
<td>Final thickness (center)</td>
<td>0.238 mm</td>
</tr>
<tr>
<td>Electroplated copper thickness (each side)</td>
<td>40 μm</td>
</tr>
</tbody>
</table>

**D.2.4. Photo masking**

Photo resistor paste is evenly daubed on the copper layer of the substrate. The whole sample is baked at 120°C for 40 minutes to dry the paste. After both sides of the substrate are covered by dry photo resistor layers, they are covered with two patterns. To accurate
control the capacitance, the strict requirement for conductor line-up of both sides is needed. Two patterns aligned by crop masks on both sides as shown in Figure D-3 are used during the sequential exposure. The exposed part of the negative photographic masks do not dissolve in the developer liquid, D4000 and remain as the masks to protect the desired copper in the chemical etching process.

![Top mask](image1.png) ![Bottom mask](image2.png)

Figure D-3 Negative photographic masks for two LC cells

**D.2.5. Chemical etching**

The schematic cross-section of the substrate with photo-mask is shown in Figure D-4. The chemical etching process parameters are shown in Table D-6.

![Schematic cross-section of the substrate](image3.png)

Figure D-4 Schematic cross-section of the substrate

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Copper etchant</td>
<td>Ferric Chloride @ 50 -60 °C</td>
</tr>
<tr>
<td>Titanium etchant</td>
<td>5% TFT solution @ room temperature</td>
</tr>
</tbody>
</table>

Table D-6 Chemical-etching parameters
Appendix D: Construction Process

For multi-winding structure, insufficient etching may leave shorts between turns. However, over etching will reduce the copper dimension by etching off the copper at the edge.

**D.2.6. Laser cutting**

To remove the redundant ceramic, the substrate is cut by laser along the edge of the copper layer with a tolerance of about 0.3 mm.

**D.2.7. LC cells**

A completed LC cell is shown in Figure D-5. The capacitance of LC cells is measured by multimeter and listed in Table D-7.

![Figure D-5 An LC cell](image)

**Table D-7 Measured capacitance of LC cells**

<table>
<thead>
<tr>
<th>Cell number</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Capacitance (nF)</td>
<td>6.32</td>
<td>5.73</td>
<td>6.51</td>
<td>6.04</td>
<td>5.22</td>
<td>6.43</td>
<td>6.42</td>
<td>5.20</td>
</tr>
<tr>
<td>Design capacitance for each cell (nF)</td>
<td>6.16</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Total capacitance of LC cells (nF)</td>
<td>47.87</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Total designed capacitance (nF)</td>
<td>49.28</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**D.3. Construction of the windings**

The stacked structure provides small space for interconnections at the two ends. The interconnections become the critical part of the integrated module. For mass production, the interconnections can be made with a series of standardized processes, which is suitable for automation and provides modulized products with small variation of parameter values. For experimental prototypes, of which the parameter values would be tuned frequently, the conventional technique, soldering is preferred for the interconnections.

The automation processes for the primary winding are illustrated in Figure D-6. The LC cells are fabricated with leads on the ceramic (Figure D-6 (a)), and stacked with the core,
and then encapsulated together with epoxy as a module. The two end surfaces of the module are polished to bare the leads. Interconnections are made on the two end surfaces using the metallization technologies addressed above.

The prototype for test as stated in Chapter 7 is made with the conventional interconnection technology. Two copper foils are soldered on each end of an LC cell as leads. Leads are soldered together to make electrical connections. The LC cell and the leads are shown in Figure D-7. The LC cell is covered with a solder mask layer for insulation. The leads are insulated by 0.2 mm Kapton tape.

The secondary wind is made of 2 mil copper foil with the width of 10 mm and covered by 0.2 mm Kapton tape for insulation.
Appendix D: Construction Process

D.4. Construction of the core with embedded heat extractors

The cores consists of I shape and E shape cores. To have heat extractors embedded into the cores, the cores are customized to shorter pieces having the dimensions shown in Figure D-8. A notch is specially made on the I shape cores for the placement of (K type) thermocouple. The heat extractors are made of 0.5mm aluminum nitride substrates and have the same cross-sectional dimensions with the core, but the thickness is 0.5 mm. The heat extractors are cut by laser and the edge is polished by diamond grinding wheel. The picture of cores and heat extractors are shown in Figure D-9. The cores and heat extractors are mechanically combined with thermal adhesive, DM4130SM, which is reworkable and provide good thermal conductivity in the interface. The DM4130SM is a two silver-loaded polymeric paste designed as electrically conductive organic replacements for high lead content solders used with surface mount devices [87]. Though this material is conductive, it does not influence the performance of the prototype, because it forms a very thin layer between the core and heat extractors. The typical properties of the material are listed in Table D-8.

![Figure D-8 core dimensions](image1)

![Figure D-9 Picture of cores and heat extractors](image2)
Table D-8 Typical properties of DM4130SM

<table>
<thead>
<tr>
<th>Properties</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistivity, $\mu\Omega$-cm</td>
<td>240</td>
</tr>
<tr>
<td>Adhesion, psi (2)</td>
<td>2500</td>
</tr>
<tr>
<td>Contact resistance, m$\Omega$/jxn</td>
<td>&lt;5</td>
</tr>
<tr>
<td>Rework Temperature, °C (4)</td>
<td>150</td>
</tr>
<tr>
<td>Thermal conductivity, W/m°C</td>
<td>5</td>
</tr>
<tr>
<td>Thermal expansion, ppm/°C</td>
<td>30</td>
</tr>
<tr>
<td>Curing profile</td>
<td>180 °C, 30 min</td>
</tr>
</tbody>
</table>

**D.5. Construction of the prototype**

The prototypes are constructed with the internal structure shown in Figure D-10. The leakage layers are applied in general designs but not needed in this design.
**D.6. Summary**

Every construction technology stated above induces dimensional tolerance. For realizable designs, the constructional limitations and dimensional tolerances must be considered in a practical design. A list of the technological limits and their present values with tolerances are given in Table D-9.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Present value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Min. dielectric thickness</td>
<td>150 μm</td>
</tr>
<tr>
<td>Min. secondary thickness</td>
<td>25.4 μm</td>
</tr>
<tr>
<td>Min. insulation thickness</td>
<td>200 μm</td>
</tr>
<tr>
<td>Min. core thickness</td>
<td>1.5 mm</td>
</tr>
<tr>
<td>Min. heat extractor thickness</td>
<td>0.5 mm</td>
</tr>
</tbody>
</table>
### Appendix D: Construction Process

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max. LC cell aspect ratio</td>
<td>0.5</td>
</tr>
<tr>
<td>Max. working temperature</td>
<td>100 °C</td>
</tr>
<tr>
<td>Fixed FPC permeability</td>
<td>8</td>
</tr>
<tr>
<td>Min. FPC layer height</td>
<td>0.2 mm</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Processing technology</th>
<th>Tolerance (experimental value)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electroplating</td>
<td>&lt; 20 μm</td>
</tr>
<tr>
<td>Photo-masking</td>
<td>&lt; 200 μm</td>
</tr>
<tr>
<td>Etching</td>
<td>&lt; copper thickness</td>
</tr>
<tr>
<td>Laser cutting</td>
<td>&lt; 300 μm</td>
</tr>
<tr>
<td>Polishing with grinding wheel</td>
<td>&lt; 500 μm</td>
</tr>
</tbody>
</table>
Appendix E: In-circuit Loss Measurement

E.1. Introduction

The accurate measurement of the in-circuit loss of the integrated LLCT prototype is importance for the evaluation of the stacked structure. Only the experimental values can be used to identify the efficiency of the practical prototype. The measured loss is compared to the calculated loss to verify the loss models used in design approaches. The loss models provide the estimation of temperature distribution and are the basic of thermal management for specific constructions.

The in-circuit loss measurement normally can be done with electrical and thermal methods. The electrical method is to measure the difference between the input and output power [88]. This accuracy of the indirect measurement method is influenced by the efficiency much. With high efficiency, the measurement accuracy is low. The general instrument to measure the in-circuit loss is the calorimeter, which records the dissipated heat in thermal steady state as the generated loss [89]. However, the high frequency operation of the LLC resonant converter limits the length of leads used to connect the LLCT prototype with the circuit. To put the prototype in the calorimeter, the leads would be long and the induced inductance will significantly change the overall resonant inductance, and thus change the circuit operation. To measure the in-circuit loss of the LLCT prototype, a specific but simple structure is developed to monitor the heat flux at the thermal steady state.

E.2. Method of the developed in-circuit power loss measurement

Considering the integrated prototype shown in Figure 7-7, if the two ends are covered with thermal insulation material, the heat generated inside the module mainly dissipate through the four side surfaces, which are flat and can be regarded as isothermal surfaces.

Now we think about a heat generating block with \( n \) flat and isothermal surfaces dissipating heat through heat sinks. If we insert a uniform thermal resistant layer with constant thermal properties between the block surfaces and the corresponding heat sinks, the layers can be regarded as thermal resistors with constant resistance. We assume the inserted layers have uniform thickness and thermal properties. Both surfaces of one layer
are isothermal. Thus, the heat flux has uniform distribution inside one thermal resistant layer. The heat dissipated through the layer at thermal steady state can be described with one-dimensional heat conduction model as

\[ q = \frac{\Delta T}{R} \]  

(E-1)

where \( \Delta T \) is the temperature drop across the two surfaces of the thermal resistant layer, \( R \) is the overall thermal resistance of the layer between the module surface and the heat sink, including the interfacial thermal resistance. Summing up the heat dissipated through all side surfaces, the total heat is obtained as

\[ Q = \sum_{i=1}^{n} q_i = \sum_{i=1}^{n} \left( \frac{k_i}{\Delta x_i} \cdot A_i \cdot \Delta T_i \right) \]  

(E-2)

where \( \Delta x_i, k_i, \) and \( A_i \) are the thickness, equivalent thermal conductivity and area of the \( i^{th} \) thermal resistant layer, respectively. The equivalent thermal conductivity \( k_i \) takes the impact of the interfacial thermal resistance into account. Its value is influenced by the pressure and the specific surface conditions. By using screws and frames, we can mechanically keep the pressure identified during the experiment. Assuming the material properties of the thermal resistant layer are constant, we regard the item \( \frac{k_i}{\Delta x_i} \) as the \( i^{th} \) characteristic parameter, a variable with constant value. The values of the characteristic parameters can be identified in the calibration experiment.

For a heat block with \( n \) heat dissipating surfaces, equation (E-2) has \( n \) unknown variables. We can measure the total heat and record the temperature drop at thermal steady state in \( n \) calibration experiments and solve the combined \( n \) equations to obtain the values of the variables. After calibration, we can apply some or all these thermal resistant layers to completely cover the under-test module before the heat sinks are attached. When the module is under test and reaches thermal steady state, the power loss in the module can be determined with the recorded temperature drop. Here we make two important assumptions related to the measurement accuracy.

- The heat is dissipated only through the monitored surfaces in both the calibration experiment and the measurement experiment.
The values of characteristic parameters remain constant in the calibration and measurement experiment.

In this power loss measurement method, the detailed heat distribution inside the module does not influence the measurement results. No matter how much heat dissipates through each surface, the heat flux penetrating each thermal resistant layer is directly related to the temperature drop across the thermal resistant layer and represents the dissipated heat after multiplying the characteristic parameter and the area of the layer. The sum of weighted temperature drops is equal to the total dissipated heat.

In the practical experiment, the thermal resistant layer attached on each surface can be made of the same material. With the same surface condition and external pressure, the values of \( \frac{k}{\Delta x} \) are the same for each layer. Then only one variable remains in equation (E-2). It can be simplified to

\[
Q = \frac{k}{\Delta x} \sum_{i=1}^{n} (A_i \cdot \Delta T_i)
\]  

(E-3)

In this case, only one calibration experiment is needed to obtain the value of the characteristic parameter \( \frac{k}{\Delta x} \). In equation (E-3), the area \( A_i \) can be regarded as the weight of the temperature drop. The equation shows that the sum of the weighted temperature drop at thermal steady state is proportional to the total heat. A set of calibration experiments can be made with various amount of heat. A line would be obtained from the calibration data by taking the total heat as the horizontal axis and the sum of the weighted temperature drop as the vertical axis. The slope is the reciprocal of the characteristic parameter. The linearity of the line shows how good the above assumptions are satisfied.

For the specific module shown in Figure 7-7, all heat dissipating surfaces have the same dimensions and area \( A \). The equation (7-3) can be further simplified to

\[
Q = C \cdot \sum_{i=1}^{n} \Delta T_i
\]  

(E-4)

where \( C \) is the characteristic parameter including the factor of the area \( A \). Based on the above assumptions, the value of \( C \) is determined by the material properties, dimensions,
interfacial conditions and pressure, which can be mechanically identified during the experiment.

**E.3. Calibration experiment**

To practically apply the measurement method in the power loss measurement of the integrated LLCT module, the proper material is needed to be selected for the thermal resistant layer. The overall equivalent thermal resistance is of the most concern. If the overall thermal resistance is large, the temperature drop across the thermal resistant layer would be large and cause high temperature inside the module. If the overall thermal resistance is small, the monitored temperature drop would be small, and the measurement accuracy will reduce due to the reduced signal-to-noise ratio.

In the original cooling structure of the integrated LLCT module, the thermal interface material, gap pad, is applied between the module surfaces and the heat sinks to improve the interfacial thermal conductivity. If the gap pad with appropriate material properties is found, we can simply use it as the thermal resistant layer. Figure E-1 shows the selection of the gap pad based on the trade-off between good heat dissipation ability and the measurement accuracy. The assembly issue is also a concern to determine the desired thickness of the gap pad. The gap pad 3000 ( \( k = 3 \text{ W/m-K}, d = 0.4 \text{ mm} \)) is selected as the thermal interface material, as well as the thermal resistant layer.

![Figure E-1 Concerned factors in the selection of thermal resistance layer material](image-url)
To keep external conditions identified and to eliminate the measurement error induced from different setups, a comparison module is made to have the same external dimensions as the LLCT module, so that the comparison module can share all the external setup with the integrated module. The comparison module is a rectangular tank made by 20 mil copper foil. After a resistor is located inside the tank, the two ends of the tank are sealed with thermal isolation covers. Two leads connect the resistor and the power supply through holes on the covers. To efficiently conduct heat between the resistor and the tank, the heat transfer fluid, paratherm NF, is injected into the tank through the two holes on the covers. To have sufficient space for the volume expansion when the fluid is heated, the two holds are connected to two short pipes. When the fluid is heated, the expanded fluid will be stored in the pipes. K type thermocouple and Agilent 34970A Data Acquisition/ Switch Unit are used to measure and record temperature.

The power loss in the comparison module varied from 2 watts to 23 watts and the temperature drops at the thermal steady state were recorded. The results are shown in Figure E-2. The temperature drops across gap pads on the four sides are different because neither the inside structure of the module nor the external cooling condition is practically symmetric. The sum of the total temperature drop shows good linear relationship with the power loss. However, the intercept of the fitted curve with the vertical axis has a large value. That means heat leakage is large. That is mainly due to two reasons. The heat transfer fluid vaporizes and takes heat away, especially when the temperature is high. The thermocouple placed on both sides of the gap pad makes the interface not flat and worsens the interfacial condition.

To fix the heat leakage problem and improve the interfacial condition, the heat transfer fluid is replaced by thermal joint compound Type 120 [90], which is a grease-like material without the evaporation problem, and the temperature drop measurement structure is also improved with the heat flux sensor structure.

As shown in Figure E-3, two copper plates are attached on both sides of the gap pad to satisfy the assumption that the surfaces of the gap pad are isothermal. With thermocouples placed on the outside surface of the copper plates, the structure functions as a heat flux sensor, which can monitor the temperature drop across the gap pad. Two thermocouples are placed on one surface. The average of the two measured values
represents the surface temperature. To improve the interfacial conditions, two additional gap pads are applied on both sides of the heat flux sensor. The structure is clamped by a frame with screws to accurately control the pressure at a constant value.

\[
y = 0.4637x + 0.326 \\
R^2 = 0.9992
\]

Figure E-2 Temperature drops on each surface vs. heat. Data from the calibration experiment with the comparison module having heat transfer fluid

Before the module is warmed up, the initial value read from every temperature measurement channel is recorded. The initial value is used to compare with the value
Appendix E: In-circuit Loss Measurement

obtained at each thermal steady state. Instead of the temperature value, the change of temperature value is used to calculate the temperature drop to avoid the steady state measurement error induced by the instrument and thermocouples.

The calibration experiment was repeated five times within 29 days to investigate the factors influencing the measurement results. As shown in Figure E-4, a linear function is applied to fit the curves. In all the five data lines, the intercepts with the vertical axis have a negligible value. That means the improved structure greatly reduces the heat leakage, which is negligible in the present experiment. The lines show very good linearity and agree with the linear relationship described in equation (E-4). Data 1 were obtained in the first calibration experiment on day 1. Data 2 and 3 were obtained on day 14 and 15 respectively with nothing changed. Data 4 were obtained on day 16 after the whole structure was un-installed and re-installed. Data 5 were obtained on day 29 with nothing changed from the previous experiment.

The corresponding values of the characteristic parameter $C$ are calculated and shown in Figure E-5. The value of the characteristic parameter $C$ increases while time passes, because the heat flux sensor is under pressure. The longer the pressure is exerted on the heat flux sensor, the better the interfacial thermal conditions, thus the larger the value of $C$. Nevertheless, the observation on the data shows that the value of $C$ does not change much within such a short time as one day. It is also observed that the un-installation and

<table>
<thead>
<tr>
<th>Data</th>
<th>Equation</th>
<th>$R^2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$y = 0.6478x - 0.0107$</td>
<td>0.9992</td>
</tr>
<tr>
<td>2</td>
<td>$y = 0.5841x + 0.0028$</td>
<td>0.9998</td>
</tr>
<tr>
<td>3</td>
<td>$y = 0.5764x - 0.0107$</td>
<td>0.9998</td>
</tr>
<tr>
<td>4</td>
<td>$y = 0.5744x + 0.0748$</td>
<td>0.9995</td>
</tr>
<tr>
<td>5</td>
<td>$y = 0.5157x + 0.1883$</td>
<td>0.9990</td>
</tr>
</tbody>
</table>

Figure E-4 Results of five calibration experiments within 29 days

The corresponding values of the characteristic parameter $C$ are calculated and shown in Figure E-5. The value of the characteristic parameter $C$ increases while time passes, because the heat flux sensor is under pressure. The longer the pressure is exerted on the heat flux sensor, the better the interfacial thermal conditions, thus the larger the value of $C$. Nevertheless, the observation on the data shows that the value of $C$ does not change much within such a short time as one day. It is also observed that the un-installation and
re-installation does not influence the $C$ value. Hence, if the calibration experiment and the loss measurement experiment are made within a short period with the same heat flux sensors, the calibrated results are valid in the calculation of the power loss.

![Measured $C$ value vs. time](image)

**Figure E-5 Calibration values of the characteristic parameter vs. time**

### E.4. Measurement experiment

By replacing the comparison module with the LLCT prototype, the whole setup can be placed in a 1 MHz 500W converter, excited with voltage and current waveforms, and the in-circuit power loss of the prototype can be measured. The experimental setup for the in-circuit loss measurement is shown in Figure E-6. The heat flux sensors occupy small volume and will not influence the original setup of the converter.

![Experimental setup](image)

(a) Cross-sectional view  
(b) In-circuit measurement setup

**Figure E-6 Experimental setup of the integrated module loss measurement**
Appendix E: In-circuit Loss Measurement

The prototype was tested at the operation frequency of 845 kHz, less than 1 MHz. The low frequency operation is due to the long leads applied in the secondary side for placing the current sensor for other experimental purposes. Due to the EM noise induced by the high electromagnetic field at 845 kHz operation, the temperature values acquired through the data acquisition unit and thermocouples are not accurate when the converter is functioning. Especially, when the power turns on or off, there is a sudden jump or drop of the read values. Figure E-7 shows the measured values from one of the thermocouples during the transient when the converter power is turned off. The converter power is turned off at second 6, when the read value of the temperature is 31.85°C. The read values in the next three seconds are 31.95°C, 31.25°C and 30.55°C, respectively. After second 9, the read values are consistent with a quadratic curve, which represents the cooling situation of the setup. As illustrated in Figure E-7, the temperature values recorded after second 9 are valid, while those recorded before second 9 are disturbed by EM noise and invalid in the data process. The transient period starts after the converter power is turned off, and ends when the disturbance from the EM noise stops. The experiment results show that the transient period is less than 3 seconds. Due to the existence of the transient period, the thermal steady state temperature is not able to be directly monitored. This induces measurement error which will be detailed in later.

Generally, high sampling rate is desired for accurate identification of the moment that the converter power is turned off. Recording signals from 18 channels at one sampling cycle, the data acquisition unit limits the shortest sampling period no less than 0.77 second. To get accurate time reference, the sampling period is set one second.

To find out the thermal steady state temperature, the data recorded from second 9 to second 13 are applied for curve fitting. The data processing result shows that the quadratic function can best represent the cooling situation with the maximum R-squared value. With the curve fitting equation, the total temperature drop at arbitrary moment can be calculated by applying the equation with the value of the time variable. As indicated in Figure E-7, the setup is at thermal steady state at second 6, but no longer at thermal steady state at second 7. That means the moment that converter power is turned off should be within second 6 and second 7. Second 6.5 is regarded the effective moment.
Appendix E: In-circuit Loss Measurement

with the time error of 0.5 second. The estimation of the total weighted temperature drops at the thermal steady state is illustrated in Figure E-8.

Figure E-7 EM noise interference on the acquired data

\[ y = 0.0149x^2 - 0.5148x + 7.4851 \]

\[ R^2 = 0.9997 \]

Figure E-8 Total temperature difference estimated by curve fitting
Appendix E: In-circuit Loss Measurement

In the power loss measurement experiment, the LLC resonant converter has an output voltage of 50V and output current of 10.6A. The measured LLCT module loss is 13.6 W. The efficiency of the integrated module is 97.4%.

E.5. Error analysis

The measurement error of this structure mainly lies in two aspects: the determination of characteristic parameter value $C$ and the temperature measurement. The value of $C$ used to calculate the power loss may not be accurate. The calibration and calculation equation (E-4) is obtained based on the assumptions: isothermal surface, uniform distribution of heat flux, unchanged physical properties of the heat flux sensor, and no heat leakage. These assumptions are satisfied through the structural design of the heat flux sensor and the mechanical design and the whole experimental setup. However, the practical setup can never reach the ideal conditions. The discrepancy between the practical situation and the ideal assumptions leads to error in determining the characteristic parameter value in calibration experiments. Since the characteristic parameter value is measured in the calibration experiment but used in the measurement experiment, the un-installation and re-installation between the calibration and measurement experiments may change the practical value of the characteristic parameter and induce error.

The measurement error induced by the above factors can be estimated by evaluating the variance of the characteristic parameter values measured in two calibration experiments made within a short time. According to the calibration experiments demonstrated above, the $C$ value from Data 2 is 1.712 [W/K], the $C$ value from Data 3 is 1.735 [W/K], and that from Data 4 is 1.741 [W/K]. The relative measurement error is represented by the variance of the $C$ values as

$$\text{error} = \frac{\sum|\Delta C|}{C/3} = \frac{(1.735 - 1.712)}{(1.741 + 1.735 + 1.712)/3} \times 100\% = 1.32\% \quad \text{(E-5)}$$

Temperature measurement error occurs with the usage of K type thermocouples and the data acquisition unit. Due to the specific situation illustrated in this file, extra measurement error is induced from the temperature sampling. Since the signals measured by thermocouples are disturbed by the EM noise from the high frequency operation of the converter, the thermal steady state temperature can not be directly and accurately
measured during the converter power is on. This problem can be fixed by replacing thermocouples with fiber optic temperature sensors which work stably with high frequency EM noise. However it will increase the cost. By using thermocouples to monitor the temperature, only the values recorded after transient period are valid. The proposed solution is to use those valid data to extrapolate the temperature at thermal steady state through curve fitting. Though curve fitting can be made with good agreement with the valid data, the error is induced when determining the exact moment that the converter power is turned off. Before the moment, the module is at thermal steady state and the data extrapolated from the fitted curve is not valid; after the moment, the module is not at thermal steady state and the dissipated heat is not equal to the power loss.

The existence of the transient period leads to uncertainty in determining the time of the latest thermal steady state moment. According to the estimation of the effective moment illustrated in the above, the induced error of time is half second. According to the data shown in Figure E-8, the induced error in this specific experiment is 3.36%. Using $f(t)$ to represent the function obtained from the curve fitting and $t$ represent to time variable, the error is calculated by the equation:

$$Error = \left(\frac{f(7) - f(6)}{2} / f(6.5)\right) \times 100\%$$  \hspace{1cm} (E-6)

Since the temperature change in cooling condition agrees with the quadratic function, higher thermal steady state temperature and higher sampling rate will lead to lower measurement error. Taking the above two errors into account, the overall measurement error is about 4.74%.

Besides the measurement of the power loss in the resistor in the calibration experiment is also a source of measurement error. When calculating the power loss in the resistor, loss dissipates on the lead connecting the dc resistance with the power supply should be considered. As estimated with the experimental data, the loss in the leads contributes about 4% of the total dc loss. Fortunately, the amount can be compensated in the calculation of resistor loss. The accuracy of the dc power measurement will affect the measured resistor loss as well. Since the induced error is small and difficult to identify, its influence is not taken into account in the error analysis.
Reference


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Vita

The author, Wenduo Liu, was born in October 1975 in Fuzhou, China. He received his B.S. and M. S. degree from Tsinghua University, Beijing China, in 1998 and 2001, respectively, both in mechanical engineering. He started Ph.D study at Center for Power Electronics Systems (CPES), Virginia Polytechnic Institute and State University and major in electrical engineering in August, 2001. He completed the research work for Ph.D degree in April 2006.